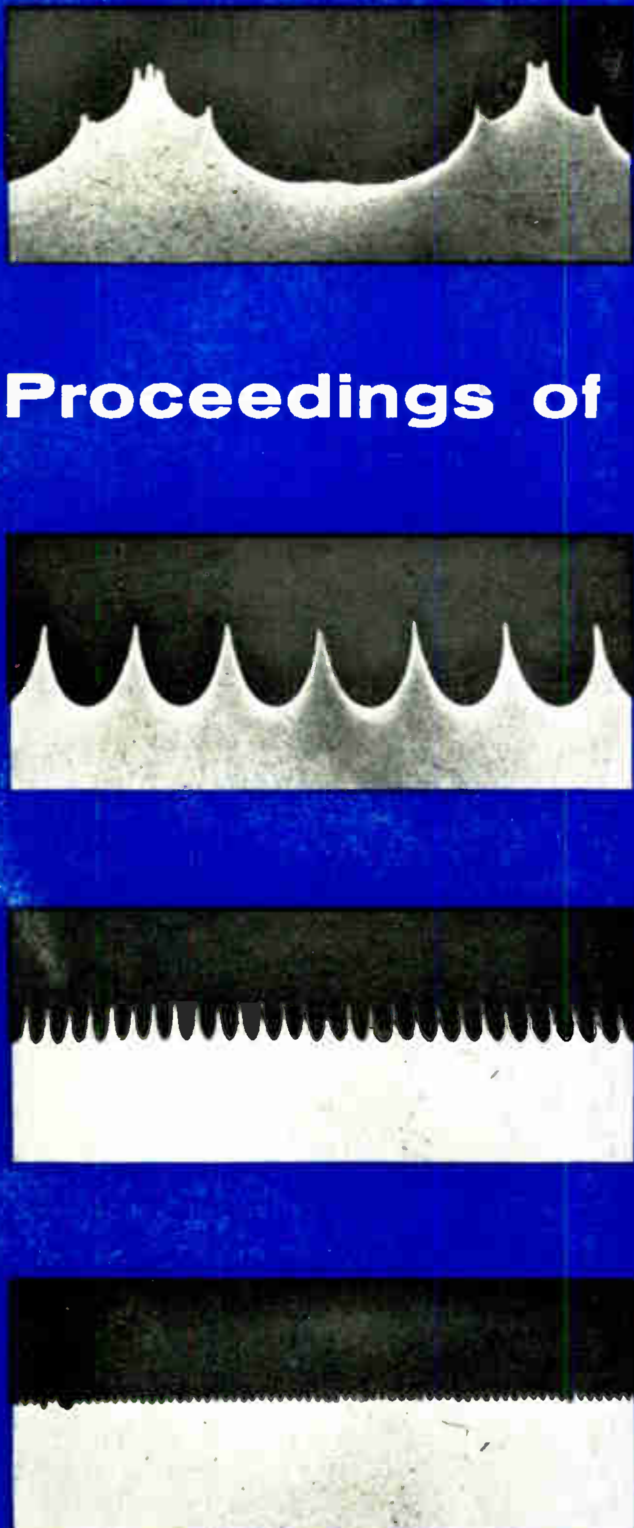

february 1961
the
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Proceedings of the IRE

in this issue

MAGNETIC TAPE READOUT WITH AN ELECTRON BEAM: page 498


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The bulk of UTC production is on special units designed to specific customers' needs. Illustrated below are some typical units and some unusual units as manufactured for special applications. We would be pleased to advise and quote to your special requirements.
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All types for frequencies from .1 cycle to 400 MC.
400 ~ telemetering, 3 db at ± 7.5%, 40 db at 230 and 700 ~, 5/8 x 1¼ x 2"
15 — BP filter, 20 db at 30 ~, 45 db at 100 ~, phase angle at CF less than 3° from —40 to +100°C.
LP filter within 1 db to 49 KC, stable to .1 db from 0 to 85°C., 45 db at 55 KC.
LP filter less than .1 db 0 to 2.5 KC, 50 db beyond 3 KC.
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Wound core unit .01 micro-second rise time.
Pulse current transformer 100 Amp.
Pulse output to magnatron, bifilar filament.
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VOLUME 49, NUMBER 2

February, 1961

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

contents

COVER

The interesting patterns on the cover were produced on a fluorescent screen by an electron beam which is reading magnetic tape recordings of sinusoidal signals. The novel readout system, developed jointly by the AIL Division of Cutler-Hammer, Clevite Corp. and the Department of Defense, is described on page 498.

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

continued

PROCEEDINGS OF THE IRE, published monthly by The Institute of Radio Engineers, Inc., at 1 East 79th Street, New York 21, N. Y. Manuscripts should be submitted in triplicate to the Editorial Department. Correspondence column items should not exceed four double-spaced pages (illustrations count as one-half page each). Responsibility for contents of papers published rests upon the authors, and not the IRE or its members. All republication rights, including translations, are reserved by the IRE and granted only on request. Abstracting is permitted with mention of source.

Thirty days advance notice is required for change of address. Price per copy: members of the Institute of Radio Engineers, one additional copy $1.25; non-members $2.25. Yearly subscription price: to members $9.00, one additional subscription $13.50; to non-members in United States, Canada, and U. S. Possessions $18.00; to non-members in foreign countries $19.00. Second-class postage paid at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of Febuary 28, 1925, embodied in Paragraph 4, Section 412. P. L. and R., authorized October 26, 1927. Printed in U.S.A.

Jans Kliphuis, Ken Siegel, and Jim Whelehan of the Department of Applied Electronics have been developing several different kinds of parametric amplifiers for use at C-band. They have obtained some unusually good experimental results with a degenerate amplifier, which is described below.

C-BAND DEGENERATE PARAMETRIC AMPLIFIER*

The parametric amplifier is a device in which amplification is obtained through the mixing action in a nonlinear reactance of an applied signal voltage and a larger amplitude, higher frequency pump voltage.[1] The mixing action results in the production of harmonics and sum and difference tones of the signal and pump frequencies, which are also produced in a conventional nonlinear resistance mixer. However, whereas the nonlinear resistance absorbs power and therefore introduces noise without providing amplification, a nonlinear reactance acts as an energy storage device that provides amplification while introducing very little noise.

Energy considerations for the nonlinear reactance show a basic difference in amplifier behavior depending on the terminating impedances provided for the various frequencies produced in the mixing process.[2,3] If for example, proper terminations are provided only at the signal and sum frequencies and the output signal is taken at the sum frequency, the amplifier has a maximum gain equal to the ratio of the sum and signal frequencies, and a positive resistance characteristic. In this mode of operation, the amplifier is absolutely stable and can have a fairly large amplification bandwidth. Alternatively, if proper terminations are provided only at the signal and difference frequencies, the amplifier has a negative resistance characteristic, and therefore can provide a large gain but at the expense of reduced bandwidth. In this case, the amplified output signal can be taken at either the signal frequency or the difference frequency. Basically, all these modes of operation exhibit low-noise characteristics.

Since the difference, or idler, frequency must be present to obtain amplification at the signal frequency, a parametric amplifier providing signal frequency amplification will have a spurious response because signals at the idler frequency that are applied to the amplifier input terminal will be converted to an output at the signal frequency. This characteristic is similar to that of a conventional mixer having an image frequency response.

For the special case where the pump frequency is exactly twice the signal frequency, the signal and idler frequencies are the same. In this degenerate mode of operation, input signals appearing below half pump frequency will generate corresponding signals above half pump frequency and vice versa. For signals at exactly half pump frequency, the amplifier gain is a function of the relative phase difference between the signal and half pump frequencies. Again, these operational characteristics are similar to those of a conventional mixer where the local oscillator frequency equals the signal frequency.

* The work reported here was performed under Contract AF30(602)-1854 with the Rome Air Development Center, Griffiss Air Force Base, New York.

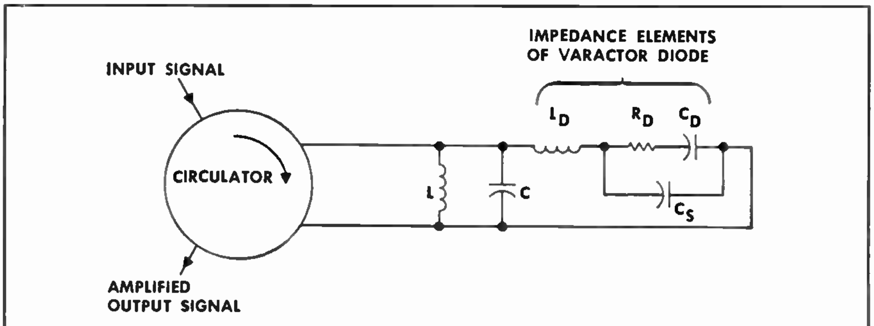

Figure 1. Schematic Diagram of Double-Tuned C-Band Degenerate Amplifier

Recently we have been working on a C-band degenerate amplifier for which the design goals are (1) a total amplification bandwidth (centered around half pump frequency) of 500 Mc, (2) a gain of at least 15 db, and (3) a broadband noise figure below 2 db. To obtain a low noise figure simultaneously with a large gain-bandwidth product (and operational stability), it was decided to use a circulator to interconnect the antenna, amplifier, and post receiver. Although the required gain-bandwidth product seemed beyond the state-of-the art, computations showed that it could be achieved by using a double-tuned input circuit in which the varactor diode and its parasitic reactances formed one resonant circuit. Although even larger gain-bandwidth products can be obtained by using more than two resonant circuits, the double-tuned circuit was chosen since (1) such a circuit is easier to design and more readily adjusted than more complex circuits, and (2) it provides the largest improvement in bandwidth with the smallest attendant increase in amplifier noise figure (because of resonant circuit losses). A schematic diagram of the amplifier is shown in Figure 1.

An MA 4254-X pill diode was used in the experimental amplifier. The measured signal frequency gain characteristic is shown in Figure 2. This curve was obtained by varying the signal frequency below and above half pump frequency (5300 Mc) and measuring the amplified output component only at the signal frequency (the idler component was rejected with a filter). The total amplifier bandwidth is 550 Mc and the peak amplifier gain is 18 db. The measured broadband noise figure of the amplifier alone is 1.6 db

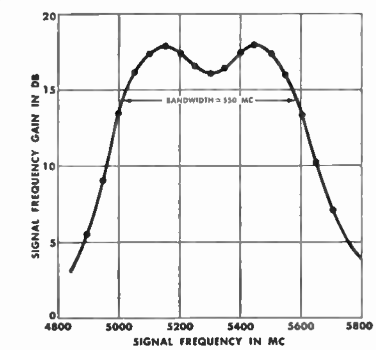

Figure 2. Measured Gain Characteristic of C-Band Degenerate Amplifier

and that for the amplifier-circulator combination is 1.9 db. The pump power required is about 50 milliwatts. Thus, this amplifier meets all the design goals.

Because of the exceptionally large amplification bandwidth and low noise figure obtained with this amplifier, it is now being evaluated for use in tunable radar systems and radio astronomy applications.

REFERENCES

[1] A van der Ziel, "On the Mixing Properties of Non-linear Condensers," Journal of Applied Physics, Vol. 19, p. 999-1006, November 1948.

[2] J. M. Manley and H. E. Rowe, "Some General Properties of Nonlinear Elements—Part I. General Energy Relations," Proc. IRE, Vol. 44, p. 904-913, July 1956.

[3] B. Salzberg, "Masers and Reactance Amplifiers—Basic Power Relations," Proc. IRE, Vol. 45, p. 1544-1545, November 1957.

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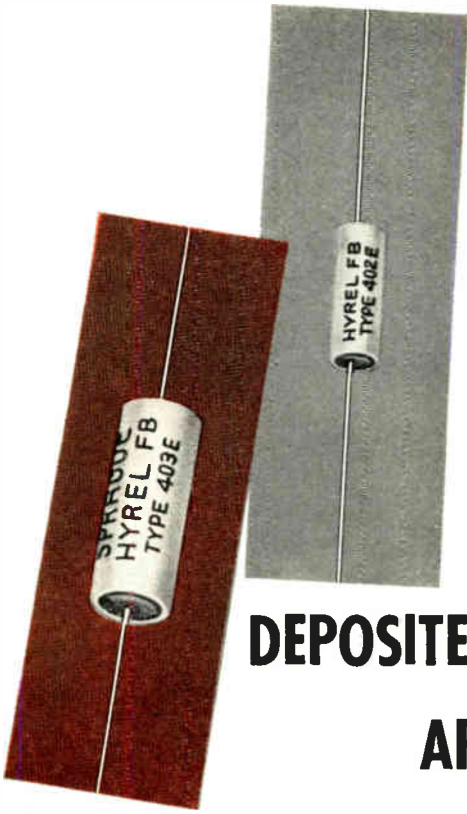

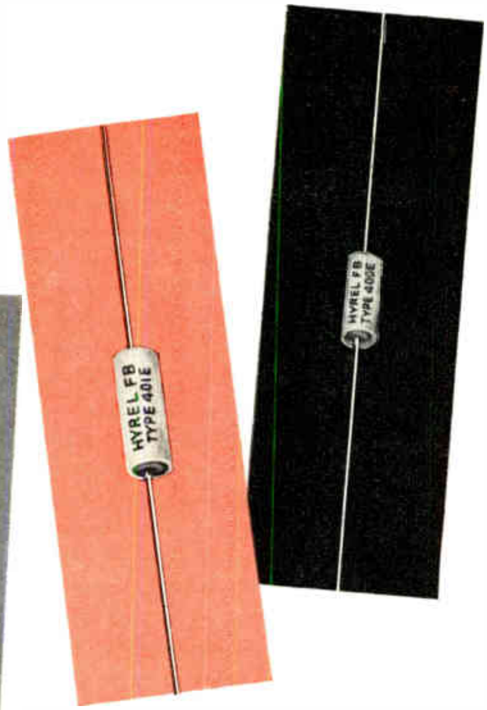

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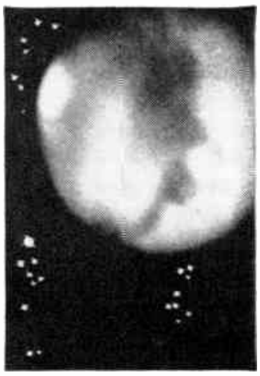

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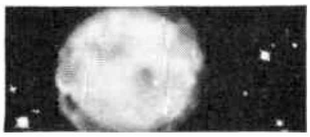

Bell Laboratories research with chilled ruby amplifiers speeds the day we may telephone via satellites

A strange combination of Nature's forces at Bell Laboratories foreshadows the day when world-wide phone calls may be relayed via man-made satellites orbiting the earth. It is a union of synthetic rubies and extreme cold, making it possible to amplify microwave signals from these satellites clearly.

Synthetic rubies possess an extraordinary property when deeply chilled and subjected to a magnetic field. They can be excited to store energy at the frequencies of microwave signals. As a signal passes through an excited ruby, it releases this energy and is thus amplified a thousandfold.

Bell Laboratories scientists chose a ruby amplifier because it's uniquely free of "noises" that interfere with radio signals. For example, it doesn't have the hot cathodes or hurtling electrons that generate noise in conventional amplifiers. It is so quiet that only the noise made by matter itself in heat vibrations remains. But at a temperature close to absolute zero, this also is silenced. Even very faint signals from satellites can be clearly amplified and studied for their possibilities.

Bell Laboratories scientists were first to discover that matter itself generates electrical noise. They also discovered that stars send radio waves, and thus helped found radio astronomy. It is particularly fitting that the same scientists, in their endless research on noise, should now battle this number-one enemy of telephony in the dramatic new field of communication via satellites. The ultimate goal, as always, is the improvement of your Bell System communications services.

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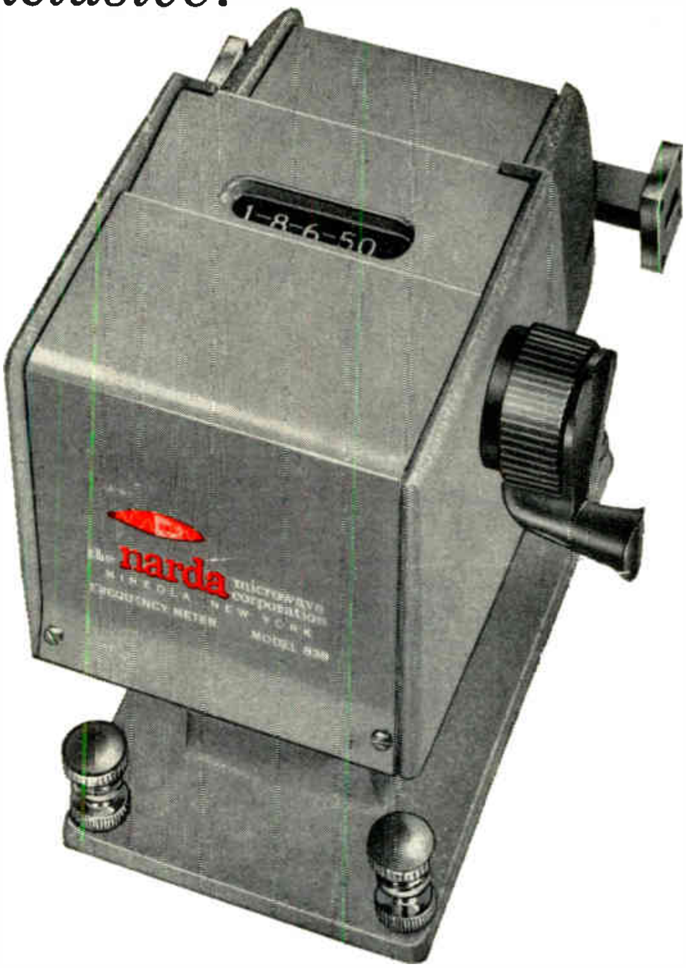

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SPECIFICATIONS

Band	X	KU	K
Frequency (KMC)	8.20-12.4	12.4-18.0	18.0-26.5
Waveguide (in.)	1 x ½	.702 x .391	½ x ¼
Accuracy	0.08%	0.1%	0.1%
Loaded Q	7000	5000	4000
Length (in.)	4	4	4
NARDA Model Number	840	839	838
Price	$195.	$250.	$275.

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Meetings with Exhibits

As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its sections and professional groups which include exhibits.

March 20-23, 1961

International Radio and Electronics Show and IRE International Convention, Waldorf-Astoria Hotel and New York Coliseum, New York, N.Y.

Exhibits: Mr. William C. Copp, Institute of Radio Engineers, 72 West 45th Street, New York 36, N.Y.

April 19-21, 1961

SWIRECO, South West IRE Regional Conference & Electronics Show, Dallas, Tex.

Exhibits: Mr. R. W. Olson, Texas Instruments, Inc., 6000 Lemmon Ave., Dallas 9, Tex.

April 26-28, 1961

Seventh Region Technical Conference and Trade Show, Westward Ho Hotel, Phoenix, Ariz.

Exhibits: Mr. G. J. Royden, 912 W. Linger Lane, Phoenix, Ariz.

May 8-10, 1961

National Aerospace Electronics Conference (NAECON), Miami & Dayton-Biltmore Hotels, Dayton, Ohio

Exhibits: Mr. R. J. Stein, 136 W. Second St., Rm. 202, Dayton 2, Ohio

May 9-11, 1961

Western Joint Computer Conference, Ambassador Hotel, Los Angeles, Calif.

Exhibits: John H. Whitlock Associates, 253 Waples Mills Road, Oakton, Va.

May 22-24, 1961

Fifth National Symposium on Global Communications (GLOBECOM V), Sherman Hotel, Chicago, Ill.

Exhibits: Mr. Robert Hajek, Commtronix Engineering, Box 62, Riverside, Ill.

May 22-24, 1961

National Telemetering Conference, Sheraton Towers Hotel, Chicago, Ill.

Exhibits: Mr. Frank Finch, 795 Gladys Ave., Long Beach 4, Calif.

June 6-8, 1961

Armed Forces Communications & Electronics Show, Sheraton Park and Shoreham Hotels, Washington, D.C.

Exhibits: Mr. William C. Copp, 72 W. 45th St., New York 36, N.Y.

June 13-14, 1961

Fifth National Conference on Product Engineering & Production, Philadelphia, Pa.

Exhibits: Mr. Paul J. Riley, Radio Corp. of America, Building 10-6, Camden 2, N.J.

(Continued on page 10A)

Meetings with Exhibits

(Continued from page 8A)

June 19-20, 1961

Chicago Spring Conference on Broadcast and Television Receivers, O'Hare Inn, DesPlaines, Ill.

Exhibits: Mr. Roy Lee, Philco Corp., 6957 West North Ave., Oak Park, Ill.

June 26-28, 1961

Fifth National Convention on Military Electronics, Shoreham Hotel, Washington, D.C.

Exhibits: Mr. L. David Whitelock, 5614 Greentree Road, Bethesda 14, Md.

July 16-22, 1961

Fourth International Conference on Medical Electronics & Fourteenth Conference on Electrical Techniques in Medicine & Biology, Waldorf-Astoria Hotel, New York, N.Y.

Exhibits: Mr. Lewis Winner, 152 W. 42nd St., New York 36, N.Y.

August 22-25, 1961

Western Electronic Show and Convention (WESCON), Cow Palace and Fairmont Hotel, San Francisco, Calif.

Exhibits: Mr. Don Larson, WESCON, 1435 LaCienega Blvd., Los Angeles, Calif.

October 4-6, 1961

IRE Canadian Convention, Exhibition Park, Toronto, Canada

Exhibits: Business Manager, IRE Canadian Convention, 1819 Yonge St., Toronto 7, Ontario, Canada

October 9-11, 1961

National Electronics Conference, Hotel Sherman, Chicago, Ill.

Exhibits: Mr. Arthur H. Streich, National Electronics Conference, 228 N. LaSalle St., Chicago 1, Ill.

October 23-25, 1961

East Coast Conference on Aeronautical & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md.

Exhibits: Dr. Harold Schutz, Westinghouse Electric Corp., Air Arm Div., P.O. Box 746, Baltimore, Md.

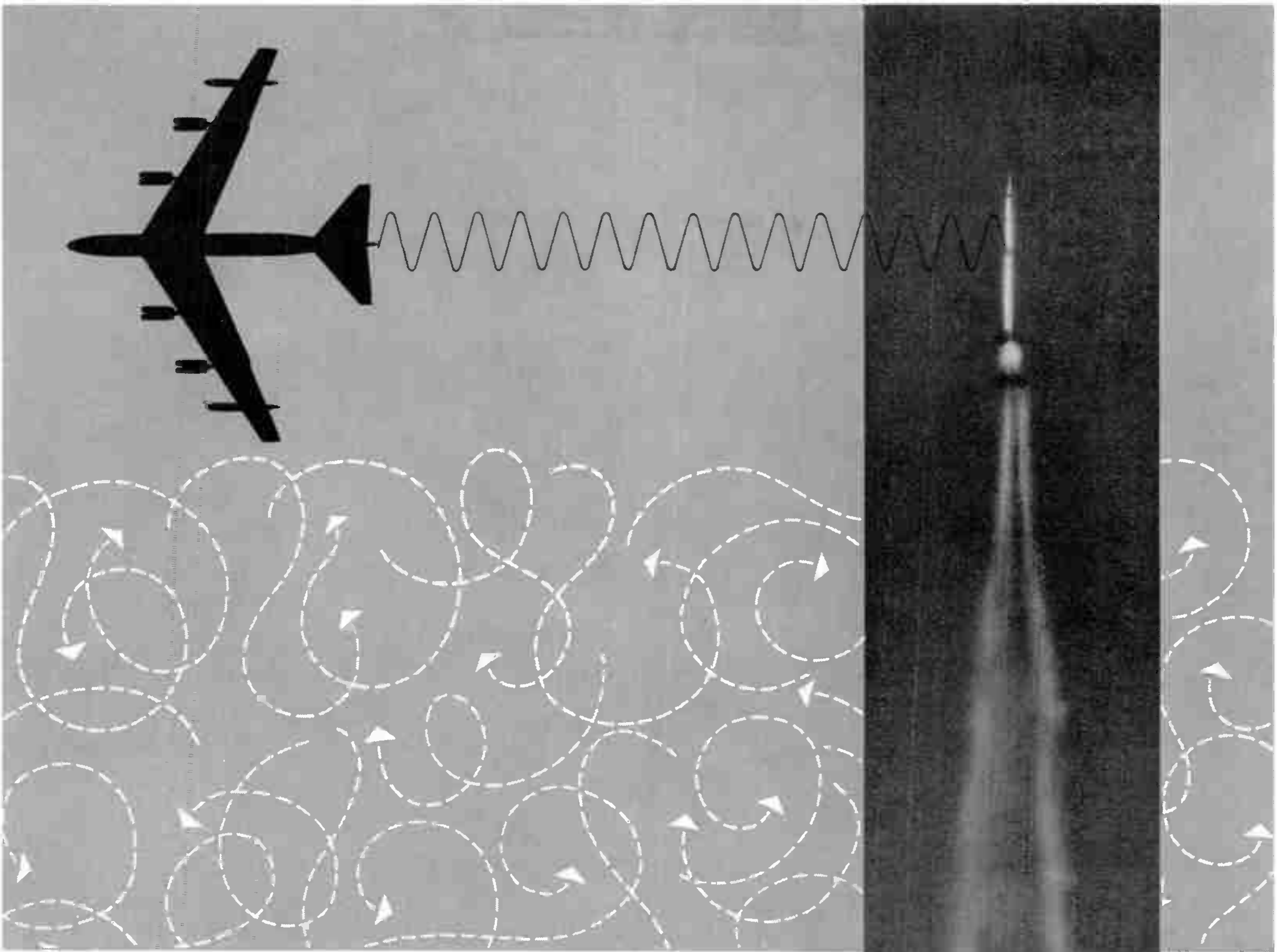
November 14-16, 1961

Northeast Electronics Research and Engineering Meeting (NEREM), Boston Commonwealth Armory, Boston, Mass.

Exhibits: NEREM, 313 Washington St., Newton, Mass.

Δ

Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.



Raytheon Subminiature Tubes Help Deliver The Message for Hughes Project Tattletale

Enemy atomic attack can scramble the ionosphere disrupting vital communications. The Air Force provides a solution in the form of Project Tattletale. A high altitude rocket containing a taped message and transmitting equipment is shot 300 miles up to provide a straight-line transmission requiring no ionospheric bounce.

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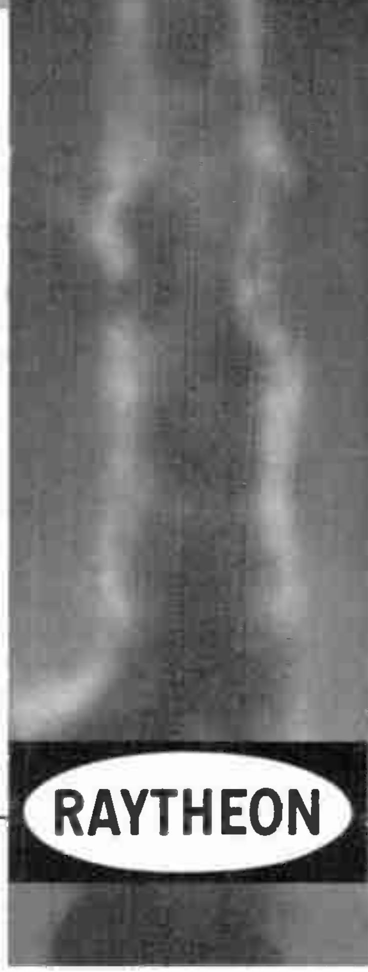

IRE News and Radio Notes

Current IRE Statistics

(As of December 31, 1960)
Membership—88,479
Sections*—109
Subsections*—28
Professional Groups*—28
Professional Group Chapters—278
Student Branches†—196

* See January, 1961 issue for a list.
† See October, 1960 issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1961

Int'l. Solid State Circuits Conf., University of Pennsylvania and Sheraton Hotel, Philadelphia, Pa., Feb. 15–17.

Annual Biophysical Society Mtg., Chase-Plaza Hotel, St. Louis, Mo., Feb. 16–18.

Int'l. Symp. on Semiconductor Devices, UNESCO, 2 Place Fontenoy, Paris, February 20–25.

Symp. on Engrg. Aspects of Magnetohydrodynamics, Univ. of Pa., University Park, Mar. 9–10.

IRE Int'l. Conv., N. Y. Coliseum and Waldorf-Astoria Hotel, New York, N. Y., Mar. 20–23.

2nd Int'l. Conf. on Quantum Electronics, Berkeley, Calif., Mar. 23–25.

PIB Int'l. Symp. on Electromagnetics and Fluid Dynamics of Gaseous Plasma, N. Y., N. Y., April 4–6.

Symp. on Information and Decision Processes, Purdue Univ., Lafayette, Ind., Apr. 12–13.

15th Ann. Spring Tech. Conf., Hotel Alms, Cincinnati, Ohio, Apr. 12–13.

SWIRECO, Dallas, Tex., April 19–21.

7th Region Tech. Conf. & Trade Show, Westward Ho Hotel, Phoenix, Ariz., April 26–28.

Electronic Comp. Conf., Jack Tar Hotel, San Francisco, Calif., May 2–4.

2nd Nat'l. Symp. on Human Factors in Electronics, Marriott Twin-Bridges Motor Hotel, Arlington, Va., May 4–5.

Workshop in Graph Theory, University of Illinois, Urbana, May 6.

5th Midwest Symp. on Circuit Theory, Allerton Park & Urbana Campus, Univ. of Ill., Urbana, May 8–9.

NAECON, Miami & Biltmore Hotels, Dayton O., May 8–10. (DL*: Abstracts, Jan. 1; Papers, March 7, A. J. Wilde, 4136 Lotz Rd., Dayton 29, Ohio.)

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Admiral Burke Heads MIL-E-CON Advisors

Admiral Arleigh Burke, USN, Chief of Naval Operations, heads a list of distinguished military officers and civilians who will serve as advisors for the Fifth National Convention on Military Electronics (MIL-E-CON 1961), to be held at the Shoreham Hotel, Washington, D. C., on June 26–28, 1961. This annual meeting is sponsored by the Professional Group on Military Electronics of the IRE.

Other advisors for MIL-E-CON 1961, as announced by Major General F. L. Ankenbrandt, USAF (Ret.), MIL-E-CON President, and member of the technical staff of Defense Electronic Products, RCA, Camden, N. J., are: J. H. Rubel, Deputy Director of Defense Research and Engineering, Department of Defense; Dr. H. L. Dryden, Deputy Administrator, National Aeronautics and Space Administration; Lieut. Gen. A. G. Trudeau, USA, Chief of Research and Development, Department of the Army; Lieut. Gen. R. C. Wilson, USAF, Deputy Chief of Staff, Development, U. S. Air Force; Vice Admiral J. T. Hayward, USN, Deputy Chief of Naval Operations (Development); Lieut. Gen. B. A. Schriever, USAF, Commander, Headquarters, Air Research and Development Command; R. H. Cranshaw, Manager, Advanced Space Projects, Light Military Electronics Department, General Electric Company, Utica, N. Y (President, 1960 MIL-E-CON); Dr. E. G. Witting, Chairman, Professional Group on Military Electronics, and Deputy Director of Research and Development, Department of the Army; D. C. Ports, Chairman, Washington, D. C. section of the IRE and Vice President, Jansky and Bailey, Washington, D. C.

Harry Davis, Deputy for Research, Assistant Secretary of the Air Force (Research and Development), Department of the Air Force, is Chairman of the Technical Program Committee.

L. D. Whitelock, Bethesda, Md., is Exhibits Chairman for the Convention.

The Professional Group on Military Electronics of the IRE, sponsor of MIL-E-CON 1961, has as its field of special interest the electronic systems, activities and services applicable to the requirements of the Armed Forces.

Attendance has increased at each of the four MIL-E-CONs held to date, and an increased attendance is expected for MIL-E-CON 1961.

JTAC Forms Committee on Space Communications

The Joint Technical Advisory Committee, recognizing the technical and policy problems concerning the United States, both nationally and internationally, in the utilization of space, has appointed Lieut. Gen. J. D. O'Connell, USA (Ret.), as chairman of a committee to study problems posed by the FCA's inquiry into the allocation of frequency bands for space communication problems of an industry-wide nature. (Docket No. 13522)

The activities of the committee will be devoted to the technical problems involved. To this end, recommendations have been made to initiate technical studies in the major problem areas. Such studies can be effective only if they are supported by members of industry who are willing to contribute actively their ideas, concepts and recommendations, and are immediately responsive to requests for information from such study groups as may be formed.

JTAC and its task groups desire to establish close communication with all agencies, activities and corporate bodies which have responsibilities in this field and to obtain an adequate and thorough consensus of the concepts and views of U. S. industry.

Working with General O'Connell on the committee are: R. Emberson, Associated Universities; R. P. Gifford, General Electric Co.; J. P. Hagen, NASA; J. W. Herbstreit, NBS Propagation Labs.; D. MacQuivey, Telecommunications Div., Dept. of State; R. Peavey, Space Science Board, NAS; A. M. Peterson, SRI; C. A. Petry, ARINC; T. F. Rogers, MIT Lincoln Lab.; and L. C. Tillotson, BTL.

The JTAC's two sponsors, the EIA and the IRE, also recognizing the need to go beyond the JTAC membership for collecting these data, are furnishing preliminary financial support. However, additional financial aid will be needed for these studies. Because of the unquestioned value of this study to industry, individual corporations will want to support this work in order to establish the U. S. position for the next international frequency allocation program.

Further information may be obtained from: L. G. Cumming, Secretary, JTAC, IRE Headquarters, 1 East 79 St., N. Y. 21, N. Y.

Biophysical Society Plans February Meeting

The Biophysical Society will have its annual meeting February 16–18, 1961, at the Chase-Plaza Hotel in St. Louis, Mo. The program includes contributed papers on many fields of biophysics, and symposia on Mutations, Membrane Transport and Computational Techniques.

For information, contact: Dr. W. Sleator, Dept. of Physiology, Washington University, St. Louis 10, Mo.

Geneva Section Established by IRE

On December 13, 1960, the IRE Executive Committee approved the establishment of a new IRE Section, to be known as the Geneva Section, and to encompass the entire country of Switzerland.

Call for Papers

1961 Western Electronic Show and Convention (WESCON)

August 22–25, 1961

Cow Palace, San Francisco, Calif.

The 1961 Western Electronic Show and Convention now issues a call for papers for its 1961 meeting, which is to be held August 22–25, at the Cow Palace in San Francisco, Calif.

Prospective authors are required to submit 100- to 200-word abstracts and 500- to 1000-word detailed summaries of their papers by May 1, 1961, in order to be considered for inclusion in the program. They will be notified by June 1, 1961, of acceptance or rejection of their papers.

Submissions should be sent to: E. W. Herold, c/o WESCON's Northern California Office, 701 Welch Road, Palo Alto, Calif.

International Conference On Medical Electronics Issues Call for Papers

The Fourth International Conference on Medical Electronics will be held July 16–21, 1961 at the Waldorf-Astoria Hotel, New York, N. Y. The conference follows three successful international meetings held in London and Paris. This year, it will be combined with the 14th Annual Conference on Electrical Techniques in Medicine and Biology. It is sponsored by the Joint Executive Committee on Medicine and Biology (IRE, AIEE, ISA) under the auspices of the International Federation for Medical Electronics and organized by the IRE through the Professional Group on Bio-Medical Electronics.

A technical program is now in preparation by an international sub-committee under the chairmanship of Dr. H. P. Schwan, University of Pennsylvania, Philadelphia. The theme of the conference covers a broad scientific area common to the engineering, medical, and biological fields. Acceptable topics for discussion include: models of biological systems, physiological monitoring, system analysis, electrical and radiation stimulation study, automation, instrumental diagnostic methods, data analysis techniques, physical-chemical procedures, etc. These subjects encompass medical specialties such as myo- and neurophysiology, cardiology, circulatory physiology, obstetrics, respiration, experimental psychology, clinical pathology, as well as engineering techniques such as microwaves, servo-systems analysis, transducers, computers, ultrasonics, telemetry, etc.

The program is now open to prospective contributors. An abstract of 300 words for preliminary review and a 50-word summary for inclusion in the advance program should be submitted before April 1, 1961.

It is planned to print a Conference Digest containing 600- to 1000-word digests of the presented communications, to be distributed to participants at the conference. Digests must be received by the Program Committee by May 15, 1961. Alternately, authors may elect to submit their digests for review in lieu of the 300-word abstract. In such cases this should be mailed, together with the 50-word summary, before April 1, 1961.

Abstracts and digests may be in English, French, German, or Russian. Prospective authors are requested to address all correspondence to: Dr. H. P. Schwan, Program Chairman, University of Pennsylvania, Philadelphia.

Another feature of the conference will be a large technical and scientific exhibit of the latest techniques and equipment in the medical electronic and electrical fields. Examples of scientific instrumentation, diagnostic and therapeutic equipment, techniques and systems will be displayed, and methods will be demonstrated.

Other highlights will include a banquet, a ladies' program, guided tours to scientific, educational and industrial laboratories and institutions and other unique points of interest in the New York metropolitan area.

The conference committee includes: L. E. Flory, chairman, David Sarnoff Research Center, Princeton, N. J.; O. H. Schmitt, vice chairman, University of Minnesota, Minneapolis; J. T. Farrar, general secretary, Veterans Administration Hospital, New York, N. Y.; C. Berkley, arrangements, Rockefeller Institute, New York, N. Y.; W. E. Tolles, publicity and liaison, Airborne Instruments Laboratory, Deer Park, L. I., N. Y.; C. G. Mayer, finance, RCA International, Westfield, N. J.; and L. Winner, conference manager, Consultant, New York, N. Y., as well as a number of representatives abroad who are acting as coordinators of the conference interests in their representative countries.

Professional Group News

At its meeting on December 13, 1960, the IRE Executive Committee approved the following new chapters: PG on **Engineering Management**—Baltimore Chapter; PG on **Human Factors in Electronics**—Los Angeles Chapter; the existing PG on **Microwave Theory and Techniques Chapter** combined with the PG on **Antennas and Propagation**—Buffalo-Niagara Chapter; PG on **Microwave Theory and Techniques**—Orlando Chapter.

Calendar of Coming Events and Author's Deadlines*

(*Continued from page 14A*)

Western Joint Computer Conf., Ambassador Hotel, Los Angeles, Calif., May 9–11.

Microwave Theory and Tech. Nat'l. Symp., Sheraton Park Hotel, Washington, D. C., May 15–17.

GLOBECOM V, Sherman Hotel, Chicago, Ill., May 22–24.

Nat'l. Telemetering Conf., Chicago, Ill., May 22–24.

Electro-Optical Devices Symp., Los Angeles, Calif., May.

3rd Nat'l. Symp. on Radio Frequency Interference, Washington, D. C., June 12–13.

5th Nat'l. Symp. on Product Engrg. and Production, Philadelphia, Pa., June 13–14.

2nd Nat'l. Conf. on Broadcast and Television Receivers, O'Hare Inn, Des-Plaines, Ill., June 19–20. (DL*: Feb. 15, 1961, N. Frihardt, Motorola Inc., 4545 W. Augusta Blvd., Chicago, Ill.)

MIL-E-CON 1961, Shoreham Hotel, Washington, D. C., June 26–28. (DL*: Feb. 1, 1961, H. Davis, SAFRD, The Pentagon, Washington 25, D. C.)

JACC, Univ. of Colorado, Boulder, June 28–30.

4th Int'l. Conf. On Medical Electronics & 14th Conf. on Elec. Techniques in Medicine & Biology, Waldorf-Astoria, Hotel, N. Y., N. Y., July 16–22. (DL*: April 1, 1961, H. P. Schwan, Moore School of E.E., Philadelphia 4, Pa.)

WESCON, San Francisco, Calif., Aug. 22–25. (DL*: May 1, E. W. Herold, WESCON North Calif. Office, 701 Welch Rd., Palo Alto, Calif.)

1961 Symp. on Transmission & Processing of Information, M.I.T., Cambridge, Mass., Sept. 6–8. (DL*: Abstracts, Jan. 1, 1961; Papers, April 1, 1961, P. Elias, M.I.T., Cambridge, Mass.)

1961 Nat'l. Symp. on Space Electronics and Telemetry, Albuquerque, N. M., Sept. 6–8.

Joint Nuclear Instrumentation Symp., North Carolina State College, Raleigh, N. C., Sept. 6–8.

9th Ann. Engrg. Management Conf., New York, N. Y., Sept. 14–16.

10th Ann. Industrial Electronics Symp., Boston, Mass., Sept. 20–21.

IRE Canadian Conv., Exhibition Park, Toronto, Can., Oct. 4–6.

Nat'l. Electronics Conf., Chicago, Ill., Oct. 9–11.

5th Nat'l. Symp. on Engrg. Writing and Speech, Kellogg Ctr. for Continuing Education, Michigan State Univ., East Lansing, Oct. 16–17.

East Coast Conf. on Aeronautical & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md., Oct. 23–25.

Elec. Tech. in Medicine & Biology Conf., Univ. of Nebraska, Lincoln, Oct. 26–27.

Radio Fall Mtg., Hotel Syracuse, Syracuse, N. Y., Oct. 30–31.

NEREM, Boston, Mass., Nov. 14–16.

MAECON, Kansas City, Mo., Nov. 14–16.

* DL = Deadline for submitting abstracts.

SWIRECO Plans Nearing Completion

A unique pre-registration campaign for the 13th annual SWIRECO, April 19–21, 1961, has already resulted in more than 2,000 paid registrations of engineers and other scientific personnel.

The precedent-setting pre-registration drive consists of personal calls on company executives, according to SWIRECO general chairman, R. W. Olson, research and engineering vice president of Texas Instruments, Inc. He explained that in several cases, management is pre-registering its entire force of electronics engineers, chemists, physicists, and others working in the forefront of advanced electronics and applied sciences. He stated further that SWIRECO's goal is more than 3,000 pre-registrations.

Mr. Olson also announced that L. V. Berkner, president of the IRE, will be a special guest speaker at the Conference.

Five moderators of key technical sessions have also been announced by O. Becklund of Texas Instruments, Inc., technical program chairman. They are: Dr. J. F. Reagan, Chance Vought, Inc., Systems, Sub-Systems and Equipment Design session moderator; Dean W. H. Hagerty, Dean of Engineering, University of Texas, Engineering Education session moderator; R. D. Alberts, chief of the Molecular Electronics Branch, Wright-Patterson A. F. B., Dayton, Ohio, moderator of the panel on Microminiature Devices and Circuits; and Major J. E. Steel, M. D., Aerospace Medical Div., Wright-Patterson A. F. B., moderator, Bionics session. Other principal sessions will include Computer and Data Systems and Communications.

The Conference and Show, which is also the meeting of Region 6 of the IRE, will be held in the Dallas Memorial Auditorium and the Baker Hotel, Dallas, Tex.

NEC Announces 1961 Executive Officers

J. J. Gershon, Director of the Resident School, De Vry Technical Institute, Chicago, Ill., has been elected president of the National Electronics Conference for 1961. Other officers named for the 1961 NEC, which will be held in Chicago at the International Amphitheatre on October 9–11, 1961, are: vice president, J. H. Kogen, GPE Controls, Inc.; secretary, Dr. T. F. Jones, Jr., Purdue University; treasurer, R. J. Parent, University of Wisconsin; and assistant treasurer, Dr. J. S. Aagaard, Northwestern University.

Mr. Gershon, past vice president and long active in NEC activities, succeeds Dr. L. Von Tersch as president. He has been particularly active in national and regional IRE activities, and has served as Chairman of the Chicago Section. He is active in professional societies and is a member of the National Society for Professional Engineers, Eta Kappa Nu, Tau Beta Pi, the ASEE, and the ISA.

Re-elected as NEC chairman of the board is W. O. Swinyard, vice president of Hazeltine Research, Inc. He is a past president of NEC and was a member of the original group responsible for the organization of the Conference.

Appointed committee chairmen are: arrangements, C. F. Valach, Amphenol-Borg Electronic Corp.; awards, Dr. M. E. Van Valkenburg, University of Illinois; exhibits, J. S. Powers, Bell & Howell; fellowship awards, Dr. D. S. Gage, Northwestern University; housing, R. R. Foley, Bell & Howell; industrial advisory, Dr. R. M. Soria, Amphenol-Borg Electronics Corp.; international activities, Prof. G. E. Anner, University of Illinois; planning, J. S. Johnson, Wayne State University; procedures, S. I. Cohn, Armour Research Foundation; Proceedings, J. L. Asmuth, University of Wisconsin; program, Dr. W. L. Firestone, Motorola, Inc.; registration, Dr. E. W. Ernst, University of Illinois; student activities, L. J. Murphy, Illinois Bell Telephone Co.; trust advisory, R. R. Jenness, Northwestern University; NEC party, H. M. Sachs, Armour Research Foundation.

More than 12,000 registrants—a record attendance—are expected at the 17th annual NEC.

The National Electronics Conference is a nonprofit organization serving as a national forum for the presentation of authoritative technical papers on electronic research, development, and application. Sponsors include the AIEE, Illinois Institute of Technology, the IRE, Northwestern University and the University of Illinois. Participants include the EIA, the Society of Motion Picture and Television Engineers, Iowa State University, Marquette University, Michigan State University, Purdue University, University of Michigan, University of Notre Dame, University of Wisconsin, and Wayne State University.

NAECON Plans Spring Meeting

Arthur J. Wilde, President of NAECON, has announced that plans for the Thirteenth Annual Conference are near completion. In keeping with the times and the Conference activities, the name of this annual conference is now the National Aerospace Electronics Conference. It will be held at the Miami and Dayton Biltmore Hotels, Dayton, Ohio, from May 8–10, 1961. The theme will be "Electronic Technology in the Aerospace Age."

Jointly sponsored by the Dayton Section of the IRE and the Professional Group on Aeronautics and Navigation (PGANE), this year's NAECON will offer a program of value to everyone interested in avionics. A new feature, a classified forum, sponsored jointly by NAECON and the Wright Air Development Division of ARDC, has been added to the conference. It was felt that NAECON attendees would welcome a forum that would give them a better understanding of the future of the military applied research program. Outstanding scientists of the Air Force will make presentations intended to answer questions such as, "What is electrical propulsion, bionics, and molecular engineering?" and "What about materials research for avionics development?" Attendees must be U. S. citizens with secret clearance. The required forms will be distributed with the NAECON advance brochures. The classified forum will be held at the WADD auditorium and seats will be limited to 500, attendance being granted on a first-come, first-served basis. For additional information, contact: R. Nordlund, Executive Vice-President, WWRNG, Wright Air Dev. Div., WPAFB, Dayton, Ohio.

The technical papers sessions will be under the leadership of R. Stimmel, WWRNC, WADD. Among the already selected technical sessions are the following: Solid State Devices, Radio Astronomy, Energy Conversion Systems, Molecular Electronics, Space Systems Integration, Bionics, Antennas and Propagation, Computers and Telemetry, Magnetohydrodynamics and Electric Propulsion.

Additional exhibit space has been acquired to handle the increased demands of industry to display its wares before the Wright Air Development Division and the Aerospace Industries.

As usual, the conference will have its lighter side with entertainment in the form of informal gatherings, luncheons, banquets and the annual ball.

For additional information on any aspects of the conference, contact: NAECON, 1414 E. Third St., Dayton, Ohio.

PGVC Announces Awards Presented At Annual Conference

The Eleventh National PGVC Conference, which was held at the Sheraton Hotel, Philadelphia, Pa., on December 1–2, 1960, matched the record of successful PGVC Conferences of recent years.

Advance registration, as well as last-minute decisions by many members to participate in this two-day conference, moved the attendance beyond the expected total. The exhibits represented efforts of over twenty companies in the vehicular communications field in an equipment display of varied interest to the professional engineer, management, and the user public.

The formal program included four technical sessions highlighting the theme, "New Horizons in Vehicular Communications."

After the dinner which was held on Thursday evening, R. P. Gifford, Chairman of the Administrative Committee, presented the following awards for recent papers:

First Prize—L. G. Schimpf, "Applications of Semiconductors in an 860 MC Receiver."

Honorable Mention—D. S. Dewire and E. A. Steere, "Precision Carrier Frequency Control and Modulation Phase Equalization of Base Transmitters in a Mobile Radio System."

Honorable Mention—A. V. Korolenko and N. H. Shepherd, "Can SSB Provide More Usable Channels in the Land Mobile Service?"

Dr. L. H. Stegg of the General Electric Company was the after-dinner speaker, and delivered a talk with films on missiles and space travel in the next twenty years.

The main feature of the second day was a luncheon during which Dr. G. H. Brown, vice president of engineering at RCA, spoke about trends toward miniaturization of electronic equipment in which the transistor has played a major role.

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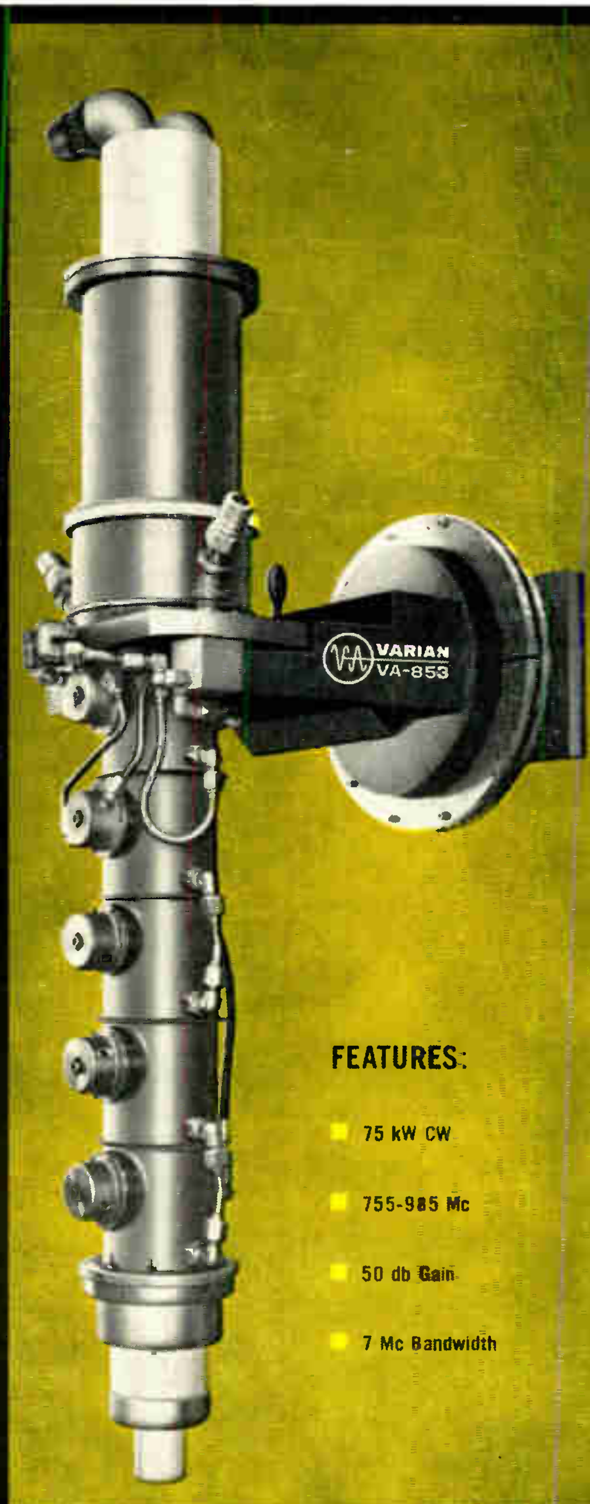

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Air Force MARS Announces Schedule

The schedule for the next month of broadcasts of the Air Force MARS Eastern Technical Net, operating Sundays from 2 to 4 P.M. EDT, has been announced as follows:

February 19—"Electronic Test Equipment Using Sonic Principles," R. Gunderson, Editor, Braille Technical Press. Mr. Gunderson is a designer of all types of meters, decade boxes, tuning devices and bridges which enable the blind to make precise measurements of passive, as well as operating circuits. These instruments will be described and demonstrated.

February 26—"Patriotic Rearmament Through Education," Vice Admiral R. B. Pirie, Deputy Chief of Naval Operations for Air. Admiral Pirie, former Commandant of the U. S. Naval Academy and presently Deputy Chief of Naval Operations, unfolds the challenge of America's future and the objectives to be sought by our science majors.

March 5—"Physics and Chemistry of Pure Metals," Dr. C. R. Kelly, Jr., Central Technical Services, Westinghouse Res. Labs. In semiconductors, parts per billion of impurities are known to have controlling importance. Dr. Kelly describes techniques which achieve levels of impurity far less than one part per million, and discusses methods of measuring the effects on mechanical and electrical properties in the presence of such apparently insignificant contaminants.

March 12—"Semiconductors," Dr. A. I. Bennett, Westinghouse Res. Labs. The electrical properties and crystalline structure of semiconductors will be discussed; mechanisms of electrical conduction in the presence of impurities will be described. Dr. Bennett will portray the preparation of semiconductor single crystals, including some recent advances in techniques of crystal growth.

Adaptive Control Systems Symposium Held on L. I.

Garden City, N. Y., was the center of a three day symposium on Adaptive Control Systems beginning October 17, 1960. The program, sponsored by the Long Island Section of the IRE and cosponsored by ARDC, ONR, Polytechnic Institute of Brooklyn, and the local branches of AIEE, ISA, ASME and PGAC, was equivalent to a short, intensive course in adaptive systems.

During the opening session, which was guided by Dr. J. R. Ragazzini, several challenging and conflicting views were set forth by J. Aseltine of Aero Space Laboratories and Y. T. Li of Massachusetts Institute of Technology, Cambridge, on the controversial definition of the problem. This set the pace for the subsequent sessions. The audience participation was unusually high during the discussion periods following many of the papers. The discussions originated from so many sources that they could have, in most cases, continued long past the time available.

Several sessions were devoted to Analytical Techniques—Theory and Application. One of the most unique papers, delivered by Dr. G. Bekey of Space Technology Laboratories, on "Adaptive Control System Models of the Human Operator" demonstrated the transfer function of man and how the various manual control systems were built to complete the over-all desired transfer function when that of man was inserted in the proper place in the block diagram.

The final session, moderated by Professor J. G. Truxal, consisted of a panel composed of the chairmen of the previous sessions. The implications, major needs and ultimate utility of adaptive system technology were discussed.

The Honorable C. D. Perkins, Assistant Secretary of the Air Force in charge of Research and Development, gave an appropriate talk at the banquet. He discussed the distribution of the Air Force budget and the need for increased participation of industry in basic research and development.

The Proceedings of the Symposium are to be published in book form. For particulars, contact Pergamon Press, Attention: D. Raymond, 122 E. 55 St., N. Y., N. Y.

Mathematics Research Center Plans April Symposium

The Mathematics Research Center will conduct a Symposium on Electromagnetic Waves on April 10–12, 1961. The topics to be considered are the propagation of waves in anisotropic media, diffraction, antenna theory, and numerical methods by which investigations of these can be submitted to electronic computation. The plan is to discuss the topics both from the practical (engineering) and the theoretical (mathematical) standpoint.

The program (when it has been fixed) may be obtained from: Mathematics Research Center, United States Army, University of Wisconsin, Madison 6, Wis.

OBITUARIES

Charles M. Clothier (S'51–A'52–M'58), a senior engineer with Sperry Gyroscope Company's Surface Armament Division was among those killed in the tragic airliner collision over New York, N. Y., December 16, 1960. He had boarded the ill-fated TWA Constellation at Columbus, Ohio.

At Sperry, he had been engaged in systems analysis, radar development work, and work on related military electronic systems. He was credited with significant contributions to radar technology and had a pending patent application in the radar art dealing with a V-beam height indicator system.

He joined Sperry in 1952 following his graduation from Bucknell University, Lewisburg, Pa., where he received the B. S. degree in electrical engineering. He also had attended Wilkes College, Wilkes Barre, Pa., and the Massachusetts Institute of Technology, Cambridge. More recently, he had attended Adelphi College, Long Island, N. Y., where he was doing graduate work in applied mathematics. During World War II, he served for five years with the Air Force.

Mr. Clothier was active in professional groups of both the IRE and the AIEE concerned with aeronautical and navigational electronics and information theory.

John V. L. Hogan (A'12–M'12–F'15)(L), a co-founder of the IRE, died recently at the age of 70. He was chairman of the Board of Directors of the Hogan Faximile Corporation, New York, N. Y.

J. V. L. Hogan

An early radio amateur, Dr. Hogan was a laboratory assistant to Lee De Forest in 1906–1907. In 1910, after studying electrical engineering at the Sheffield Scientific School of Yale University, New Haven, Conn., he became a telegraph engineer for the National Electric Signaling Company. He was appointed manager of the company in 1917 when its name was changed to the International Radio Telegraph Company.

In 1921, he left the firm to become a consulting engineer, and in 1929, became President of Hogan Laboratories, Inc., a firm principally engaged in television and facsimile development.

Radio Station WQXR (then W2XR) of New York was founded, owned and operated by Dr. Hogan from 1934–1944, when *The New York Times* acquired it.

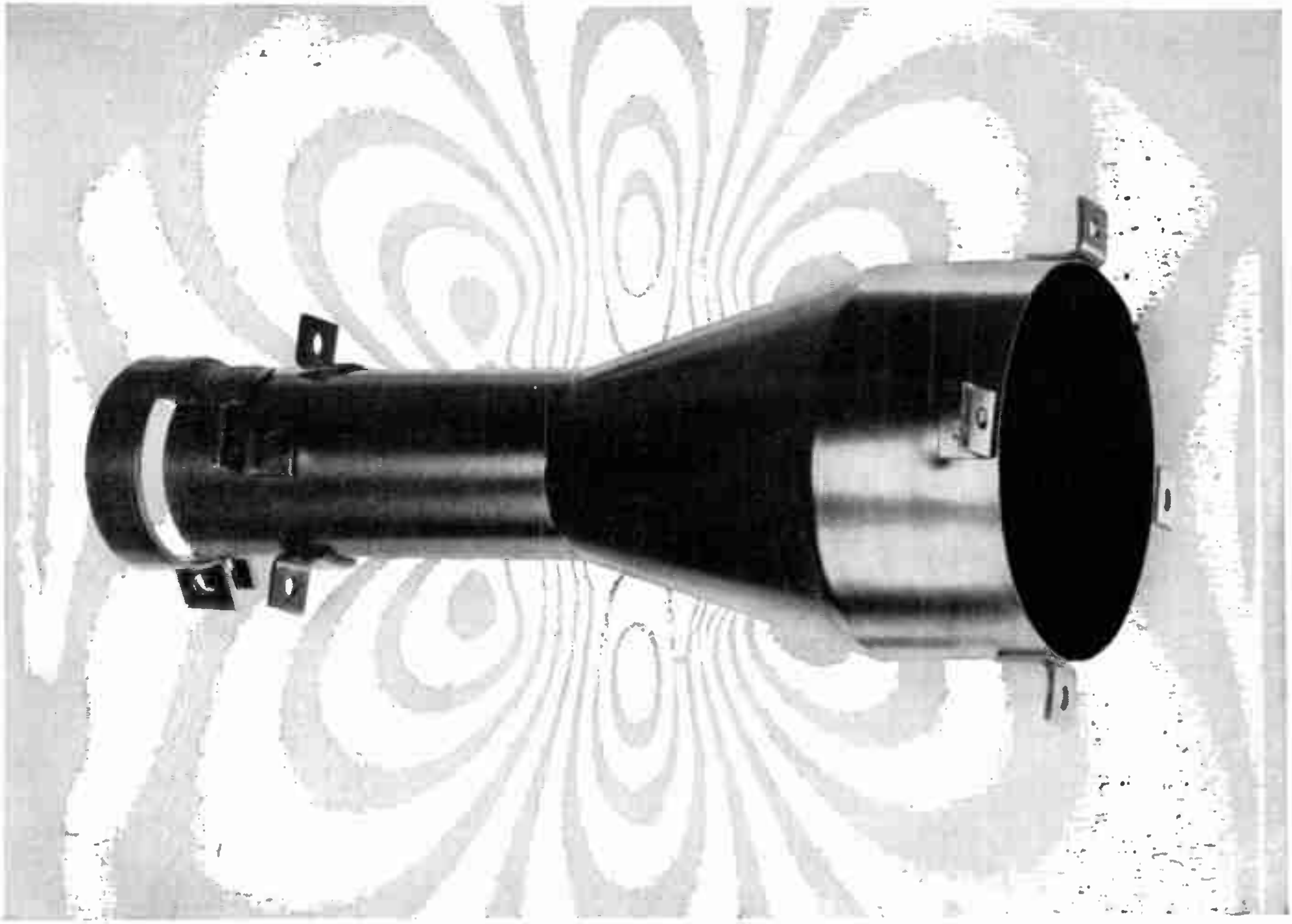
One of three original founders, he helped to combine the Society of Wireless Telegraph Engineers and the Wireless Institute to form the IRE in 1912. In 1920, he became its President. In 1956, he was awarded the IRE Medal of Honor. The Brooklyn Polytechnic Institute, Brooklyn, N. Y., awarded him the honorary degree of Doctor of Engineering in 1957.

Author of *The Outline of Radio*, Dr. Hogan contributed through numerous inventions to the radio, television and facsimile fields.

He had been a member of the Joint Technical Advisory Committee from 1948–1960, and had served as chairman of this committee on more than one occasion—most recently from 1959–1960. Also, he had been a member of the Patent Compensation Board of the U. S. Atomic Energy Commission, and served on the Special Technical Advisory Group under the Joint Chiefs of Staff and the Department of Defense, and as chairman of the Research and Development Board.

During World War II, he was special assistant to Dr. Vannevar Bush, Head of the Office of Scientific Research and Development.

Dr. Hogan had been a member, in various capacities, of the Board of Directors of the IRE from 1912–1936 and from 1948–1950. His service to the IRE and his contributions to its development extended over a period of nearly a half-century. During this time, he was a member of over twenty-seven IRE committees, frequently as chairman of these bodies.



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George W. O. Howe (F'55), who was born in 1875 at Charlton, Kent, England, died at Glasgow, Scotland, in November, 1960, at the age of 84. He received his technical education at the Woolwich Polytechnic Institute, and during this seven-year period was an apprentice with the firm of Messrs. Siemens Brothers, Woolwich. Later he went to Armstrong College, Durham University, from which he graduated with honors in electrical engineering in 1900. After spending two years with Messrs. Siemens & Halske at Charlottenberg, Germany, he returned to England as a lecturer at Hull Technical College. In 1909, he became assistant professor at the City & Guilds Engineering College, London, a position he held until 1920, when he was appointed head of the electrical standards and measurements department of the National Physical Laboratory. In 1921 he was elected as the first James Watt Professor of Electrical Engineering at the University of Glasgow. He held this position until his retirement in 1946. In 1947, he was made an honorary LL.D. of Glasgow University.

G. W. O. Howe

During his career, Professor Howe carried out research on the high frequency resistance of coils, the properties of condensers, and on the propagation of radio waves from transmitting antennas. He read several papers on these subjects before the Royal Society of Arts, British Association, Physical Society and the Institution of Electrical Engineers, of which latter body he had served at one time as chairman of the Wireless Section. In 1956 he was awarded the Faraday Medal of the Institution of Electrical Engineers for his pioneering work in the study and analysis of high frequency oscillations, on the theory of radio propagation and for his outstanding contributions to engineering education.

In addition to his academic work, he was a critical technical writer of considerable distinction. From 1926–1954 he was a consultant and technical editor of the *Wireless Engineer*, published in London. During this period, he wrote a monthly editorial on some topic of electrical, and usually radio, interest. The range of subjects discussed was very great, from specialist studies of radio to some discussions of the basic concepts of electrical science.

Dr. Howe was one of the small band of pioneers in the field of radio, and the professional engineer of today is greatly indebted to the contributions to this field which he made over a period of nearly half a century.

Second Symposium on Engineering Aspects of Magnetohydrodynamics

University of Pennsylvania, Philadelphia, March 9–10, 1961

Under the joint sponsorship of the AIEE, the AIR, the IRE and the University of Pennsylvania, the Second Symposium on Engineering Aspects of Magnetohydrodynamics will be held at the University Museum, University of Pennsylvania, Philadelphia, on March 9–10, 1961.

The primary purpose of the symposium is to exchange information among related disciplines involving magnetohydrodynamics. It is assumed the participants will have some degree of competence in one or more of the broad areas of program coverage. No tutorial sessions or papers are planned.

The program of the symposium and advance registration forms will be mailed to the symposium mailing list early in February. Names will be added to the list upon request to: C. Manual, G. E. Missile and Space Vehicle Dept., 3750 "D" St., Philadelphia 24, Pa.

A tentative program of papers to be presented is as follows:

Flight Applications Session

"Some remarks concerning MHD applications to the re-entry problem," *R. X. Meyer, Space Technology Labs.*

"Continuous duty dc accelerators," *W. Powers, Avco-Everett Res. Lab.*

"Experimental magnetogasdynamic engine for argon, nitrogen and air," *S. T. Demetriades, Northrup Corp.*

"The design, fabrication and test of a pulsed pinch plasma engine for space propulsion applications," *J. Pearson, C. Cavalconte, W. Guman and I. Granet, Republic Aviation Corp.*

"Experiments in steady-state high-density plasma acceleration," *A. F. Carter, G. P. Wood, A. P. Sabol and R. H. Weinstein, NASA Langley Field Res. Ctr.*

"Heat transfer to cold electrodes in a flowing ionized gas," *J. S. Fay and W. T. Hogan, Dept. of M.E., M.I.T., Cambridge.*

Power Conversion Session

"MHD power research in the United Kingdom," *B. C. Lindley, Parsons, Ltd.*

"Vortex generators," *C. P. Donaldson, Aero. Res. Assoc. of Princeton.*

"MHD Power Studies," *B. Blackman, MHD Res. Corp.*

"Measurements of the electrical conductivity of flame gases seeded with alkali metals and application to MHD power plant designs," *Dibelius, Luebke, and Mullaney, GE Co.*

"An electrodeless MHD generator," *Bernstein, Fanucci, Fischbeck, Jarem, Kulsrud, Lessen, and Ness, Forrestal Res. Ctr. and RCA Labs.*

"Progress in MHD power generation," *Brogan, Kantrowtiz, Rosa and Stekly, Avco-Everett Res. Lab.*

"Electrode boundary layers in direct current plasma accelerators," *J. L. Kerrebrock and F. E. Marble, Calif. Inst. Tech., Pasadena.*

Communications and Diagnostics Session

"Space charge waves and plasma diagnostics," *A. W. Trivelpiece, E.E. Dept., Univ. of Calif., Berkeley.*

"Diagnostic measurements of a highly ionized, steady state plasma," *A. L. Gardner, Lawrence Radiation Lab., Livermore.*

"RF reradiation spectrum broadening of a monochromatically excited plasma sheet due to the random motion of plasma constituents," *S. Zivanovic, Bendix,*

"Plane magneto-acoustic wave propagation through a stratified medium," *Gen, City College of New York, New York, N. Y.*

"Some effects of a nonlinear electrical conductivity on propagation of electromagnetic waves," *M. Epstein, U. S. C. Engineering Center.*

"Communication through a magnetoplasma," *R. L. Phillips and R. G. DeLosh, Bendix Systems Div.*

Fusion Session

"Plasma and fusion reactors," *M. B. Gottlieb, Project Matterhorn, Forrestal Res. Ctr.*

"Some aspects of economics for fusion reactors," *R. F. Post, Lawrence Radiation Lab., Livermore.*

"Thermonuclear power and super-conductivity," *R. G. Mills, Project Matterhorn, Forrestal Res. Ctr.*

"The hollow cathode arc," *L. D. Smullin, W. D. Getty, and A. Starr, E.E. Dept. and Res. Lab. of Electronics, M.I.T., Cambridge.*

"Magnetohydrodynamic shear heating," *M. C. Gourdine, Plasmadyne Corp.*

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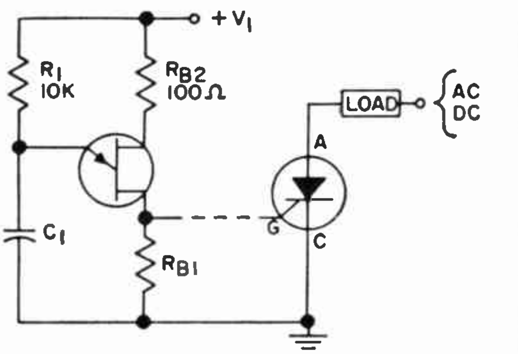

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Transistor Time Delay Circuit	Unijunction Circuit Equivalent
3 transistors*	1 unijunction transistor
1 diode	1 Zener diode
1 Zener diode	1 relay
1 relay	1 capacitor
2 capacitors	4 resistors
8 resistors	

SAVINGS: $3.30 (*germanium transistors)
$16.50 (*silicon transistors)

Transistor Voltage Detector	Unijunction Circuit Equivalent
2 transistors*	1 unijunction transistor
1 potentiometer	1 potentiometer
2 capacitors	1 capacitor
5 resistors	3 resistors

SAVINGS: $0.40 (*germanium transistors)
$9.20 (*silicon transistors)

Types 2N1671, 2N1671A, and 2N1671B are the latest additions to the General Electric line of silicon unijunction transistors . . . at a price comparable to germanium transistors. A unique type of semiconductor device proved in over three years of production and application, silicon unijunction transistors feature stable negative resistance characteristics, extremely low trigger currents, stable trigger voltages, and high pulse current capabilities.

The 2N1671 is intended for general purpose industrial applications where circuit economy is the prime consideration. The 2N1671A features a guaranteed minimum pulse amplitude, and is characterized for use in firing circuits for Silicon Controlled Rectifiers. The 2N1671B features low emitter leakage current and low trigger current, making it the most sensitive semiconductor triggering device available.

Price reductions have also been made on standard unijunction types 2N489 through 2N494. For complete technical data and application information for all unijunction transistor types, see your G.E. Semiconductor District Sales Manager. Or write section 54B87. Semiconductor Products Dept., General Electric Company, Electronics Park, Syracuse, New York. In Canada: Canadian General Electric Company, 189 Dufferin Street, Toronto, Ontario. Export: International General Electric Company, 150 E. 42nd St., New York, N. Y.

CONDENSED SPECIFICATIONS OF UNIJUNCTION TRANSISTORS 2N1671, A, B

Maximum Total Power Dissipation		450 mw
Opeating Temperature Range		—65°C to 140°C
Intrinsic Standoff Ratio (η)		0.47 to 0.62
Interbase Resistance (R_{BB})		4.7K to 9.1K
Peak Point Emitter Current (I_P)	2N1671,A	25 μa max.
($V_{BB}=25$ volts)	2N1671B	6 μa max.
Emitter Reverse Current (I_{EO})	2N1671,A	12 μa max.
($V_{EB2}=30$ volts)	2N1671B	0.2 μa max.
Base—One Peak Pulse Voltage	2N1671A,B	3 volts min.

(In SCR circuit shown, $V_1=20V$, $C_1=0.2$ μfd, $R_{B1}=20\Omega$)

For fast delivery at factory-low prices, call your G-E distributor.

Progress Is Our Most Important Product

GENERAL ELECTRIC

Industrial Engineering Notes*

Associated Activities

The EIA-sponsored educational television experiment at Hagerstown, Md. is currently under critical evaluation from a cost standpoint by the County Commissioners of Washington County, Md. As the five-year experimental project comes to a close next summer, the Washington County Commissioners must decide whether to include sufficient funds in the county's school budget to cover operating costs of the educational television system which was established at the beginning of the 1956 school year. EIA originated the program and its members donated the necessary transmitting, distributing and receiving equipment and components. The project was supported financially by the Ford Foundation on a five-year basis in cooperation with the Washington County School Board. While no definite date has been set for the actual determination by the Washington County Commissioners as to the future support of the project, the subject is now under intense consideration by the people of Hagerstown and Washington County. Confronting the County Commissioners is the question of financing the continuation of TV teaching—or of scrapping the TV equipment assembled in Hagerstown, the carefully acquired "know-how," and the staff of skilled TV instructors and technicians. School Superintendent W. M. Brish calls the Hagerstown results "exceedingly good." Superintendent Brish and the school board have asked the Commissioners to allocate funds for TV in next year's school budget. They estimate the cost at $90,000 for the last six months of 1961 and about $181,000 a year thereafter. Mr. Brish and the school board point out that it will cost more money to drop the project than to continue it. It was emphasized that ending of TV instruction would require the hiring of enough additional teachers to wipe out any financial saving which would accrue. Nearly 20,000 students in the county have been receiving instruction via television.

Government and Legislative

The Federal Aviation Agency announced a revision of the Very High Frequency Deployment Plan which will extend to January 1, 1966, the date when the FAA will implement unrestricted channel assignments using 50 kc separation. A Technical Standard Order will not be adopted as a requirement for communication transmitters and receivers for general aviation aircraft, FAA said. Instead, a guide for general aviation communications equipment manufacturers is being developed by Special Committee 93 of the Radio Technical Commission for Aeronautics. The Federal Communications Commission requirements for transmission stability and quality govern general aviation communication equipment. The aviation agency said the plan provides for continuing both VFR and IFR (below 24,000), communications service on the 100 kc channels between 118–127 Mc until January 1, 1966. VFR communications service will continue thereafter on the 100 kc to the extent feasible. The plan continues to provide service on 50 kc channeling in the 127–135 mc band for users with this tuning capability. Beginning January 1, 1961, frequency assignments of 50 kc separation will be progressively implemented below 127 Mc on a case-by-case basis, as may be required for both terminal and en route services. However, until January 1, 1966, 50 kc channel assignments will in most cases be installed in high density traffic areas. Starting in 1966, unrestricted channel assignments using 50 kc separations will be implemented for all communication functions as air traffic control requirements may dictate. Airport control towers and flight service stations will continue to provide communications service for VFR traffic on 100 kc channels to the extent feasible. . . . **Shipment of electronic components continued at first quarter levels during the second quarter of 1960, the Electronics Division of the Business and Defense Services Administration has reported.** A decline in shipments of receiving tubes, transistors, capacitors, and connectors was counterbalanced by increased shipments of diodes and rectifiers, power and special purpose tubes, television picture tubes, transformers, and quartz crystals. Shipment of other components continued at first quarter levels. For the first time since the Electronics Division started collecting these data, output of semiconductor devices (transistors, diodes and rectifiers, and related devices) failed to increase. . . . **The National Bureau of Standards has established an Advisory Committee on Engineering and Related Standards to handle national needs for such standards as codes, specifications, standard test methods, and standard data on the properties of engineering materials.** The committee, which consists of three members each from the American Standards Association and the American Society for Testing Materials, will also be concerned with maintaining awareness of the efforts of private organizations in the standardization areas, fostering cooperative programs, and recommending use of the bureau's special competence for the development and application of engineering and related standards. . . . **A move which would have the effect of making obsolete most existing television sets**

(Continued on page 24A)

* The data on which these Notes are based were selected by permission from *Weekly Reports*, issues of November 28, December 5 and 19, 1960, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

SWEEP SIGNAL SOURCES

X775A

DELIVERY FROM STOCK

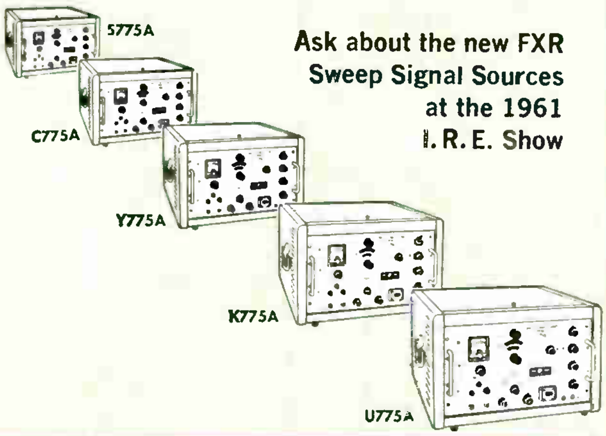
- Direct Reading Frequency Dials for Setting Upper and Lower Band Ends
- Integral AGC Circuit Keeps Output Flat to ±0.5 db
- AGC Provision Eliminates Need for Slow Response Ratiometer Set-ups
- Convenient, Portable, Versatile

for fast, visual reflectometer tests ranges from 2 to 40 KMC

Another breakthrough in measuring convenience has been achieved by the newly expanded FXR family of self-contained, direct reading sweep signal sources. Coverage is now provided as high as 40 KMC. Each unit utilizes a permanent magnet BWO as the tunable RF source.

Output power can be equalized at any detection point relative to the response of the detection element through use of FXR's exclusive built-in AGC circuit. This circuit provides a flat (±0.5 db level) on modulated signal throughout the swept frequency range when used with matched bolometers and directional couplers. This AGC provision eliminates the need for using slow response ratiometers, and allows for visual VSWR or Reflection Coefficient tests.

FXR FAMILY OF ALL ELECTRONIC SWEEP SIGNAL SOURCES



Ask about the new FXR Sweep Signal Sources at the 1961 I. R. E. Show

Model Number	Frequency Range (KMC)	Approx. Minimum Power Out.	OUTPUT Waveguide Type	OUTPUT Connector	Price
S775A	2.0-4.0	70 mw	(⅜" Coax Type N)		$2750.
C775A	4.0-8.0	20 mw	(⅜" Coax Type N)		$2800.
X775A	8.2-12.4	20 mw	WR-90	UG-39/U	$2900.
Y775A	12.4-18.0	10 mw	WR-62	UG-419/U	$3300.
K775A	18.0-27.0	5 mw	WR-42	UG-595/U	$3700.
U775A	27.0-40.0	5 mw	WR-28	UG-599/U	$4300.

Characteristics and prices subject to change without notice.

GENERAL SPECIFICATIONS

SWEEP RATE (Resolution): 0.3 to 300 KMC/sec linear with time
SWEEP WIDTH: approximately 200 KC to full frequency range
OUTPUT SIGNAL: CW, square wave (internal 600 to 1200 cps)
FREQUENCY DIAL ACCURACY: ±1% for fixed frequency operation
±2% for sweep frequency operation
POWER REQUIREMENTS: 115/230 v., 50/60 cycles, 200 w.
DIMENSIONS: 12⅞" high x 21¾" wide x 18" deep
WEIGHT: 78 lbs.

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REPRESENTATIVES IN ALL MAJOR CITIES THROUGHOUT THE WORLD

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PRECISION MICROWAVE EQUIPMENT • HIGH-POWER PULSE MODULATORS • HIGH-VOLTAGE POWER SUPPLIES • ELECTRONIC TEST EQUIPMENT

Industrial Engineering Notes

(Continued from page 22A)

was proposed by Federal Communications Commissioner R. E. Lee at the Winter Conference as a means of making transmission channels available for some 200 stations now kept off the air for lack of broadcast frequencies. Specifically, the commissioner advocated shifting all TV transmission into ultra high frequency bands within 5 to 7 years, or approximately the life of TV receiving sets. Most TV transmission now is at very high frequency and only about 6 per cent of the approximately 60 million TV sets now in use can receive UHF signals. The military services, he pointed out, twice have turned down FCC pleas for more VHF space "in no uncertain terms and for good and sufficient reasons." The commission's long-range-50-channel VHF plan, he said, "is definitely out the window." "I have yet to find an FCC engineer who does not believe that the only answer to the problem of getting new TV channels is a gradual shift to UHF," the commissioner said. He added that crowding more stations into presently available frequencies through channel splitting, as has been proposed, would result in interference which, in the opinion of engineers, would be "intolerable." Mr. Lee said that there is a "good chance" Congress will pass FCC-sponsored legislation which would have the effect of requiring TV manufacturers to produce only sets capable of receiving both UHF and VHF signals. He admitted the proposal runs counter to his opposition to Government interference with private enterprise, but contended the legislation to be justified in view of the "critical nature of the present situation." The 12 channels now being used by TV, he said, could be assigned, following the shift to UHF, to mobile radio communications services now badly needed by industry, police, fire, forestry, educational institutions and "innumerable private and public agencies which touch upon our welfare in every conceivable way." . . . **An electronic computer-plotter that mechanically draws a complete weather map in less than three minutes has been put into regular Weather Bureau service.** The "weather plotter" produces a completed Northern Hemisphere weather map about seven times faster than the former hand drawing method. The new automatic electronically controlled method produces maps said to be more accurate than handdrawn maps. The electronic unit reads weather information supplied in numerical form on magnetic tape, presents the information to a digital-to-analog converter that in turn instructs a plotting board to automatically draw contours on a 30-by-30 inch map of the Northern Hemisphere. The plotter is equipped with a mechanical "hand" which guides an inked stylus on instruction from the analog impulses. The weather plotter was developed and produced by Electronic Associates, Inc. of Long Branch, N. J. Two units, one of which will be used on a stand-by basis, are now at the National Meteorological

(Continued on page 26A)

Industrial Engineering Notes

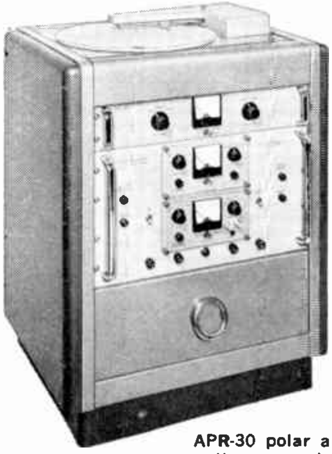
(Continued from page 24A)

Center. . . . **The electronics industry may well become the nation's top industry in the next decade, Secretary of Commerce Frederick H. Mueller told the 14th Annual Conference of Bank Correspondents in Chicago recently.** "Most people—even businessmen—do not realize that the electronics industry today is our fifth largest industry; it could become first in size in the next 10 years," said Mr. Mueller. The Commerce secretary called the application of the science of electronics the most far-reaching change in modern private industry since the development of steam, electricity, and oil. "It already is revolutionizing industry by presenting the advance guard of the completely automatic factory, automatic office, and earth-controlled vehicles." Working toward the "automatic office" of the future, the Commerce Department's Bureau of Standards has developed high-speed electronic computers which handle large, complex figures "several thousand times faster" than an expert mathematician with a desk calculator, Mr. Mueller said. One of these, the SEAC, is solving scientific problems of ballistics and atomic energy in a matter of seconds and at an "enormous" reduction in labor costs. Mr. Mueller said NBS engineers are studying the possibilities of applying SEAC to banking, insurance, mail order houses, and other large record collecting agencies, as well as government censuses, social security filings, tax listings and military records. . . . **The Federal Communications Commission has informally decided to press for early passage of the bill it sponsored last year to force TV set manufacturers to make all-channel receivers only as a means of encouraging the establishment of more UHF stations.** This agreement was reached recently at a Commission meeting, although no official announcement of the decision has been made. Last year the FCC was blocked in its efforts to push the legislation because of an adverse report from the Department of Commerce. Whether this opposition will be withdrawn under the new Administration is not known. Normally, the U. S. Budget Bureau will not permit contradictory recommendations to go to Congress from different executive agencies. The EIA Consumer Products Division has decided to oppose the legislation and testify against it if given the opportunity on the ground that it would compel consumers to pay more for TV sets, whether or not they wanted or could use the UHF tuner, and because it would give the FCC authority to establish commercial standards for receivers. It was also reported that the FCC plans to encourage present VHF stations to simultaneously transmit on UHF and to create a "pool" of UHF channels by removing them from the table of allocations and alloting them when needed. . . . **Authorization for a new experimental pay TV station in Los Angeles has been granted by the Federal Communications Commission.** The Commission issued a permit for

(Continued on page 30A)

Flexibility and Refinement

APR-20 rectangular antenna pattern recorder



APR-30 polar antenna pattern recorder

...The reason most antenna pattern recorders come from S-A

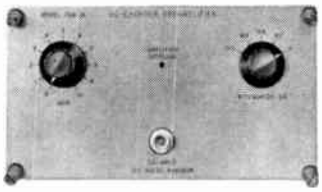
It's the little things that make the difference. Little things, refinements, "extras," and top-notch workmanship all add up to preference for S-A instrumentation.

Things Like Plug-In Balancing Potentiometers...

Series P plug-in pen balancing potentiometers

Series P potentiometers are used in both rectangular and polar coordinate pattern recorders. By interchanging potentiometers together with the appropriate pen function amplifier, different responses—linear, square-root, and logarithmic—are obtained. Interchanging these new self-aligning potentiometers can be accomplished in less than thirty seconds. Stocking spare units cuts downtime. Of dust and dirt proof construction, Series P plug-in balancing potentiometers are offered with exchange pricing.

DC Amplifiers...



DCA-21 amplifier for dc input signals

Scientific-Atlanta's DCA-21 amplifier lets APR 20/30 recorders accept dc input signals. A narrow band amplifier preceded by an electromagnetic chopper, the sensitive DCA-21 has a linear dynamic range of 80 db. The unit is directly interchangeable with Series CBA-20 Crystal-Bolometer amplifiers.

Recorder Pen Programmers...

Up to five different pen writing codes can be selected by adding the Model RPP-1 Recorder Pen Programmer to an APR 20/30 installation. Compact, lightweight, and rack mounted, the programmer provides solid line, dot, dash, dash-dot, and space-dot-dot codes at an adjustable code rate of 30 to 90 cycles per minute.

Modification C, Chart Compression...

Modification C, which must be ordered at the time of recorder purchase, provides both standard and compressed cycle charts from a single APR 20 Rectangular pattern recorder. Standard chart cycle is 20 inches, compressed 8 inches. Compressed recordings are conveniently sized to fit standard 8½ x 11 notebooks and reports.

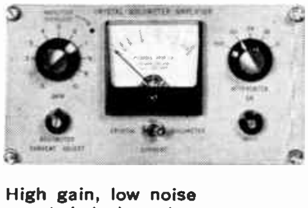
Chart Paper, Recording Pens, Ink, and Accessories...

Scientific-Atlanta offers its customers one-day service by stocking, for immediate delivery, a wide variety of chart paper, recording pens, and other recording necessities.

But above all, it's the engineering philosophy of a company run by antenna engineers for antenna engineers.

Call your nearby S-A engineering representative for more information on S-A pattern recorders and accessories. For complete technical information, please write to Box 86.

Crystal Bolometer Amplifiers...



High gain, low noise crystal-bolometer antenna

Sensitive, narrow-band Crystal-Bolometer amplifiers are miniaturized units designed for use as preamplifiers in S-A polar and rectangular pattern recorders. Five models, CBA-21 through CBA-25 are available. Features include bolometer burnout protection, low noise figure, triaxial signal ground return, up to 108 db gain, 80 db linear dynamic range, adjustable bandwidth (CBA-23), high rejection (CBA-24), variable center frequency (CBA-25).

SCIENTIFIC-ATLANTA, Inc.

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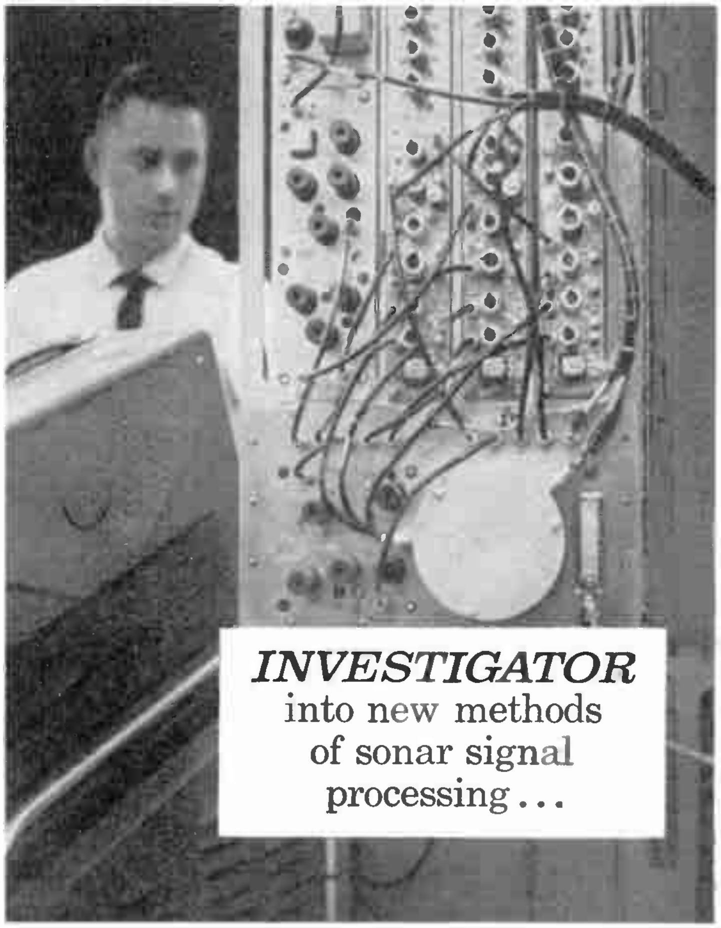
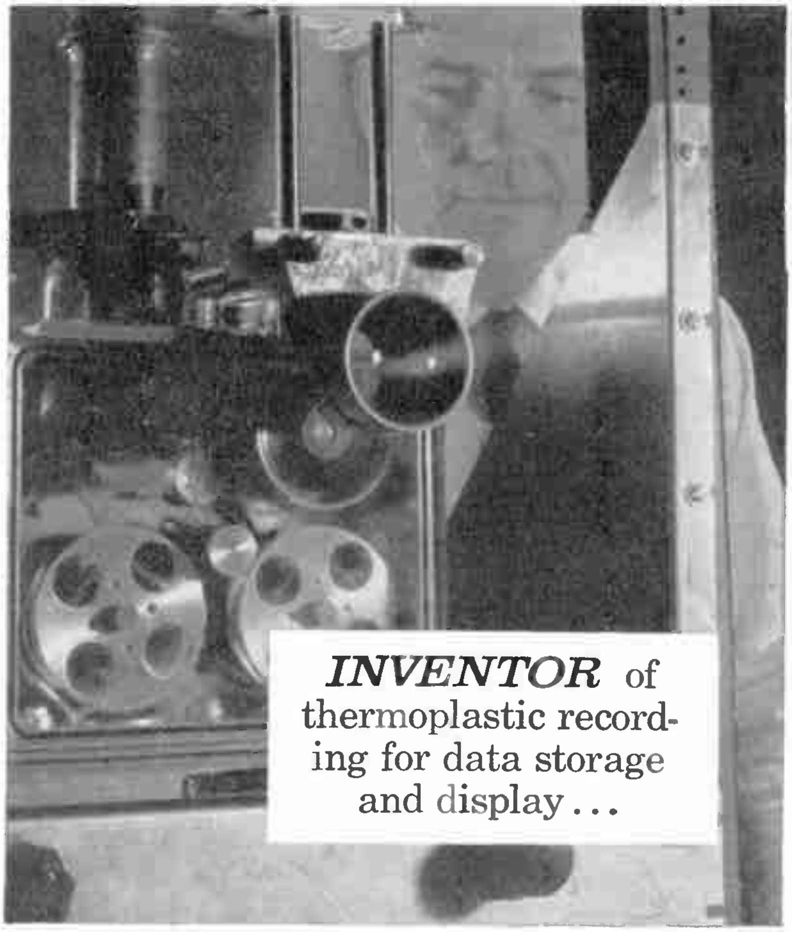


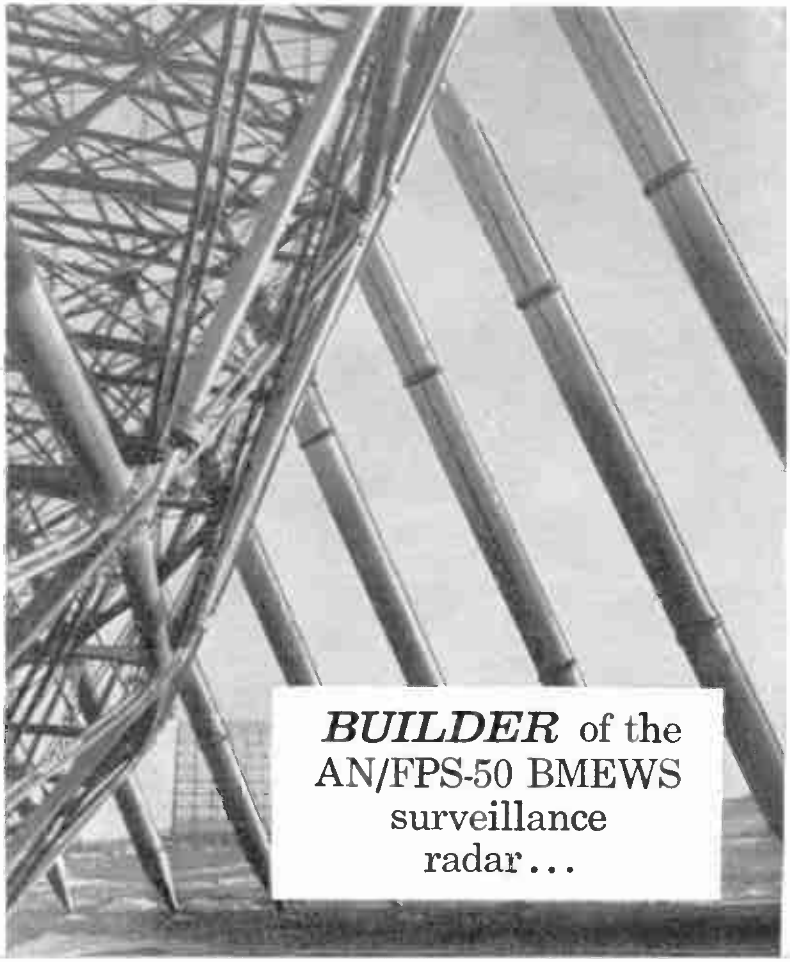
INVESTIGATOR into new methods of sonar signal processing . . .



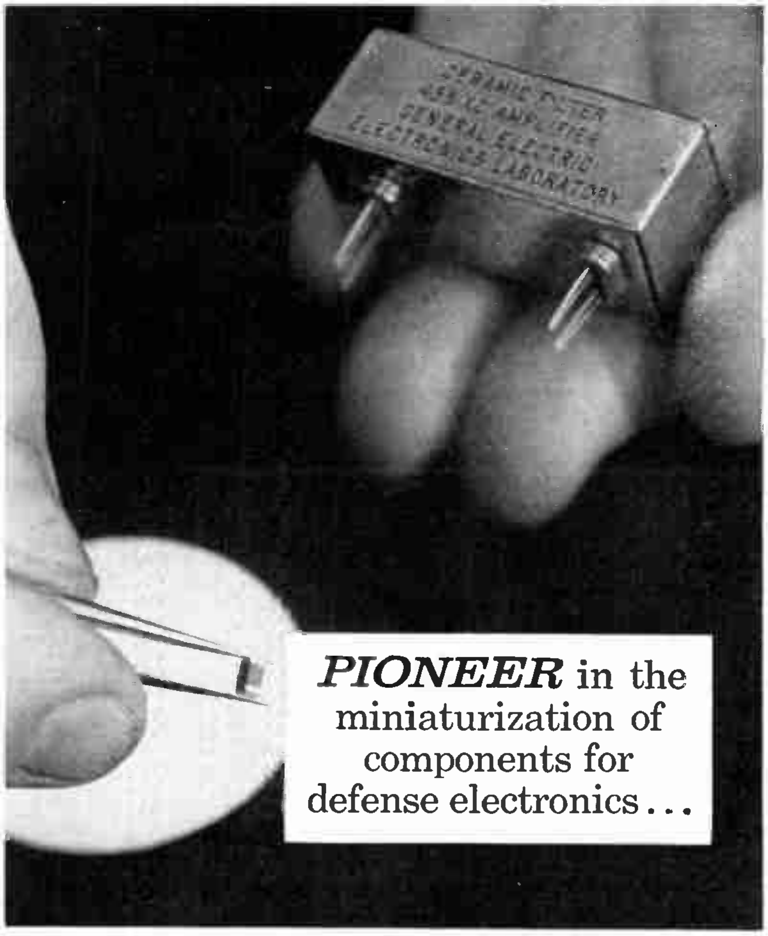
INVENTOR of thermoplastic recording for data storage and display . . .

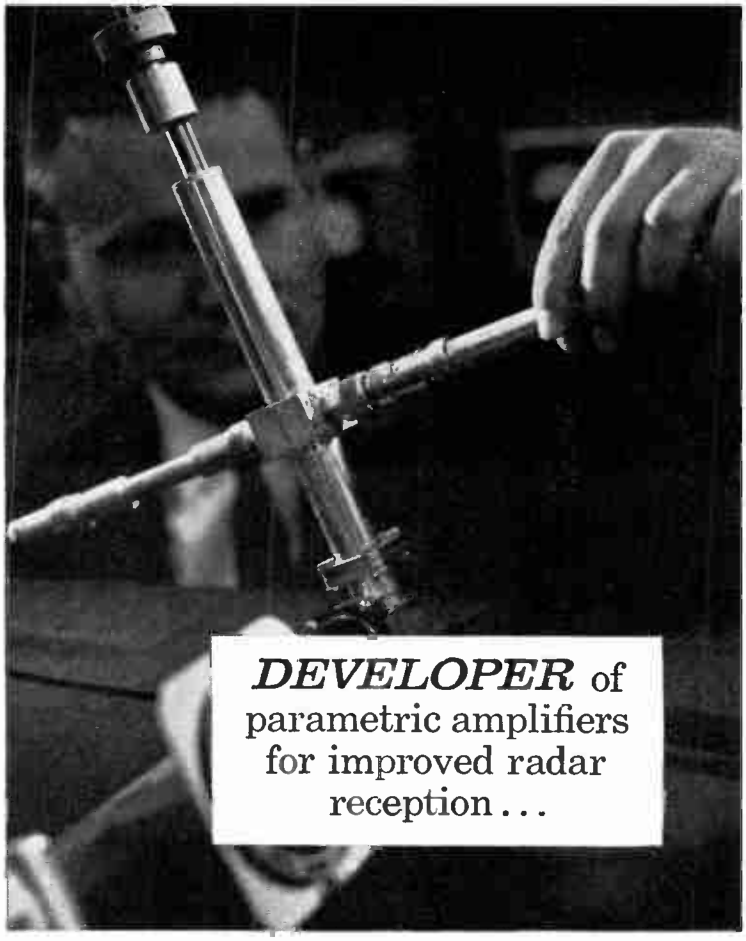
The Many

CONSTRUCTOR of the first fully automatic 3-D radar data processor . . .

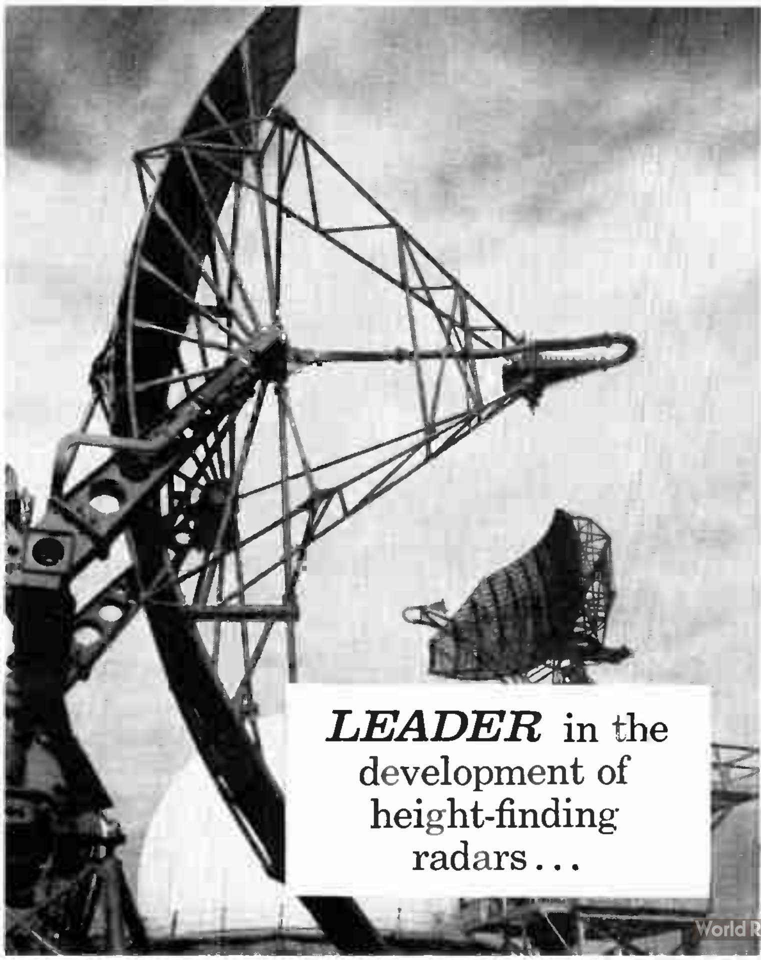


BUILDER of the AN/FPS-50 BMEWS surveillance radar . . .





Roles of HMED*



**General Electric's Heavy Military Electronics Department*

Seven examples of the roles that General Electric's Heavy Military Electronics Department is playing in contributing to U.S. defense strength are illustrated here. Of course, the full spectrum of HMED activities is much broader. It includes work in radar, sonar, missile guidance and control, and computers; in data handling, communications, counter measures, and ground warfare; in air defense, missile defense, and product service. 176-10

Progress Is Our Most Important Product

GENERAL ELECTRIC

DEFENSE ELECTRONICS DIVISION, SYRACUSE, N. Y.

Industrial Engineering Notes

(Continued from page 26A)

field tests of a system developed by the International Telemeter Co., a division of Paramount Pictures Corporation. Test broadcasts will be made over station KTLA between 1:30 a.m. and 8 a.m. There will be no public participation. Telemeter intends to install an experimental low power transmitter which will send scrambled signals which can be picked up in intelligible form only by the experimenter's receivers.

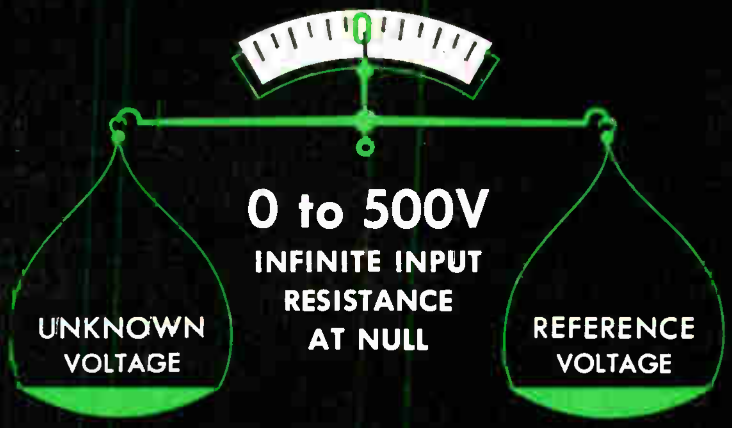
Industry Marketing Data

Electronic components manufacturing in the United States is now a $3 billion annual business—almost three times the output of a decade ago—the Business and Defense Services Administration, U. S. Department of Commerce, reported recently. In its first major study of the electronic components industries, BDSA's Electronics Division says that more than 40 per cent of the total output of these industries is now for military end-use—for the manufacture of military electronic equipment or for maintenance purposes. Most electronic component producers are small, and a relatively few firms account for most of the total output, according to a report of the study. About 75 per cent of the total output is produced in seven states. In 1959 almost 60 per cent of the total value of shipments originated in 16 metropolitan areas located in 11 states. While primarily presenting a picture of the industries in the past decade when they have experienced their greatest growth, the study gives some statistics bearing on the earlier days of electronics starting in 1923. In text and tables, the study shows such items as information on estimated 1959 output of all major components; detailed 1959 shipments data, in units and dollars for both military and nonmilitary end-use, for over 50 categories of electronic components; comparable value of shipments from 1952–59; variations in rate of growth since 1955; estimated numbers of firms, by product; concentration of output by product; output by state and metropolitan area; major markets; price trends; manpower materials, and capital requirements; and analysis of future trends. The report, "Electronic Components, Production and Related Data, 1952–59," is for sale by the Government Printing Office, Washington 25, D. C. Price: 20 cents. . . . **The EIA Marketing Data Department's totals for October showed factory sales of transistors decreasing under the year's high set in September, but remaining above the 12 million mark for the second month in a row.** A total of 12,168,632 transistors were sold in October, compared with the record 12,973,792 sold during the previous month. Revenue from October factory sales stood at $25,945,195, against the $28,442,229 total recorded for September. Through October, 102.4 million transistors have been sold as compared to 74.5 million

(Continued on page 32A)

a GOOD MEASURE of user benefits...

NEW



MODEL 801H

NEW! 1MV FULL SCALE NULL SENSITIVITY—RECORDER OUTPUT—AUTOMATIC LIGHTED DECIMAL—IN-LINE READOUT—TAUT BAND SUSPENSION METER

precision DIFFERENTIAL DC VOLTMETER

Using the same accurate principal to find the unknown as the beam balance, the Model 801H Differential Voltmeter gives you balanced accuracy, guarantees a "good measure for your money."

Like all jf differential voltmeters, the Model 801H provides infinite input impedance at null over the entire 0-500 Volt range. This jf feature is unique on today's voltmeter market and is of prime consideration when making precise DC measurements. The source loading of 1 to 10 megohms above a nominal 10 volts, which is inherent in other differential voltmeters now available, cannot be tolerated when 0.05% or better accuracy is to be maintained.

PARTIAL 801H SPECIFICATIONS

Voltage Ranges:	0.5, 5, 50 and 500V DC
Accuracy:	0.05% from 0.1 to 500V 0.1% or 50uv, whichever is greater, below 0.1V
Null Sensitivity Ranges:	10V, 1V, 0.1V, 0.01V and 0.001V
Maximum Meter Resolution:	5uv
Input Impedance:	Infinite at null
Dimensions:	Cabinet 9¾" W x 13½" H x 14" D Rack 19" W x 8¾" H x 13¼" D
Weight:	Cabinet 25 lbs., Rack 28 lbs.
Price:	Cabinet $555.00, Rack $575.00 Prices F.O.B. factory, Seattle

Extreme accuracy and stability are achieved by advanced circuit design which incorporates a chopper stabilized null amplifier and a standard cell reference.

Write for complete specifications.
Also ask for information on the jf A-70 recorder, companion to the 801H, and the Model 803.

Prices and technical data subject to change without notice.

SEE US AT BOOTH #3237
I.R.E. SHOW MARCH 20-23, 1961

JOHN FLUKE MFG. CO., INC.
P.O. BOX 7161 SEATTLE 33, WASH.

Industrial Engineering Notes

(Continued from page 30A)

during the same period in 1959. . . . **October was the best month since January for retail sales of monaural phonographs, with a total of 126,807 sets sold—10,944 more than in September, according to the EIA Marketing Data Department's latest figures.** Sales of stereo phonos also rose, but at a considerably slower rate which was first noted in September. In October, 272,101 stereos were sold, an increase of only 7,465 over the previous month's total. . . . **A ten-country survey of the market for electron tubes and semiconductors manufactured in the United States shows sales continuing at a good level the Business and Defense Services Administration reported recently** While the somewhat lower prices quoted on the competing products of foreign manufacturers have some impact in some areas, the reputation enjoyed by U. S. producers generally is holding up the trade in their products, according to a comprehensive survey by BDSA's Electronics Division. The survey constitutes the third and last report in a series based on material from Foreign Service dispatches, and brings the total countries studied to 28. The current report covers Australia, Canada, Japan, Netherlands, New Zealand, Spain, Taiwan, Turkey, Union of South Africa, and West Germany. The complete report of the survey, "Electron Tubes and Semiconductors—Production, Consumption Trade, Selected Foreign Countries," is for sale by the Government Printing Office, Washington 25, D. C. Price: 30 cents.

Military Electronics

Construction of the Strategic Army Communications System (STARCOM), designed to give almost split-second control over the far-flung Army commands, has been completed with the opening of the world's largest automatic relay station at Fort Detrick, Frederick, Md., the Army has announced. STARCOM can send a message through the East Coast Relay Station in three seconds. The world-wide communications network permits rapid flow of information necessary for commanders to make quick decisions and to take immediate action. The network includes radio relay station, communications centers and long distance radio, wire and cable circuits. The $25 million control station at Fort Detrick, largest in the world, makes the Army's communications network fully automatic in the continental United States. The East Coast Relay Station eventually will be a part of the Defense Communications Agency and will service messages for the other government agencies. The East Coast Relay Station completes the STARCOM network in the United States. Other stations are the Midwest Relay Station at Fort Leavenworth, Kan., and the West Coast Relay Station at Davis, Calif. The network control station can handle 275,000 messages a day. It can store 5,000 messages at a time. For high-speed automatic handling, messages are recorded on perforated tape and are converted back to printed messages at their destination. . . . . **Sixty-nine per cent of the Navy's prime contract awards during fiscal year 1960 went to 100 contractors, and 18 per cent of that total was allotted to electronics work, according to Rear Adm. D. F. Smith, the Navy's Chief of Information.** Adm. Smith told the Pittsburgh chapter of the Association of Industrial Advertisers that contracts worth $5.1 billion were awarded the top 100 firms. The highest expenditure was for aviation, at 48 per cent, followed by electronics. Fifteen per cent went for petroleum, 12 per cent for shipbuilding, and seven per cent for all other classifications. Of the grand total, small business received prime contracts worth $1.4 billion, according to Adm. Smith.

IRE People

Bertram D. Aaron (S'40–A'45–M'48–SM'54) has announced the reorganization and expansion of his field engineering and management consulting activities into Bertram D. Aaron and Company, Roslyn, L. I., N. Y.

B. D. Aaron

Since 1943, he has specialized in microwave theory and applications, for the past eight years as manufacturer's representative for Bomac Laboratories, Inc., and other firms, first in the Los Angeles region and more recently in the New York City Metropolitan area. Bertram D. Aaron and Company will continue to concentrate on microwave applications, but will also handle other compatible types of electronic equipment over an expanded territory that will include northern New Jersey and Long Island, as well as the New York Metropolitan area.

He received the degree in electrical engineering from Virginia Polytechnic Institute, Blacksburg, and spent several years in the U. S. Army Signal Corps and the NACA (now the NASA) at Langley Field, Va. He is perhaps best known in the microwave field for his work as Secretary in charge of the Defense Procurement Agency's Integration Committee on Hydrogen Thyratrons—commonly known as the Hydrogen Thyratron Council. He also directed several programs for the Signal Corps which set up production facilities for special types of film and certain military electronic equipment.

Mr. Aaron is at present Chairman of the Long Island Chapter's Professional Group on Microwave Theory and Techniques.

❖

Dictograph Products Inc. has appointed **Harold H. Beizer** (SM'58), President and Chief Engineer of Bellaire Electronics Inc., Red Bank, N. J., to the newly-created post of Vice President in charge of manufacturing and engineering. He will direct the company's new engineering, research and development program.

H. H. Beizer

Before joining Bellaire in 1951, he served nine years as a civilian employee of the Signal Corps and was chief engineer of the Signal Corps Engineering Laboratory in Detroit during World War II. He specialized in radio interference reduction and represented the U. S. Army on the Research and Development Board Panel on Radio Interference, the AIEE-IRE joint sub-committee on radiation measurement, and the American Standards Association Committee on interference measurement. He was also a United States delegate to the International Special Committee on Radio Interference of the International Electro-Technical Committee.

He was previously employed in various capacities from production to research and development with the Radio Corporation of America, General Electric Co., Philco Corporation and International Telephone and Telegraph Co.

Mr. Beizer is a member of the Armed Forces Communications and Electronics Association. He is a graduate of the Moore School of Electrical Engineering of the University of Pennsylvania, Philadelphia, and has the Master of Science degree in communications from the Massachusetts Institute of Technology, Cambridge.

❖

The election of **Dr. Lloyd V. Berkner** (A'26–M'34–SM'43–F'47), President of the IRE, 1961, as President of the Graduate Research Center, Dallas, Texas, was announced November 30, 1960, by J. E. Jonsson, Chairman of the Board of Trustees of the Center. Dr. Berkner will join the Center from Associated Universities, Inc., New York, where he has been President for ten years.

In announcing the election of Dr. Berkner as President of the Graduate Research Center, Mr. Jonsson stated that the Trustees foresee substantial increase in activities of the Center and development of large-scale research facilities in the Southwest. This development will be closely related to the needs of important colleges and universities in Texas and the neighboring states with the view to greatly extending the opportunities for graduate education and post-doctoral research.

During Dr. Berkner's tenure as its president, Associated Universities, Inc. managed the Brookhaven National Laboratory, Upton, L. I., N. Y., under contract with the Atomic Energy Commission and developed the National Radio Astronomy Observatory, Green Bank, W. Va., under contract with the National Science Foundation.

Dr. Berkner's photograph and biography appeared on page 2, as the Frontispiece, of the January, 1961, issue of Proceedings.

❖

(Continued on page 36A)

IRE People

(Continued from page 34A)

Narda Microwave Corporation, Mineola, L. I., N. Y., recently announced the appointment of **Adolph Brenner** (A'55), who will be in charge of the development of advanced designs of frequency meters, filters, and ridged waveguide equipment, as well as the improvement of attenuators, frequency meters, terminations detectors and other products.

He joined Narda as a project engineer from the City College of New York, N. Y., and has taken graduate work at the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. He has directed and carried out projects such as a television relay transmitter and receiver plumbing with automatic frequency control, directional couplers, direct reading frequency meters, reference cavities and broadband fixed-tuned bolometer mounts. He has also worked on projects such as harmonic mixers, externally tuned reflex klystron oscillators, UHF amplifiers, an ECM system, ridge waveguide components and K-band receiver RF head.

Mr. Brenner is a member of the New Type Waveguide Committee of the EIA, and Treasurer of the New York Chapter of the Professional Group on Microwave Theory and Techniques.

❖

John M. Norton (A'47–M'56) has been promoted to technical director in the Advanced Systems Development Division of the IBM Corporation.

J. M. Norton

In his new position, he will direct all the division's technical efforts at laboratories in Westchester, N. Y., and San Jose, Calif.

IBM has assigned the ASD Division the mission of identifying new business opportunities, and the division's engineers and scientists are solving technological and computer system problems identified by market studies.

Mr. Norton has been with IBM since 1953. He has held important assignments in advanced systems development and research. From 1957 to 1959 he was resident manager of the IBM Research Laboratory, Ossining, N. Y. His previous assignment with ASDD before his latest promotion was manager of systems engineering.

A graduate of the University of Michigan, Ann Arbor, he received bachelor's degrees in electrical engineering and mathematics and the master's degree in electrical engineering. He holds several patent applications in telemetering and automation and is a member of Eta Kappa Nu.

❖

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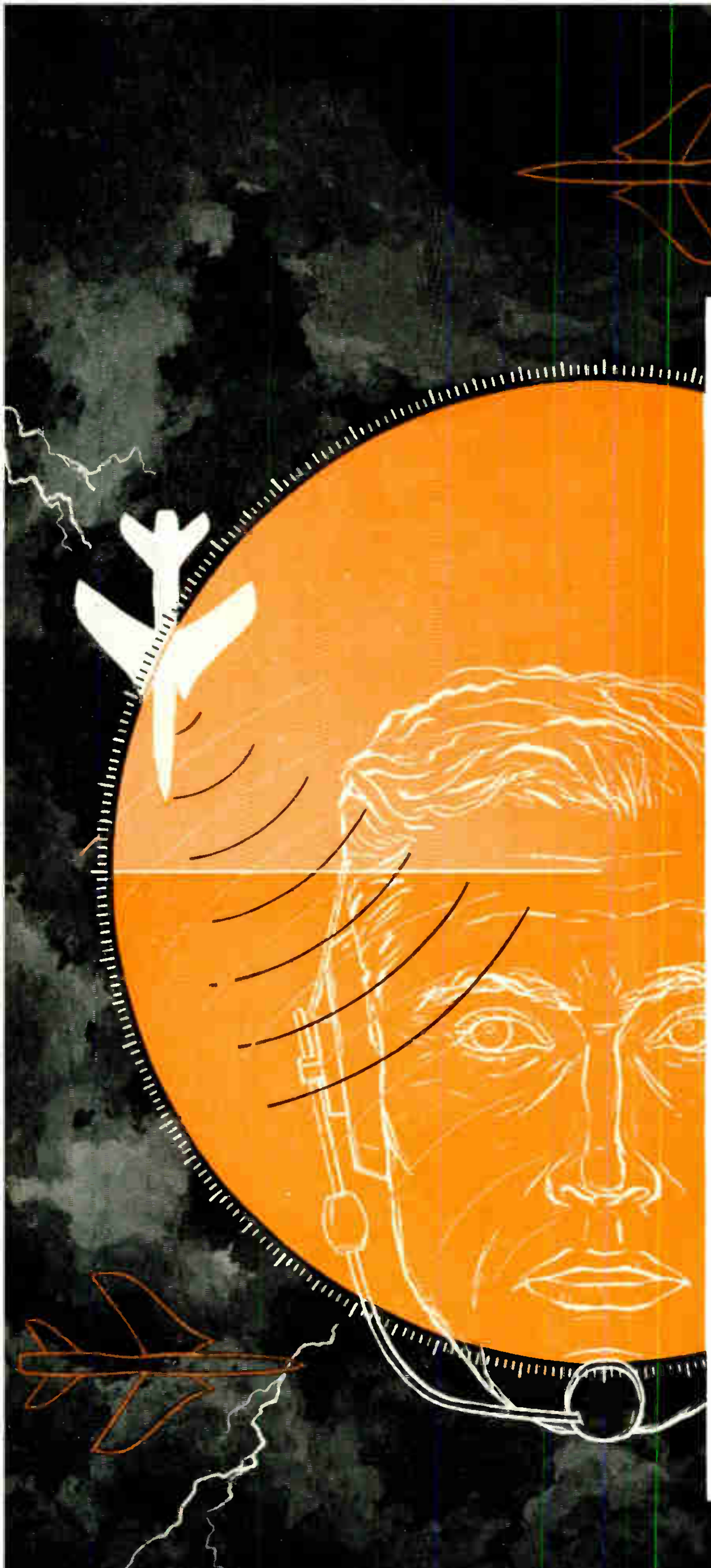

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is experienced... in Ground Controlled Approach Systems

LFE is prime contractor to the U.S.A.F. for an advanced aircraft landing-assistance system. Designated as AN/TPN-12, this system was developed for tactical missions anywhere in the world.

Lightweight . . . portable by truck, ship or helicopter . . . the AN/TPN-12 can be set up and operational within 2½ hours. It has been reliability-proved under the most severe environmental conditions in several recent Arctic military maneuvers. Sweden has purchased this system for installation in airports used jointly by the Royal Air Force and commercial airlines.

LFE's capabilities in GCA have advanced during the past 15 years . . . from the development of MTI Radar Techniques to the present position as major supplier of tactical GCA equipment to the U.S.A.F. Complete details are available upon request.

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ELECTRONIC DATA PROCESSING • MICROWAVE INSTRUMENTATION • AUTOMATIC CONTROLS • AIR TRAFFIC CONTROL

(Continued from page 36A)

Howard S. Bertan (M'59), currently manager of the High Power Electronics Division of the Narda Microwave Corporation, Mineola, L. I., N. Y., will assume additional responsibilities as group leader for the Electronic Products Division, it was recently announced.

He joined Narda in 1959 as section head of the Modulator Department. He holds the B. S. degree in electrical engineering from Massachusetts Institute of Technology, Cambridge, and has taken graduate work at Columbia University, New York, N. Y. His past experience includes research and development work on microwave components and research in the field of high power electronic equipment such as high voltage power supplies and high power pulse modulators.

Mr. Bertan is a member of the Professional Groups on Microwave Theory and Techniques and Electron Devices.

❖

Donald Christiansen (S'48–A'50–M'55–SM'59) has been named to the new post of manager of publications for the CBS Electronics division of Columbia Broadcasting System, Inc., according to a recent announcement.

D. Christiansen

He was previously manager of information services, a position to which he was appointed in January, 1958. He joined CBS Electronics in 1950 as an electron tube engineer and was subsequently made a senior engineer and an engineering group leader. He has written and edited several articles published in electronics publications.

He was graduated from Cornell University, Ithaca, N. Y., with the degree in electrical engineering. He is a member of Eta Kappa Nu, Mu Sigma Tau, and the Cornell Society of Engineers.

He is an officer of the Merrimack Valley Subsection and a member of the Professional Groups on Engineering Management and Engineering Writing and Speech.

Mr. Christiansen is also a member of the American Institute of Electrical Engineers. In 1959, he was elected to the executive board of the Merrimack Valley Subsection, AIEE, and is currently a representative to the Lynn Section Institute Membership Committee.

❖

National Company, Inc., Malden, Mass., recently announced the appointment of **Samuel J. Davy** (SM'56) as director of engineering.

Before coming to the National Company in 1960, he served as manager of the Base Activation Department for the

(Continued on page 40A)

(Continued from page 38A)

ARMA Division of the American Bosch Arma Corporation, Garden City, N. Y. He was responsible for the installation and testing of inertial guidance systems in Atlas ICBM operational bases throughout the nation. He has also served with the Western Electric Company as product engineer and with the U. S. Navy.

Mr. Davy received the B. S. degree from Lehigh University, Bethlehem, Pa., the M. E. degree from New York University, New York, N. Y., and the S.M. degree from Massachusetts Institute of Technology, Cambridge.

❖

Announcement of the promotion of Senior Engineer **Arthur L. De Bolt** (S'47–A'50–M'58) to the position of Senior Group Engineer of the Electronic Engineering Company of California, Santa Ana, Calif., was made recently.

A. L. De Bolt

He has had ten years experience as an electronics engineer. He was formerly with the Hughes Aircraft Company in Tucson, Ariz., where he was responsible for the development of automatic missile systems tests for the Falcon family of missiles. His first professional work was on the Bomarc ground control system at Willow Run Research Center at the University of Michigan, Ann Arbor.

He is a graduate of the Georgia Institute of Technology, Atlanta, and received the B.S.E.E. degree in 1952.

❖

Dr. J. Howard Dellinger (F'23)(L), one of the five Honorary Presidents of the International Scientific Radio Union (URSI), added another honor to an already long list at the 1960 Fall Meeting of URSI, held in Boulder, Colo., December 12–14, 1960. Dr. J. P. Hagen, vice president of the USA National Committee of URSI, presented Dr. Dellinger with a citation for long and meritorious service to the organization since its founding in 1913. A banquet audience of nearly 300 applauded the award.

J. H Dellinger

America's "grand old man of radio" is the senior member in point of service of the USA National Committee. During the last three decades, he has been the strong moving force in building the organization

(Continued on page 42A)

The Highest Sensitivity
and the lowest noise.....

Advanced electronic research by Hitachi technicians has now resulted in the development of a superb frame grid type twin triode (6R-HH8) with excellent high gain and low noise characteristics. As a component of the tuner, the 6R-HH8 ensures an excellent picture with a remarkable degree of definition.

fig. 1 Gain characteristics

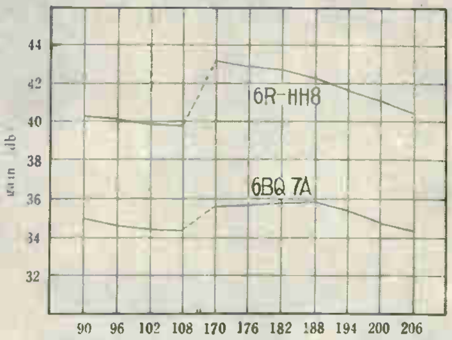
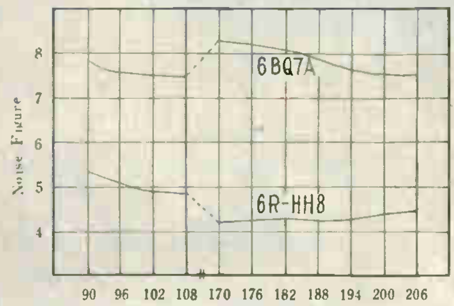
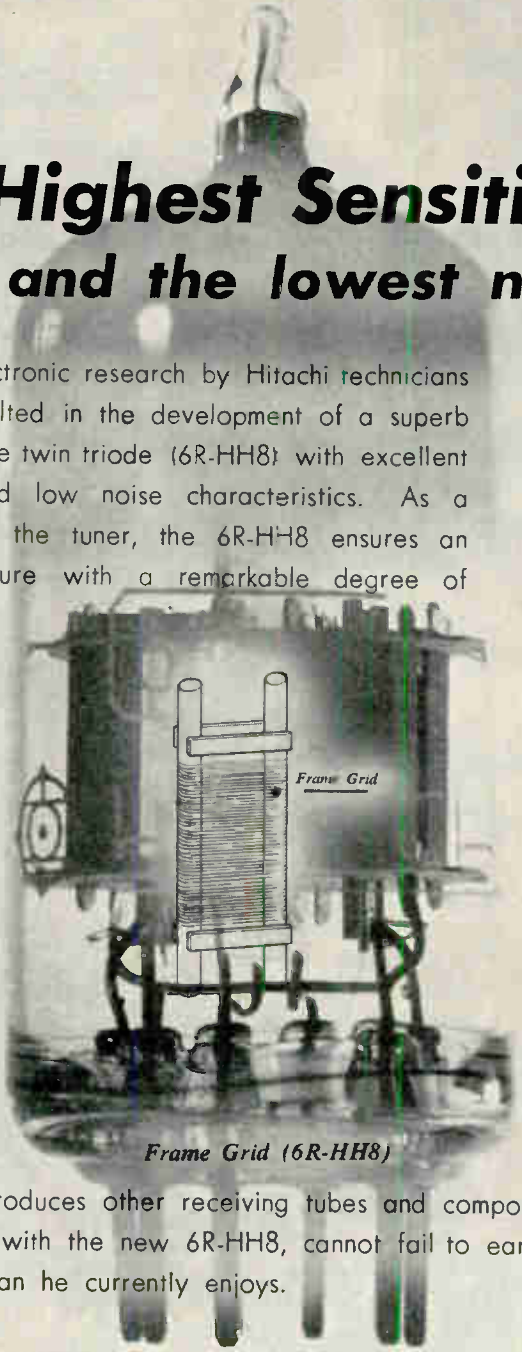


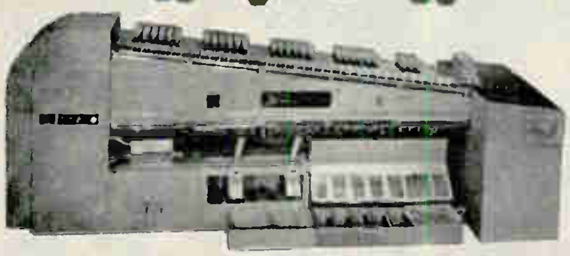
fig. 2 Noise characteristics





Frame Grid (6R-HH8)

Hitachi also produces other receiving tubes and components for television which, when used together with the new 6R-HH8, cannot fail to earn any maker a market reputation even better than he currently enjoys.



Automatic tube testing equipment

Hitachi, Ltd.
Tokyo Japan
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IRE People

(Continued from page 40A)

to its present strength, and in keeping both the national and international organizations alive during World War II.

When radio was in its infancy, he introduced and organized radio research at the National Bureau of Standards in Washington, D. C. He headed the first radio section of the Bureau beginning as early as 1921. As radio gained importance in the work of NBS, his responsibilities increased until, in 1946, he was named first chief of the Central Radio Propagation Laboratory which has now grown to five divisions located at the Boulder Laboratories of NBS in Boulder, Colo.

For a long period, he directed NBS work in the wide field of radio propagation research and also directed and contributed personally to work on radio navigation, particularly aircraft navigation.

Dr. Dellinger's own research in this field culminated in the 4-course radio beacon system, which, until the advent of modern navigational aids, was widely used across the United States for safe navigation of military and commercial aircraft.

His background and status in radio have made him a leader in national and international planning of scientific radio work. His efforts played a significant part in the creation and development of international cooperation in radio research through the URSI, which has now spread to more than 30 countries.

❖

H. J. Elias (M'59) has been named manager of signal diode manufacturing in General Electric's newly-formed Signal Diode Project in its Semiconductor Products Department.

H. J. ELIAS

Immediately prior to his promotion, he was manager of quality control for the Department's Syracuse Plant. He will also continue in that post until a successor has been appointed.

He will be responsible for all phases of the project's manufacturing operations including shop operations, quality control, manufacturing engineering and materials selection and procurement.

He is a native of Rochester, N. Y. He was graduated from the University of Illinois, Urbana, in 1947 with the bachelor of science degree in electrical engineering.

He joined the General Electric Company at Schenectady, N. Y., in 1947 and had various assignments in the company's test engineering program, becoming a process engineer at the Cathode Ray Tube Department's Buffalo plant in 1949. In 1952, he became supervisor of process

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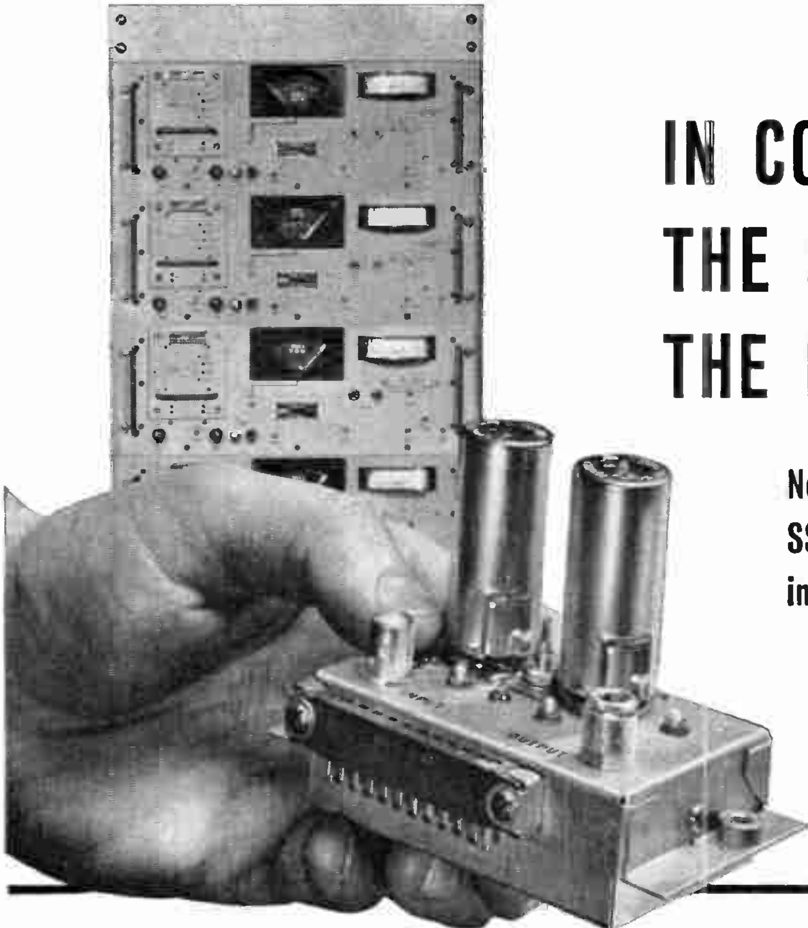

IN COMMUNICATIONS... THE SIMPLER THE BETTER

New Hallicrafters all-modular SSB strip receiver cuts costs, increases reliability.

- **50% less maintenance**
- **Far greater stability and reliability**
- **Down time almost entirely eliminated**
- **Lower initial cost**

Hallicrafters' new SX-116 SSB Receiver is the essence of simplicity—key to reliability in the Hallicrafters Series 116 communication system.

The SX-116 is entirely modular in construction, *virtually eliminating "down time" and cutting maintenance cost by over 50%.* The unit is quickly and easily adaptable to existing systems, entirely compatible with future requirements.

It is extremely stable—1 part in 10^6 per month (standard) or 1 part in 10^8 per month (special) . . . it permits, for the first time, continuous, unattended operation with *maximum reliability.*

The SX-116 weighs in at just 36 lbs.—equally practical for fixed, mobile, air or seaborne installations. *And its initial cost is very substantially lower.*

Finding a better and simpler solution to complex communications problems has been a Hallicrafters habit for over a quarter-century.

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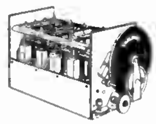
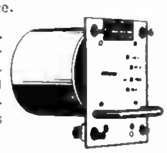
Military Electronics Division, Chicago 24, Illinois

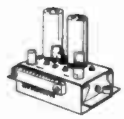
100% modular construction—only seven basic components



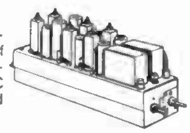
RF Module. Image and IF rejection maintained at better than 70 db. Single tuned circuit between antenna and RF grid. Four-channel, continuous tuning—2.0 mc. to 30 mc. range.



HF Crystal Oscillator Module. Stability: 1 in 10^6 per month. Capacity is four crystals; designed for HC-6/U metal or glass crystal holders. Oven temp. varies less than ±0.01°C.

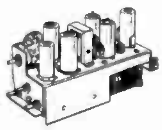


Injection Amplifier Module. AGC-controlled wide-band video amplifier provides constant injection level.

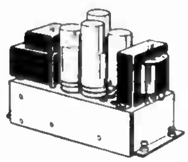


IF Module. Allows simultaneous reception of upper and lower sideband, independent AGC control of upper and lower sideband.

BFO Module. Operates at 1650.000 kc. Oscillator frequency stabilized in separate oven. Plate and filament voltages regulated.



Audio Amplifier Module. Features duel, independent 100-milliwatt line amplifiers. Hum level is 80 db. below 100 milliwatts. Harmonic dist. less than 0.5%.



Power Supply Module. Separate transformers provided for regulated and non-regulated voltages. Local oscillator and BFO filament supply are regulated for ±10% line voltage variations!

Write for complete specifications on the SX-116.

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LOOKING FOR A CHALLENGING OPPORTUNITY? We have openings for communications systems and equipment design engineers at all levels. Please write to William F. Frankart, Director of Engineering.

IRE People

(Continued from page 42A)

planning and engineering, and in 1954, supervisor of quality control and engineering. In November, 1958, he joined the Semiconductor Products Department as manager of quality control, the post he held immediately prior to his latest promotion.

Mr. Elias is a member of the American Society for Quality Control.

❖

Promotion of **Clarence E. Elkins** (S'58–M'59) to senior project engineer for Lynch Communication Systems Inc., San Francisco, Calif., was recently announced.

Mr. Elkins joined Lynch after receiving the B.S.E.E. degree in 1958 from the University of California, Berkeley. At Lynch, he participates in the engineering development of telephone, data transmission and control equipment which the company designs and manufactures.

❖

Narda Microwave Corporation of Mineola, L. I., N. Y., has announced that **Donald S. Elkort** ('59) has been named group leader for ferrite devices and will be in charge of designing and developing ferrite devices such as circulators, isolators, modulators, and phase shifters.

He joined Narda early in 1960 as a microwave engineer, and was formerly an Associate Project Engineer with the Microwave Electronics Division of Sperry Gyroscope Company. He has extensive experience in research and development of microwave ferrite components for high power radar and countermeasures systems, both ground and airborne. At Sperry, he was responsible for the design and development of microwave ferrite devices such as isolators, attenuators, and phase shifters. His previous experience also includes the design of stripline and coaxial line components for use in automatic test equipment and the design and development of components for use in extremely high power, high resolution pulse compression radar systems.

Mr. Elkort received the B. A. degree in physics from New York University, New York, N. Y., and is a member of the Professional Group on Microwave Theory and Techniques, and of the American Physical Society.

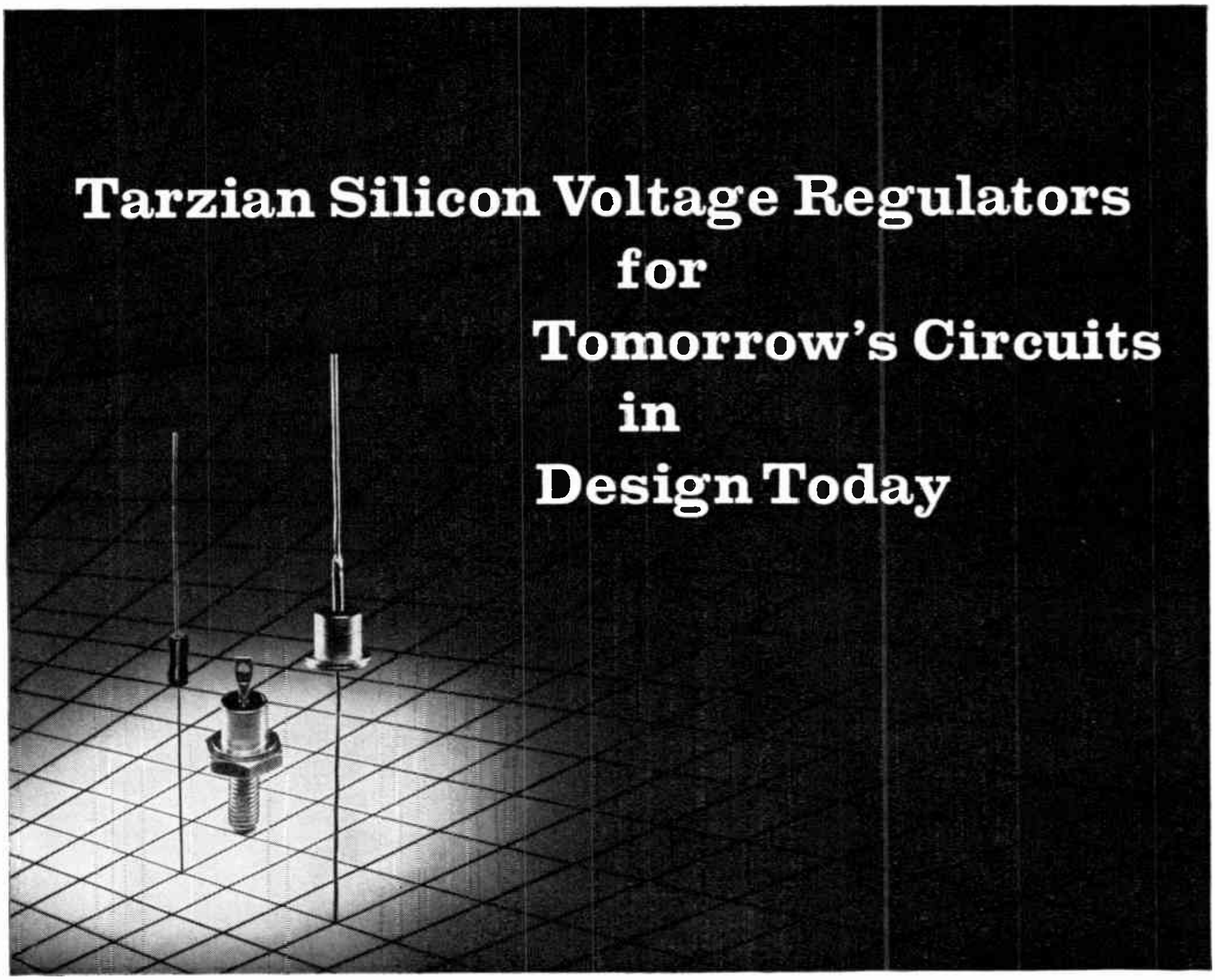
❖

National Company, Inc., Malden, Mass., has recently announced the appointment of **Saul Fast** (A'47–M'55–SM'55) as technical assistant to the president.

He joined the National Company in 1955 as chief engineer in the Communications Systems Department, where he helped to develop several types of receivers and transmitter systems. In 1958, he was promoted to director of the Technical Liaison Division

(Continued on page 46A)

Tarzian Silicon Voltage Regulators for Tomorrow's Circuits in Design Today



TARZIAN SILICON VOLTAGE REGULATORS—in ¼-watt, 1-watt and 10-watt classifications—will serve you well as DC power regulators, AC clippers and limiters, and as protective devices in a wide variety of component protection circuits.

Their small size, inherent ruggedness, and physical simplicity are distinct improvements over other types of regulators. Sharp and instantaneous breakdown (avalanche) characteristics and instantaneous recovery are invitations to inventiveness in circuit design.

Tarzian silicon regulators, like Tarzian silicon rectifiers, are made with a unique blend of care and ingenuity that delivers excellent quality at a realistic price. They are available now from factory or warehouse stocks in sample or volume production quantities. Application engineering service is offered without obligation.

¼ Watt Voltage Regulators Specifications 25°C					1 Watt Voltage Regulators Specifications 25°C.				10 Watt Voltage Regulators Specifications 25°C.				
Tarzian Type	Zener Volt (V)	Test Cur (Ma)	Dyn Imp (Ohms)	Jedec Type	Tarzian Type	Zener Volt (V)	Test Cur (Ma)	Dyn Imp (Ohms)	Tarzian Type	Zener Volt (V)	Test Cur (Ma)	Dyn Imp (Ohms)	Jedec Type
.25T5.6	5.6	25	3.6	1N708	1T5.6	5.6	100	1.2	10T5.6	5.6	1000	1	1N1803
.25T6.2	6.2	25	4.1	1N709	1T6.2	6.2	100	1.5	10T6.2	6.2	1000	1	1N1804
.25T6.8	6.8	25	4.7	1N710	1T6.8	6.8	100	1.7	10T6.8	6.8	1000	1	1N1805
.25T7.5	7.5	25	5.3	1N711	1T7.5	7.5	100	2.1	10T7.5	7.5	1000	1	1N1806
.25T8.2	8.2	25	6.0	1N712	1T8.2	8.2	100	2.4	10T8.2	8.2	1000	1	1N1807
.25T9.1	9.1	12	7.0	1N713	1T9.1	9.1	50	3.0	10T9.1	9.1	500	1	1N1808
.25T10	10	12	8.5	1N714	1T10	10	50	3.5	10T10	10	500	2	1N1351
.25T11	11	12	9.0	1N715	1T11	11	50	4.2	10T11	11	500	2	1N1352
.25T12	12	12	10	1N716	1T12	12	50	5.0	10T12	12	500	2	1N1353
.25T13	13	12	11	1N717	1T13	13	50	5.8	10T13	13	500	2	1N1354
.25T15	15	12	13	1N718	1T15	15	50	7.6	10T15	15	500	2	1N1355
.25T16	16	12	15	1N719	1T16	16	50	8.8	10T16	16	500	3	1N1356
.25T18	18	12	17	1N720	1T18	18	50	11	10T18	18	150	3	1N1357
.25T20	20	4	20	1N721	1T20	20	15	13	10T20	20	150	3	1N1358
.25T22	22	4	24	1N722	1T22	22	15	16	10T22	22	150	3	1N1359
.25T24	24	4	28	1N723	1T24	24	15	18	10T24	24	150	3	1N1360
.25T27	27	4	35	1N724	1T27	27	15	23	10T27	27	150	3	1N1361
.25T30	30	4	42	1N725	1T30	30	15	28	10T30	30	150	4	1N1362
.25T33	33	4	50	1N726	1T33	33	15	33	10T33	33	150	4	1N1363
.25T36	36	4	60	1N727	1T36	36	15	39	10T36	36	150	5	1N1364
.25T39	39	4	70	1N728	1T39	39	15	45	10T39	39	150	5	1N1365
.25T43	43	4	80	1N729	1T43	43	15	54	10T43	43	150	6	1N1366
.25T47	47	4	95	1N730	1T47	47	15	64	10T47	47	150	7	1N1367
.25T51	51	4	115	1N731	1T51	51	15	74	10T51	51	150	8	1N1368
.25T56	56	4	140	1N732	1T56	56	15	88	10T56	56	150	9	1N1369
.25T62	62	2	170	1N733	1T62	62	5	105	10T62	62	50	12	1N1370
.25T68	68	2	200	1N734	1T68	68	5	125	10T68	68	50	14	1N1371
.25T75	75	2	240	1N735	1T75	75	5	150	10T75	75	50	20	1N1372
.25T82	82	2	280	1N736	1T82	82	5	175	10T82	82	50	22	1N1373
.25T91	91	1	340	1N737	1T91	91	5	230	10T91	91	50	35	1N1374
.25T100	100	1	400	1N738	1T100	100	5	260	10T100	100	50	40	1N1375

NOTES: Standard tolerance is ± 10% however, closer or wider tolerances are available on request.
Also available on request
(a) Special voltage ratings.
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Write for your free copy of the new 6-page Tarzian voltage regulator catalog. It contains helpful design and test information and application notes.

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IRE People

(Continued from page 44A)

In announcing this appointment, J. H. Quick, president of National, added that the efforts of his newly appointed assistant had indicated a broad knowledge of engineering and had contributed to the rapid growth of the company.

Before joining National, Mr. Fast worked for the W. L. Maxson Corporation and the U. S. Signal Corps. He received the B.S. degree from the City College of New York and the master's degree in electronic engineering from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. He is the author of several technical papers.

❖

Manfred Gale (A'41–M'57) was cited recently for the second consecutive year for his fine work at the U. S. Army Engineer Research and Development Laboratories, Fort Belvoir, Va. He received an "Outstanding" rating for his work in the Mine Detection Branch of the Laboratories. Last year, he was the recipient of a "Sustained Superior Performance" award.

M. GALE

A veteran of World War II, he attended the University of Virginia, Charlottesville, and received the bachelor of science degree in electrical engineering in 1949. In October, 1950, he accepted a position at the Fort Belvoir Laboratories, which is the principal field agency of the Corps of Engineers for the research and development of new materiel, methods, and techniques required for military operation.

❖

Donald E. Garr (SM'55) has been named corporate director of engineering for Raytheon Company, Waltham, Mass., it was recently announced. He has resigned as manager of engineering operations for General Electric Company's armament and control section.

D. E. GARR

In his new post, he will report to the vice president—engineering and research.

Associated with General Electric since 1936, his engineering background includes management of development programs in industrial electronic systems, military fire control and aircraft electronic systems, data handling systems, human engineering and computer development.

A 1936 graduate of Kansas State College, Manhattan, he holds the B.S.E.E. degree and is a licensed professional engineer in the state of New York. He is a Fellow

(Continued on page 48A)

Professional Group on Ultrasonics Engineering

One of the greatest assets of the IRE Professional Group plan is its flexibility in being able to serve equally well the many branches of the electronics field, regardless of how new, small, or specialized they may be. This is convincingly demonstrated in the case of the Professional Group on Ultrasonics Engineering.

Although a relatively young field, ultrasonics already embraces many diverse fields. In the laboratory ultrasonic techniques are used for studying the properties of gases, liquids, and solids. Marine applications include ultrasonic depth indicators and underwater object locators. In the medical field ultrasonic diathermy instruments, tumor locators, dental caries locators, and even a device for drilling teeth are being investigated and developed. In industry ultrasonics is finding application in nondestructive testing of materials, acceleration of chemical reactions, emulsification, coagulation, and sterilization. Ultrasonics delay lines and electromechanical filters are being used in radio, radar and digital computers. There may one day be ultrasonic washing machines in the home, as there now are in the laboratory.

How then is one to keep abreast of these tremendous strides? Only by receiving an authoritative technical publication devoted exclusively to this subject; only by attending meetings at which recent advances are discussed; only by exchanging ideas with other workers in the field.

It was with this purpose that the Professional Group on Ultrasonics Engineering was formed seven years ago. Thanks to the *Transactions* and the Group sponsored meetings, the workers in this field are now able to keep fully informed of the diverse developments in this rapidly growing field.

Ernst Weber

Chairman, Professional Groups Committee

IRE People

(Continued from page 46A)

and past officer of the American Institute of Electrical Engineers and a member of both the American Society of Mechanical Engineers and the New York State Society of Professional Engineers.

Mr. Garr holds 19 patents in the fields of amplidynes and magnetic amplifiers.

❖

Dr. Ernst A. Guillemin (A'41–SM'48–F'49), Webster Professor of Electrical Engineering at the Massachusetts Institute of Technology, Cambridge, who has been chosen by the IRE to receive its highest technical award, has been named vice president and director of research of Burnell & Co., Inc. The firm, located in Pelham Manor, N. Y., is a producer of electronic filter networks.

E. A. Guillemin

Recognized as one of the world's leading authorities and consultants on communications and electronic network analysis, Dr. Guillemin was recently announced as recipient of the IRE 1961 Medal of Honor. He holds a Presidential Certificate of Merit for outstanding scientific contributions to the country during World War II. A member of the faculty of M.I.T. since 1928, he recently was advanced from professor of electrical communications to the coveted Webster chair.

After receiving the B.S. degree from the University of Wisconsin, Madison, in 1922, Dr. Guillemin took advanced studies at M.I.T. and in 1924 received the master's degree in electrical engineering. He was awarded the 1924 Saltonstall Traveling Fellowship for study at the University of Munich (Germany), where he received his doctorate in 1926. He returned to M.I.T. as an instructor in 1926, became an assistant professor in 1928, associate professor in 1936, and a full professor in 1944.

He took over administrative responsibilities of the Communications Section of the Department of Electrical Engineering at M.I.T. in 1941. In this capacity and as a consultant to the Microwave Committee of the National Defense Research Committee, he worked with the Radiation Laboratory at M.I.T. on various electrical problems. One of the results was development of a network for production of radar pulses.

Dr. Guillemin is the author of several volumes concerned with communications networks. He is a Fellow of the American Institute of Electrical Engineers and of the American Academy of Arts and Sciences, and he is a member of the American Society for Engineering Education.

❖

John L. Heins (A'32–VA'39–SM'59), well known in government and defense circles, has joined Servo Corporation of America, Hicksville, L. I., N. Y., as Director, Defense Systems.

J. L. Heins

Prior to joining Servo, he was Vice President, engineering, with G. B. Electronics Corporation, a subsidiary of General Bronze Corporation. He has also held engineering management positions with Republic Aviation Corporation, Burroughs Corporation, International Telephone and Telegraph Laboratories, the Sperry Gyroscope Company, and the U. S. Government.

Mr. Heins has worked with microwave, radio direction finding, tracking systems, radar, aircraft instrument landing systems, computers, communications, and other electronic advances. He is credited with a number of inventions and significant developments in these fields. He is considered an expert in ECM and ECCM.

❖

Blain A. Holmlund (M'60), B.E., lecturer in electrical engineering at the University of Saskatchewan, Saskatoon, Sask., Canada, has been appointed to full membership on the Faculty of the College of Medicine as lecturer in biomedical engineering. He will retain his place on the faculty of the College of Engineering, thus establishing a positive liaison link between the two disciplines.

B. A. Holmlund

He has had an active part in much of the medical research carried out by the joint efforts of the College of Medicine and the College of Engineering, and his keen interest and ready understanding of medical terminology made him the logical choice for the appointment.

In addition to lecturing and actively participating in research, he will advise and assist technicians in the various laboratories and departments, design and supervise construction and installation on such apparatus and control systems as may become necessary in research or to improve diagnostic and therapeutic techniques. He will also be responsible for repair, calibration and modification of electronic equipment presently in use.

❖

Armig G. Kandoian (S'35–A'36–SM'44–F'51), vice president and general manager of ITT Laboratories, Nutley, N. J., has been appointed an advisor in electrical engineering to the board of trustees, Newark College of Engineering, Newark, N. J.

An authority on worldwide communications, he is one of four new advisors who will each serve for three years. In all, 24

(Continued on page 50A)

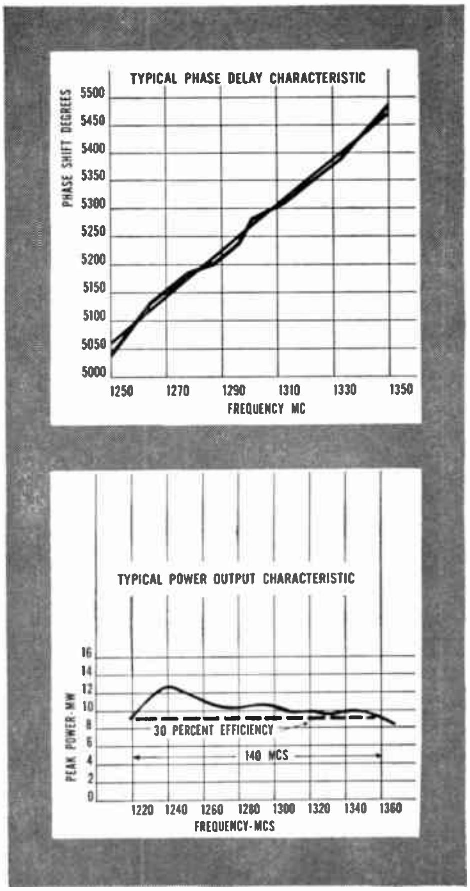
New Broadband Klystrons

140 MEGACYCLES - (1db) BANDWIDTH AT L-BAND
10 MEGAWATTS - PEAK POWER OUTPUT

New additions to the Litton Industries Broadband Klystron family extend broadband performance to even higher power levels as shown in the typical performance curves to the right. These tubes, like all those produced by Litton Industries, are conservatively designed and rated; and rigorously processed to provide many thousands of hours of reliable operation. Using Litton developed broadbanding techniques, it is now possible to achieve wide bandwith, high peak and average rf power output and linear phase shift versus frequency characteristics simultaneously. This latter feature enables the radar equipment designer to utilize pulse compression techniques to attain improved system performance.

Litton Klystrons providing these outstanding performance characteristics can be supplied in both the L and S-bands at peak rf power levels ranging from 2 to 20 megawatts. Typical of the performance obtained with Litton Klystrons is that of the L-3035, a 2.2 megawatt L-band Klystron, whose average operating life in field service is approaching 3,000 hours. Some of these tubes are continuing to provide excellent service after having operated for more than *17,000* hours.

Should you require high power broadband amplifier tubes to satisfy your system requirements, please write to us at Litton Industries, Electron Tube Division, 960 Industrial Road, San Carlos, California. Our telephone number is LYtell 1-8411.



LITTON INDUSTRIES
Electron Tube Division
MICROWAVE TUBES AND DISPLAY DEVICES

"Capability that can change your planning"

IRE People

(*Continued from page 48A*)

advisors in electrical, mechanical, civil and chemical engineering inform the NCE board and departments on current industrial developments in their fields.

Formerly Vice President and Director of Communications Systems at ITT, he has been with the division since 1935 when he joined it as a student engineer. His work has covered research and development in world-wide communications systems, navigation aids, ultra high frequency techniques, radar and antennas. He holds 44 patents in these phases of engineering.

Mr. Kandoian is also chairman of the editorial board of "Reference Data for Radio Engineers," a source book of basic telecommunications material published by ITT. Responsible for organizing the book, he is also a frequent contributor and has published numerous articles in other journals and trade magazines.

Before joining ITT, he received the B.S. and M.S. degrees at Harvard University, Cambridge, Mass., in 1934 and 1935, respectively. Both are in electrical communication engineering. In 1943, he received the honorable mention award from Eta Kappa Nu, national honorary society of electrical engineers, recognizing him as an outstanding engineer.

Mr. Kandoian is a member of Tau Beta Pi, the national engineering honor society, Harvard Engineering Society, the American Ordnance Association, and the Columbia Executive Association.

❖

Stanley P. Lapin (S'45–A'48–M'54–SM'55) was recently appointed Director of the Industrial Products Division of Adler Electronics, Inc., New Rochelle, N. Y. The newly-formed division will be responsible for microwave, translator and transmitter systems for the broadcast, community, and educational TV fields.

S. P. Lapin

Prior to joining Adler in early 1960 as Assistant to the President, he was Operations Manager of Motorola's Microwave Department in Chicago.

Mr. Lapin received the B.S.E.E. degree from the Illinois Institute of Technology, Chicago; the M.S.E.E. degree from the Massachusetts Institute of Technology, Cambridge; and the M.B.A. degree from Northwestern University, Evanston, Ill. He is a member of the Electronic Industries Association, the American Marketing Association, and the AIEE.

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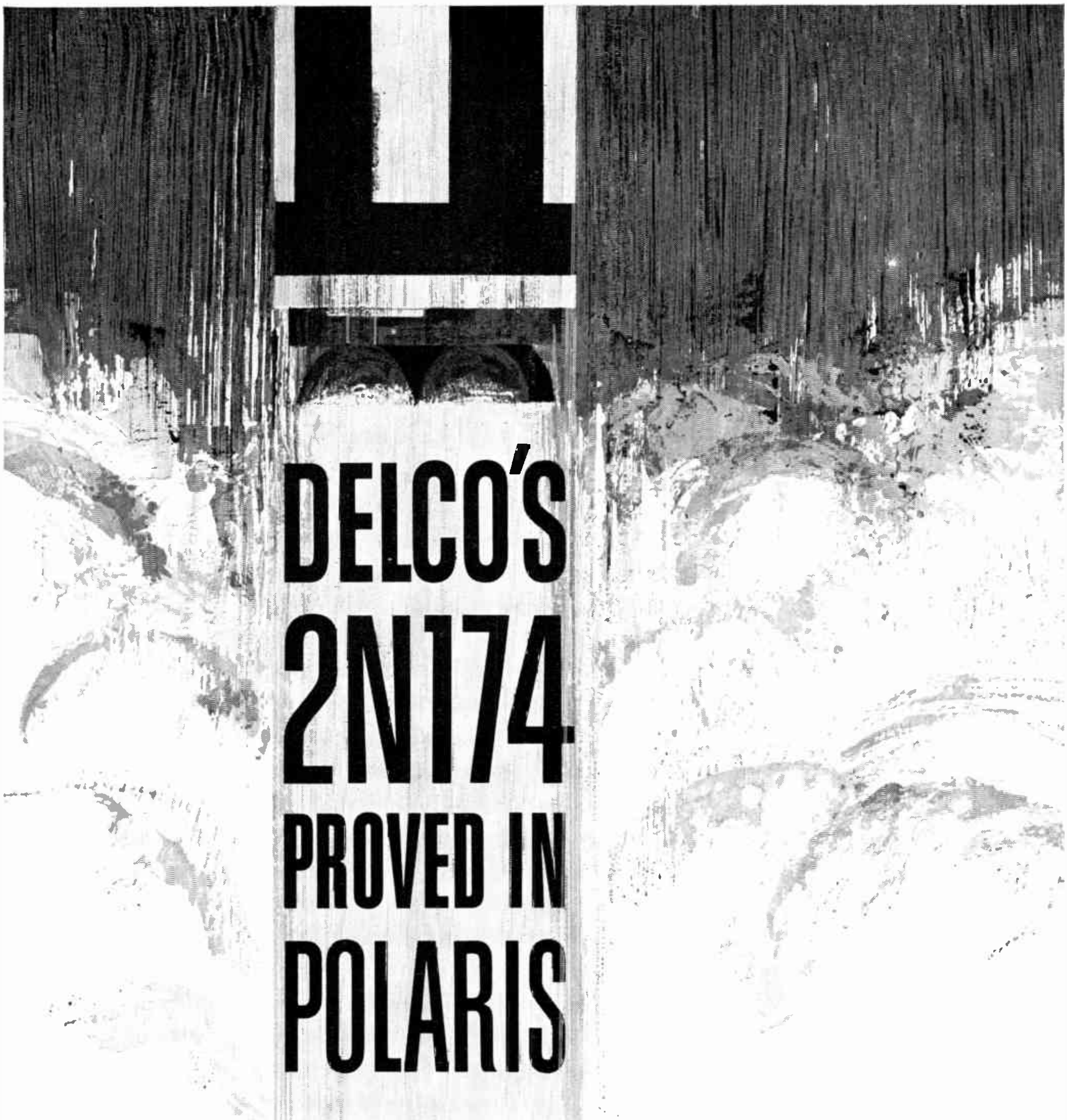

. . . and Minuteman and Talos and Atlas and Jupiter and Thor and Titan and Bomarc and Zeus and Pershing and hundreds of other military and industrial applications.

For Delco Radio's highly versatile family of 2N174 power transistors meet or exceed the most rigid electrical and extreme environmental requirements.

Over the past five years since Delco first designed its 2N174, no transistor has undergone a more intensive testing program both in the laboratory and in use, in applications from mockups for commercial use to missiles for the military. And today, as always, no Delco 2N174 leaves our laboratories without passing at least a dozen electrical tests and as many environmental tests before and after aging.

This 200 per cent testing, combined with five years of refinements in the manufacturing process, enables us to mass produce these highly reliable PNP germanium transistors with consistent uniformity. And we can supply them to you quickly in any quantity at a low price.

For complete information or applications assistance on the Military and Industrial 2N174's or other application-proved Delco transistors, just write or call our nearest sales office.

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DELCO RADIO
DEPENDABILITY
RELIABILITY

Division of General Motors · Kokomo, Indiana

IRE People

(Continued from page 50A)

Dr. Nick Holonyak, Jr. (S'51–A'55–M'59) was recently appointed head of the Structure Studies Unit of the General Electric Advanced Semiconductor Laboratory. This unit is one of five new units into which the Laboratory has been reorganized in order to meet increasing demands of semiconductor technology.

N. Holonyak, Jr.

Dr. Holonyak, who joined the ASL in 1956, was engaged in semiconductor device physics prior to his new assignment.

❖

Dr. Wilbur A. Lazier (A'54–M'59) was recently appointed Senior Vice President, Technical Director, of the Sprague Electric Company, North Adams, Mass.

He is a graduate of the University of Illinois, Urbana, and also holds the Ph.D. degree in chemistry from the University of Wisconsin, Madison. He joined Sprague Electric in 1953 as Vice President and Technical Director. Prior to joining Sprague, he served as Director of Chemical Research and Development for the Charles Pfizer Company, and as a member of the firm's Board of Directors. He also had been associated for many years with E. I. du Pont de Nemours Company and with the Southern Research Institute.

Dr. Lazier is currently a member of the Research Committee, National Association of Manufacturers.

❖

FXR, Inc., of Woodside, N. Y. has announced the appointment of **Stanley Lehr** (A'56–M'59) as an engineering section head. He will work on an expanded program of microwave component and system development.

S. Lehr

He formerly designed microwave components and special test equipment for Doppler Navigators, and earlier was with General Precision Laboratories and the Microwave Research Institute.

A cum laude graduate of the City College of New York, Mr. Lehr received the master's degree in electrical engineering from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. He is a member of Eta Kappa Nu and Sigma Xi engineering societies.

❖

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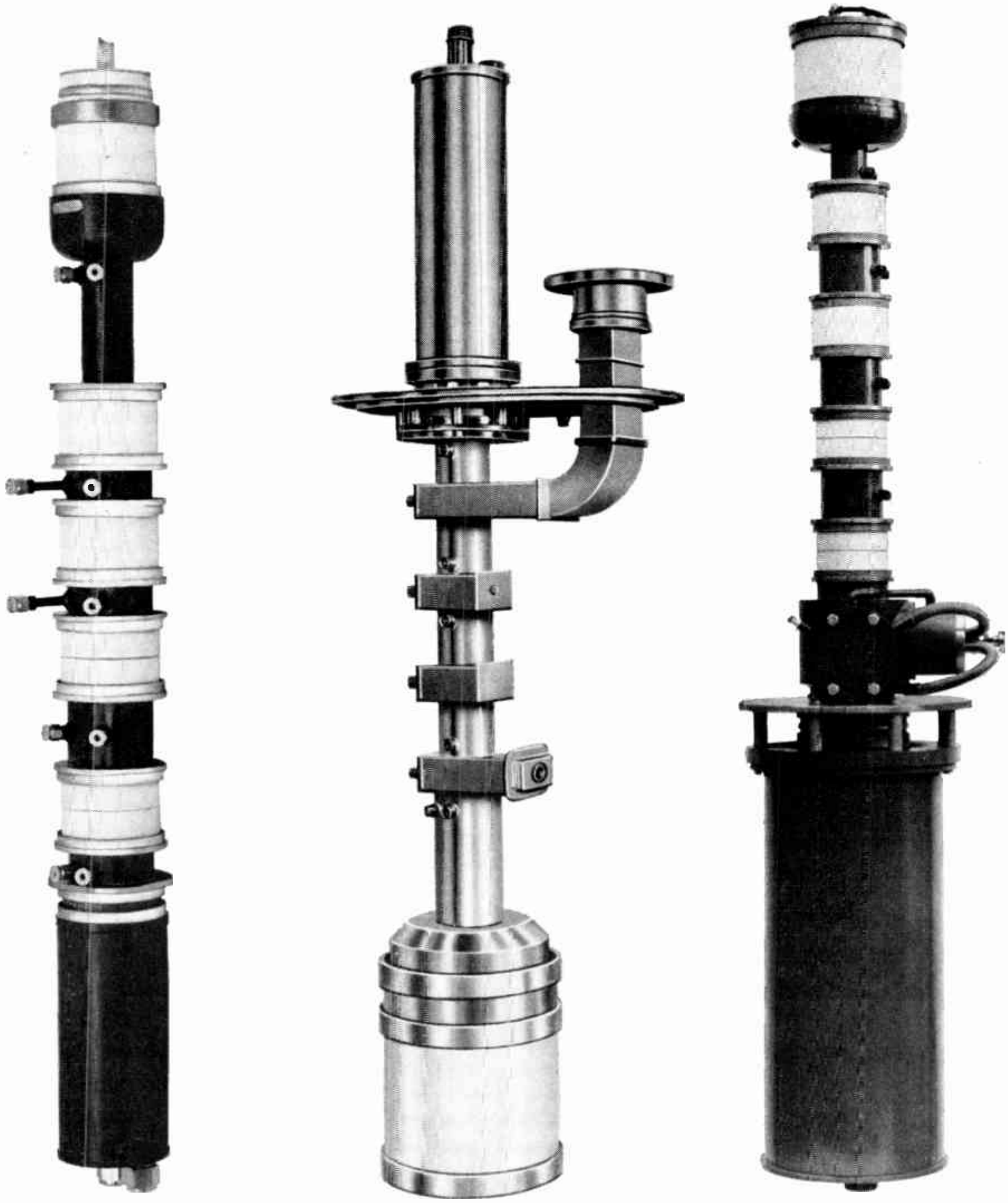

IRE People

(Continued from page 52A)

Dr. I. Arnold Lesk (S'47–A'52–M'56–SM'59) has been named to manage the Microdevices Studies Unit of the General Electric Advanced Semiconductor Laboratory. The Microdevices Studies Unit is one of the five principal research areas into which the Laboratory has been reorganized in order to meet increasing demands of semiconductor technology.

I. A. Lesk

Mr. Lesk joined the G. E. Electronics Laboratory in 1951 and, in 1957, joined the ASL. Prior to assuming his new assignment, he was a consulting scientist for semiconductor devices and techniques.

❖

Milton Magid (S'48–A'49–M'55–SM'57) has joined Microwave Dynamics Corp., a subsidiary of Talley Industries, Inc., as senior staff engineer, it was recently announced.

M. Magid

He will be responsible for planning and execution of a program for development of a proprietory line of microwave components and instruments. Conception and development of these products will be aimed at advancing the state of the microwave art.

He was formerly head of Microwave Product Engineering at FXR, Inc., where he was responsible for microwave component and instrument quality by establishing improved measurement techniques and evaluating existing product performance. In addition, he contributed to new product development and initiated a standards laboratory program.

Prior to this, he was associated with Hughes Aircraft Company for ten years. As supervisor of the Microwave Section of the Primary Standards Laboratory, his department had the prime responsibility of certifying accuracy of precision components and equipment by developing advanced techniques and employing instruments known to conform to primary standards maintained by the National Bureau of Standards.

His group also engaged in applied research leading to improvement of measurement techniques and the establishment of new techniques where none were available, thus advancing the state of the art.

Other original areas of work also included microwave phase shift measurements, dielectric constant measurements, impedance measurements and directional coupler directivity measurements. Some special mathematical functions required for this effort were computed and tabulated after being programmed on IBM 704 and 709 computers utilizing the FORTRAN system.

In addition, need for special instrumentation not commercially available led to development of an improved klystron power supply and modulator, a frequency stabilized klystron source, a linear narrow band pre-amplifier for bolometric attenuation measurements, a high-resolution expanded scale voltmeter, and a driver for a ferrite-type modulator

Work of his group led to numerous improvements in microwave components and instruments available to the industry as a whole.

Mr. Magid, who was chairman of the technical session on microwave standards and calibrations at the June, 1960 Conference on Electrical Standards and Measurements held at Boulder, Colo., has presented or published a number of technical papers including: "Recent Developments in Precision Measurements Techniques in the Microwave Region," "Broad Band Frequency Stabilization of a Reflex Klystron by Means of an External High-Q Cavity," "Precision Microwave Phase Shift Measurements," and "Highlights of the Microwave Standards and Calibration Session of the recent Conference on Electronic Standards and Measurements."

He is an associate member of RESA, and was elected to Eta Kappa Nu, Tau Beta Pi and Sigma Xi honorary societies while at the University of California at Berkeley.

He received the B.S. degree (with honors) from the University of California at Berkeley and the M.S. degree from UCLA. His thesis earned a University Citation for excellence.

❖

Austen H. Madeson (A'55–M'60) has been elected Vice President-Marketing for Solid State Materials Corporation of East Natick, Mass., it was recently announced. In this newly created position, he has over-all responsibility for sales, advertising, and market research for both the equipment and materials lines of the company.

A. H. Madeson

Prior to joining Solid State Materials Corporation, he was Sales Manager of the Crystal Filter Division of Hermes Electronics Co., a division of Itek Corporation. He has also been a Field Engineer for Hughes Aircraft Co., and a Research Scientist for the United States Navy. Mr. Madeson received the B.S. degree in Physics from the City College of New York and did post-graduate work at the University of Connecticut, Storrs, and at Boston University, Boston, Mass.

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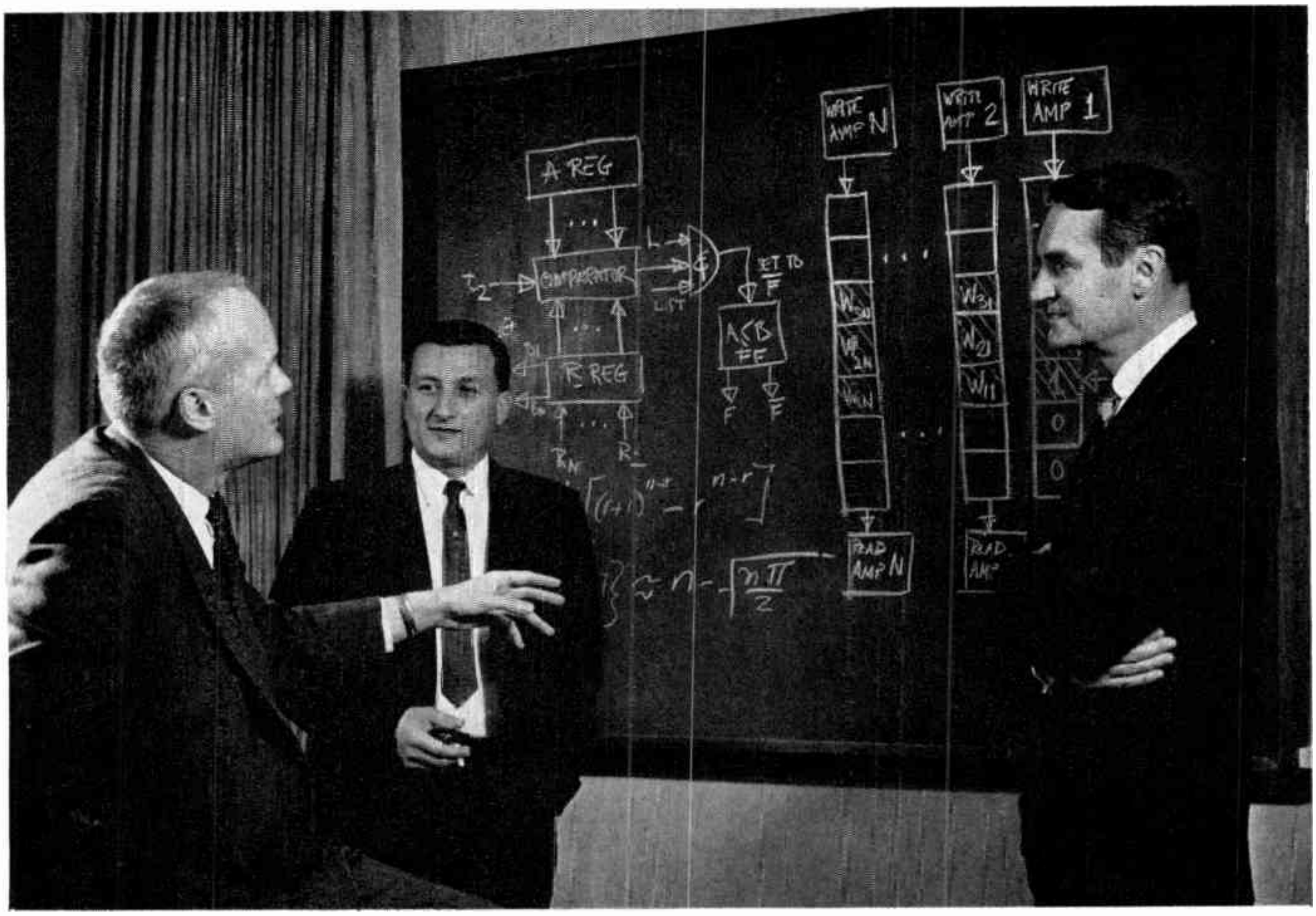

Mr. R. J. Shank, President (right), with Dr. R. E. Fagen (center) and Dr. R. B. Dawson (left) of the Information Sciences Division.

A Report from American Systems Incorporated...

A New Organization for Advanced Systems Technology

American Systems Incorporated was launched a year ago for research and development in the electronic systems field. With an across-the-board interest in systems technology, the Corporation has formed five Divisions:

ELECTROMAGNETIC SYSTEMS

Electromagnetic physics; electronic and mechanical scanning antenna systems; development and manufacturing of special microwave components; design, development, and manufacturing of complete sensor systems.

COMMAND AND CONTROL SYSTEMS

Logic of command and control complexes; systems design and development; data acquisition, processing and display; communications.

COMPONENT DEVELOPMENT

Advanced component technology; materials and processes; computer component development; chemical deposition of magnetic materials on drums, disks, rods, tapes.

INFORMATION SCIENCES

Mathematical and statistical research; computer programming, and development of advanced programming systems; computation services; digital system studies; logical design of military and industrial systems; advanced systems analysis.

RESEARCH LABORATORIES

Solid state physics and systems; thin-film research and subsystems; components for information processing.

We are gratified that the past year has been one of significant growth. Operations were started in a leased 10,000-square-foot building. Recently, we moved into our own 27,000-square-foot plant, on a 13-acre site in Hawthorne, California. This plant, which is the first unit in a long-range building program, has custom technical and scientific facilities, including an ultraclean laboratory for thin-film developments and advanced devices research projects.

We are proud that we have been able to attract an outstanding staff of technical people. We believe that scientists and engineers are our primary resource, and it is to utilize this resource that we have founded this corporation. Our operating concept has been to establish an organization which both sought new ideas and provided the facilities in which the creative mind could also be a builder, seeing his ideas through to a practical product.

Qualified scientists and engineers who are interested and experienced in our fields of activity are encouraged to investigate career opportunities with American Systems.

AMERICAN SYSTEMS INCORPORATED

AMERICAN SYSTEMS *Incorporated*

1625 East 126th Street, Hawthorne, California

IRE People

(Continued from page 54A)

Raymond G. Loughlin (A'49–SM'55) has been appointed manager of the weapons systems division at PRD Electronics, Inc., Brooklyn, N. Y., manufacturer of microwave and electronic instrumentation. PRD is a subsidiary of Harris-Intertype Corp., Cleveland, Ohio.

R. G. LOUGHLIN

He has been employed in the field of microwave and electronic development for nearly 20 years. Prior to joining PRD, he held the position of engineering section head with the Sperry Gyroscope Company, where he was responsible for the development of radar and electronic test equipment for many years. In 1953, he developed a family of field test sets for the Navy which resulted in a new concept of integrating the functions of several test sets within a single package. His recent activities were in connection with the B58 Hustler program; he planned and coordinated the test and evaluation of the Automatic Ground Support Equipment.

Mr. Loughlin received the B.E.E. degree from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. During World War II, he served in the Navy as an electronic specialist. He is a member of the PGME and PGEM.

❖

Hewlett-Packard Company has announced the formation of a new division for the development and manufacture of precision components used in electronic instrumentation. The new Precision Components Division will be headquartered at the company's main plant in Stanford Industrial Park, Palo Alto, Calif. **William D. Myers** (SM'52) who has been with Hewlett-Packard for 16 years, has been named manager of the new division. He was formerly manager of the Electrical Engineering Section of the company's Manufacturing Engineering Department.

W. D. MYERS

Mr. Myers will have overall responsibility for the new Precision Components Division. He is an electrical engineering graduate of Stanford University, Stanford, Calif. Prior to joining Hewlett-Packard in 1944, he was associated with the Radio Corporation of America and with Packard Bell.

Succeeding Mr. Myers as manager of the Electrical Engineering Section will be **Frank K. B. Wheeler** (S'44–A'46–M'55–SM'58), formerly head of Hewlett-Packard's quality assurance program.

❖

MacLeod Instrument Corporation, Fort Lauderdale, Fla., instruments and electronics manufacturer, announces the appointment of **James B. O'Maley** (M'54) as vice president, engineering. He will direct the research, development and production engineering activities of the corporation.

J. B. O'MALEY

He was formerly associated with Kollsman Instrument Corporation as automation consultant to the president and also research engineer. He was responsible for the development of a number of new advanced electromechanical and electronic systems for the automation of instrument manufacture. He also pioneered in the field of celestial guidance systems for aircraft and missiles. A number of patents for automatic star followers have been granted to him. He was also associated with the development of the Gyrosyn compass, an electronic aircraft compass. At the U. S. Army Signal Corps Laboratories, he worked with design of light beam telephone equipment.

Mr. O'Malley is a member of the American Ordnance Association.

❖

FXR, Inc., of Woodside, N. Y., has recently announced the appointment of **Robert E. Othmer** (M'57–SM'58) as an engineering section head. He will direct development of electronics test equipment for microwave measurement. He did similar work for Narda Microwave Corporation, and before that was with Aeroflex Research Corporation and Emerson Radio and Phonograph Corporation.

R. E. OTHMER

Mr. Othmer received the degree in engineering from the Pratt Institute, Brooklyn, N. Y. He is an associate member of the AIEE.

❖

Sprague Electric Company recently announced the appointment of **David B. Peck** (A'53–M'58) to Vice President, Special Products. He has been associated with Sprague Electric since 1943, and is a graduate of Rensselaer Polytechnic Institute, Troy, N. Y., where he majored in chemical engineering.

Prior to joining the Sprague firm, he was with E. I. du Pont de Nemours as a Chemical Engineer. With Sprague, he served as a Patent Engineer and as a Supervisory Engineer in the Research and

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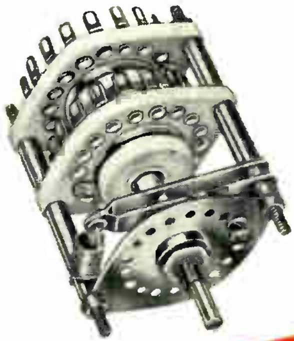

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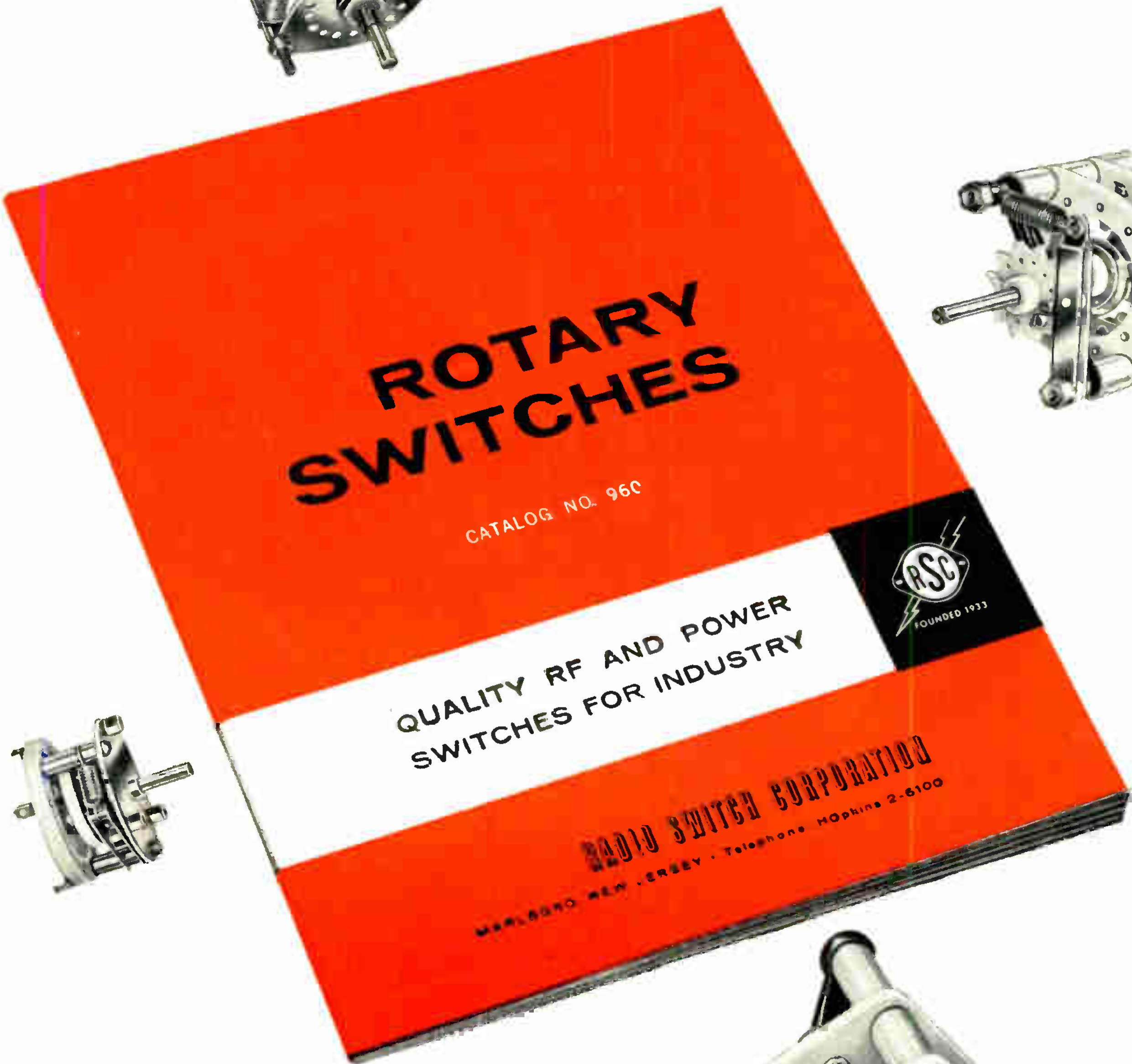


Request your copy today!

The new 1961 RSC Catalog No. 960 will be of interest to every designer and engineer concerned with radio frequency and power switching. It contains complete specifications on each switch in the RSC Line. Write for your copy today.

IRE People

(Continued from page 56A)

Engineering Department. When the Special Products Division was established in 1958, he was named to head it as its Manager.

❖

Appointment of **James A. Robinson** (M'55) as Senior Reliability Engineer for Bendix Computer Division has been announced.

Prior to joining Bendix Computer, he had served as Chief Engineer for the Walkirt Company. Previously, he had been with General Communications Company, and with the Canadian Government.

Mr. Robinson is an engineering graduate of the University of Saskatchewan, Sask., Canada.

❖

Dr. N. H. Enenstein, vice-president and laboratory director of the Data Systems Laboratory of Litton Systems, Inc., Beverly Hills, Calif., has announced appointment of **Martin Rubin** (A'47) as a senior staff engineer of the laboratory.

In his Litton Systems post, Mr. Rubin will be a member of an executive team responsible for developing simulators, trainers, and automatic checkout equipment for advanced systems.

He was formerly project engineer for all Navy automatic checkout equipment built at Autonetics.

He received the bachelor's degree in electrical engineering from Clarkson College of Technology, Potsdam, N. Y., and the master's from the University of Southern California, Los Angeles. He was a navigator with the Air Force during World War II.

❖

Stanley L. Rudnick (M'60) has been appointed Sales Manager of Ace Electronics Associates, Inc., Somerville, Mass., manufacturers of precision potentiometers. He was previously General Sales Manager of the National Company, Malden, Mass.

S. L. Rudnick

For the past fifteen years, he has held administrative positions in sales management and marketing, and is well known in the electronic equipment and radio parts industries.

He is a member of the Armed Forces Communications and Electronics Association, Sales Manager's Club, and American Society of Military Engineers.

Mr. Rudnick is a graduate of William and Mary College, Williamsburg, Va., and Boston University, Boston, Mass.

❖

(Continued on page 61A)

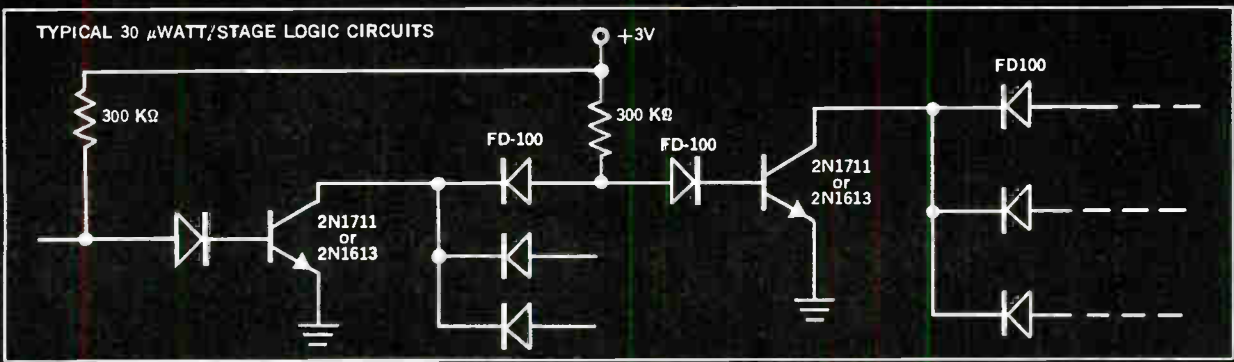
Low leakage and useful h_{FE} at very low collector currents permit low power operation — as low as 30 microwatts per stage. High performance PLANAR transistors and diodes use simplified circuitry (see illustration), keep costs down, reduce power requirements, and permit high-density packaging. Prime applications: missile and space vehicle guidance and instrumentation.

For your down-to-earth needs, where weight, space and power are less critical, similar logic circuits will operate with switching rates greater than 5 mc.

mμA leakage in Fairchild **PLANAR** transistors and diodes permits

MICROWATT LOGIC

FOR SPACE VEHICLES



MAXIMUM RATINGS AND ELECTRICAL CHARACTERISTICS (25°C except where noted)

	Total diss.	V_{CBO}	V_{EBO}	h_{FE} (I_C=150mA) (V_{CE}=10V)	h_{FE} (I_C=0.1mA) (V_{CE}=10V)	C_{ob} (I_E=0) (V_{CB}=10V)	I_{CBO}	I_{CBO} (V_{CB}=60V) (T=150°C)	I_{CBO} (V_{CB}=50V) (T=125°C)
2N1613	3.0W	75V	7.0V	40-120	20 min.	18pf typ. 25pf max.	0.8mμA typ. 10mμA max. (V_{CB}=60V)	1.0μA typ. 10μA max.	—
2N1711	3.0W	60V	7.0V	100-300	35 min.	25pf max.	10mμA max. (V_{CB}=50V)		10μA max.
FD100	WIV	P diss.		T_A	T_{stg}	I_R (V= −50V)		R_E (100 mc)	
	50V	250mW		−65° to +175°C	−65° to +200°C	0.1μA		45%	

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A WHOLLY OWNED SUBSIDIARY OF FAIRCHILD CAMERA AND INSTRUMENT CORPORATION

IRE People

(Continued from page 58A)

Raymond A. Rugge (SM'57) has been appointed Director of Engineering for Kollsman Instrument Corporation, it was recently announced.


R. A. Rugge

He will direct both the design and development programs and supervise the Engineering and Product Engineering Laboratories.

Prior to his appointment with Kollsman, he was Vice President and Director of the W. L. Maxson Corporation. In this capacity, he was in charge of both the Engineering and Research and Development Divisions and was active in the design and development of radar systems, microwave antennas, electronic and electromechanical systems for missiles such as Thor, Nike, Hercules, Hawk and others.

He is a graduate of the University of Kansas, Lawrence, Kans., with the B.S. degree in electrical engineering. He has published numerous articles in technical journals on aircraft equipment and engineering management operation. He is an Associate Fellow of the Institute of Aero-Space Sciences, and a member of AIEE, IES, AOA and of the New York Academy of Sciences. In addition, he has been an active participant in several professional groups of the IRE and committees of the AIEE.

❖

Election of **Dr. Hector R Skifter** (A'31–M'36–SM'43–F'51) as a member of the board of directors of American Research and Development Corporation has been announced by G. F. Doriot, president. Dr. Skifter is president of Airborne Instruments Laboratory, a Division of Cutler-Hammer, Inc., and also a vice president and director of Cutler-Hammer.


H. R. Skifter

He is presently a member of the Scientific Advisory Panel, Department of the Army, and the Scientific Advisory Committee, Department of Ordnance, U. S. Army. He also serves as a consultant to the President's Science Advisory Committee, Executive Office of the President, and to the Director of Defense Research and Engineering, Department of Defense. He has previously served as Assistant Director of Defense Research and Engineering (Air Defense), DOD, and Chairman of the Committee on Air Navigation of the Research and Development Board, DOD. He has been a member of the Technical Advisory Panel on Electronics, DOD; Tech-

(Continued on page 64A)

MOST MAGAZINES GET DATED

. . . with **Proceedings** it takes a little longer . . .

WHEN radio-electronics engineers want up-to-date information on any subject in their field, they turn with confidence to back copies of **Proceedings** — copies that in fact may be many months old! That's because **Proceedings'** articles are so authoritative and forward-looking that they stay timely much longer than one would expect, especially in our swiftly-changing field of radio-electronics.

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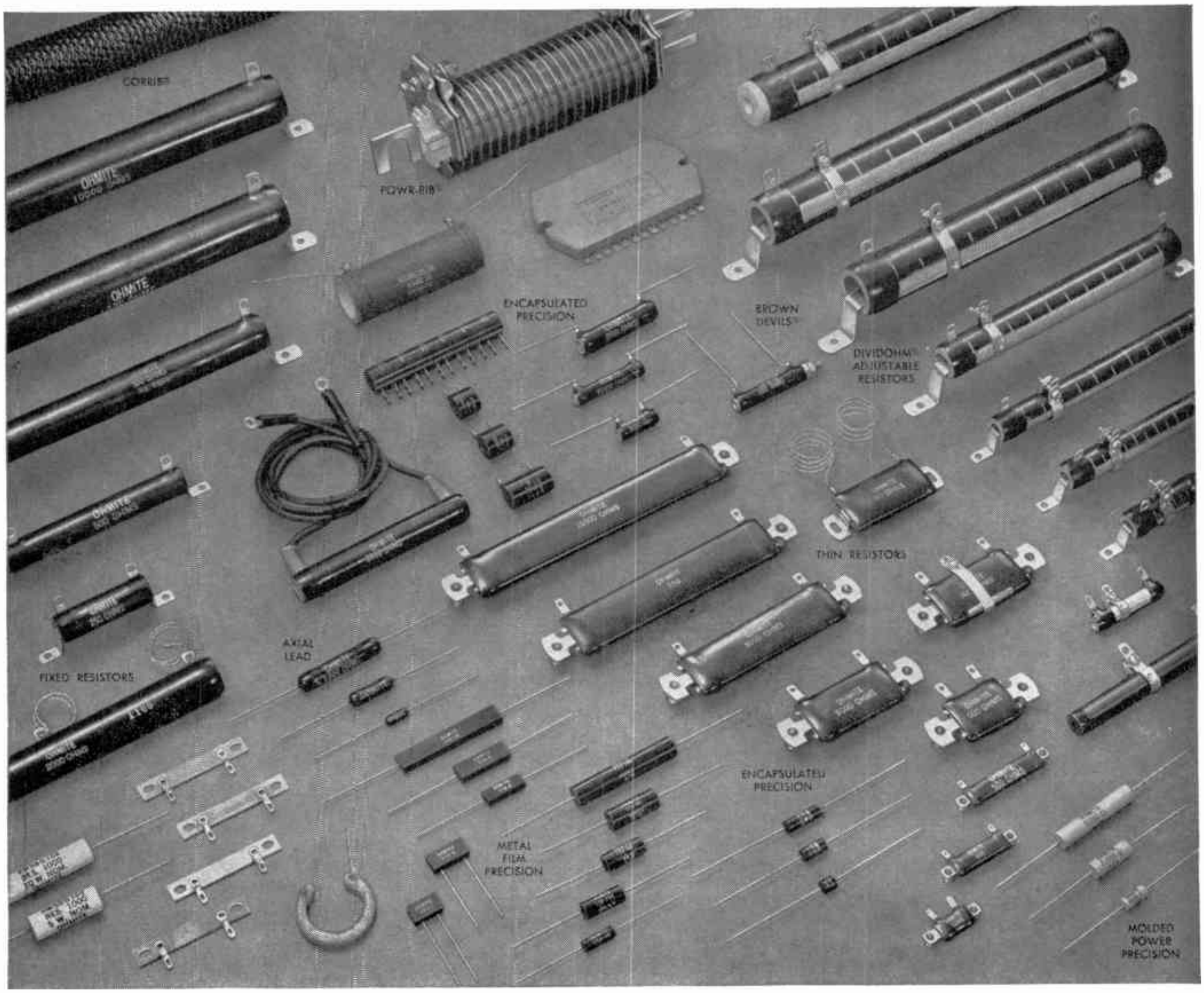

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IRE People

(Continued from page 61A)

nical Advisory Panel on Electronic Countermeasures, DOD; the Gaither Security Resources Panel of the President's Science Advisory Committee; Consultant to the Assistant Secretary of Defense, Research and Development; and Consultant to the Air Navigation Board.

During World War II, he was associate director of the Airborne Instruments Laboratory of Columbia University, New York, N. Y., operated for the Government's Office of Scientific Research and Development. In 1945, he was made president of Airborne Instruments Laboratory, Inc., after the organization was separated from Columbia University. Following acquisition of AIL by Cutler-Hammer in 1958, he became vice president and director of the parent company as well as president of Airborne Instruments Laboratory.

Dr. Skifter brings to the board of directors of American Research and Development Corporation technical and scientific capabilities as well as extensive background in the management of technically-based growth companies such as those in which the corporation's investments are made.

❖

Dr. Richard C. Sirrine (M'57) was recently appointed manager of the Surface Studies Unit of the General Electric Advanced Semiconductor Laboratory. The Surface Studies Unit is one of the five principal research areas into which the Laboratory has been reorganized in order to meet increasing demands of semiconductor technology.

R. C. SIRRINE

Dr. Sirrine joined the General Electric Company in 1956 as a physicist in the Electronics Laboratory. Since becoming a member of the ASL staff, he has specialized in the investigation of surface properties of semiconductor materials.

❖

Robert Strauss (M'52–SM'58) has recently joined ARINC Research Corporation, Washington, D. C. He will be associated with the reliability programs that ARINC conducts for the Armed Forces and Industry. His responsibilities will be, in part, devoted to tasks in the electron tube device field with particular emphasis on microwave type devices.

R. STRAUSS

Prior to joining ARINC, he was an Engineering Supervisor in the Microwave Tube Group of the Sperry Rand Electronic Tube Division, Gainesville, Fla. For the past six years, he has been engaged in the development, product refinement, and production of traveling wave type tubes and klystrons. From 1948 to 1954, he was a member of the Technical Staff of Bell Telephone Laboratory, Inc., associated with the development of various electron devices.

Mr. Strauss received the B.M.E. degree from Cornell University, Ithaca, N. Y., and the M.S.E.E. degree from Lehigh University, Bethlehem, Pa. He is a member of ASQC and Eta Kappa Nu and is a registered professional engineer in the State of New York.

❖

Dr. Jerome B. Wiesner (S'36–A'40–SM'48–F'52), director of the Research Laboratory of Electronics at the Massachusetts Institute of Technology, was elected to the board of directors of Sprague Electric Company. He has been a consultant to the company since 1957. **Dr. Jerrold R. Zacharias** (SM'57), distinguished nuclear physicist, was named as a consultant.

Dr. Wiesner has been a leader in the rapid development of communication sciences and is a member of President Eisenhower's Science Advisory Committee. In 1948, he was awarded the President's Certificate of Merit, the second highest civilian award, in recognition of "outstanding services to his country." He served on the committee that prepared the Gaither Report and was a staff director of the American delegation to the 1958 Geneva Convention on the prevention of surprise attack.

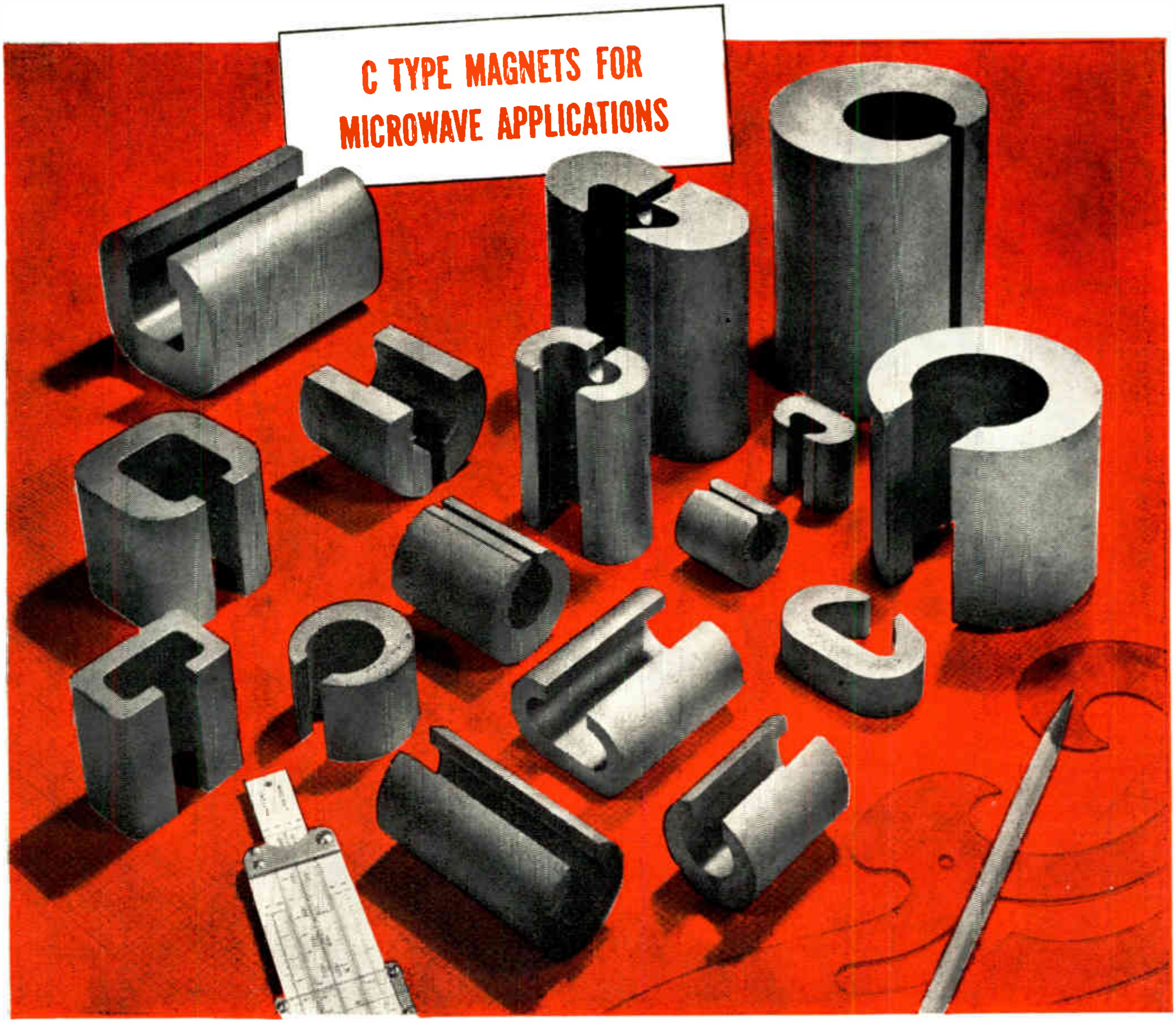
He was born in Detroit, Mich., in 1915, and received the degrees of bachelor of science, master of science, and doctor of philosophy from the University of Michigan, Ann Arbor, in 1937, 1938 and 1950, respectively.

In 1940, he was appointed chief engineer of the Acoustical and Record Laboratory in the Library of Congress, Washington, D. C. There, under a Carnegie Corporation grant, he assisted in developing sound recording facilities and associated equipment.

Shortly after the beginning of World War II, he joined the staff of M.I.T.'s Radiation Laboratory as associate leader of the radio frequency development group. Later he became project engineer of a key radar development program and a member of the laboratory's steering committee. In 1945, he joined the staff of the Los Alamos Laboratory, where he served for a year. Upon his return, he resumed his duties as assistant professor at M.I.T.

Eta Kappa Nu Association, the electrical engineering honor society, voted Dr. Wiesner an honorable mention as an outstanding young electrical engineer for the year 1947. He is a Fellow of the American Academy of Arts and Sciences and a member of the Acoustical Society of America. He is a consultant for the Department of Defense and a member of the Army Scientific Advisory Panel.

(Continued on page 66A)



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IRE People

(Continued from page 64A)

Dr. Zacharias is currently a professor of physics at the Massachusetts Institute of Technology, Cambridge. After earning his Ph.D. degree at Columbia University, New York, N. Y., in 1932, he dedicated himself to the study of the shapes of nuclei. Later he joined the staff of the Radiation Laboratory, established at M.I.T., for the wartime development of radar. In 1944, he went to Los Alamos to direct the engineering division in work on the atomic bomb. He returned to M.I.T. in 1945 to establish the laboratory for Nuclear Science, which he subsequently directed for ten years, and saw the development of major contributions in the field of cosmic rays, neutron physics, and theoretical physics.

In 1948, he was appointed associate director of Project Lexington to study the problem of nuclear powered flight. This was the first of a series of summer studies and projects in which he became involved. In 1950 he directed Project Hartwell, which dealt with undersea warfare, and in 1951 was associate director of Project Charles, a study of air defense. This led to further studies in 1952 under Project Lincoln, out of which grew the M.I.T. Lincoln Laboratory. He initiated and directed a summer study in 1952 on the Distant Early Warning Line. He was awarded the President's Certificate of Merit in 1948 and in 1955 he received the Department of Defense Certificate of Appreciation—its highest civilian honor. He is also a member of the National Academy of Sciences.

In addition to directing the laboratory for Nuclear Sciences, teaching and participating in various study groups, Dr. Zacharias was busy with his own researches, which were primarily concerned with the shape of atomic nuclei, using a molecular beam technique that had been developed during his work before the war with Dr. Rabi. One outgrowth of that work—and incidental to it—was the cesium atomic clock.

❖

Eugene L. Woodcock (S'47–A'50–SM'54) has been appointed to the Senior Technical Staff of the Electro-Optical Division, Perkin-Elmer Corporation, Los Angeles, Calif., and will be responsible for the development of infrared systems. Perkin-Elmer, with headquarters in Norwalk, Conn., established its West Coast Operations facility a year ago to serve space and defense industries in the West.

E. L. Woodcock

Until his recent assignment, he was Chief of the Development Section concerned with military infrared detecting equipment and instrumentation at Perkin-Elmer's Norwalk plant. In this capacity, he was responsible for the development of airborne infrared detecting systems, a series of controlled infrared sources, high temperature gas radiation detectors and an infrared secondary standard calibration system. Before joining P-E, he was associated with Sperry Gyroscope Company, Great Neck, L. I., as an electronic engineer assigned to the development of missile and radar systems.

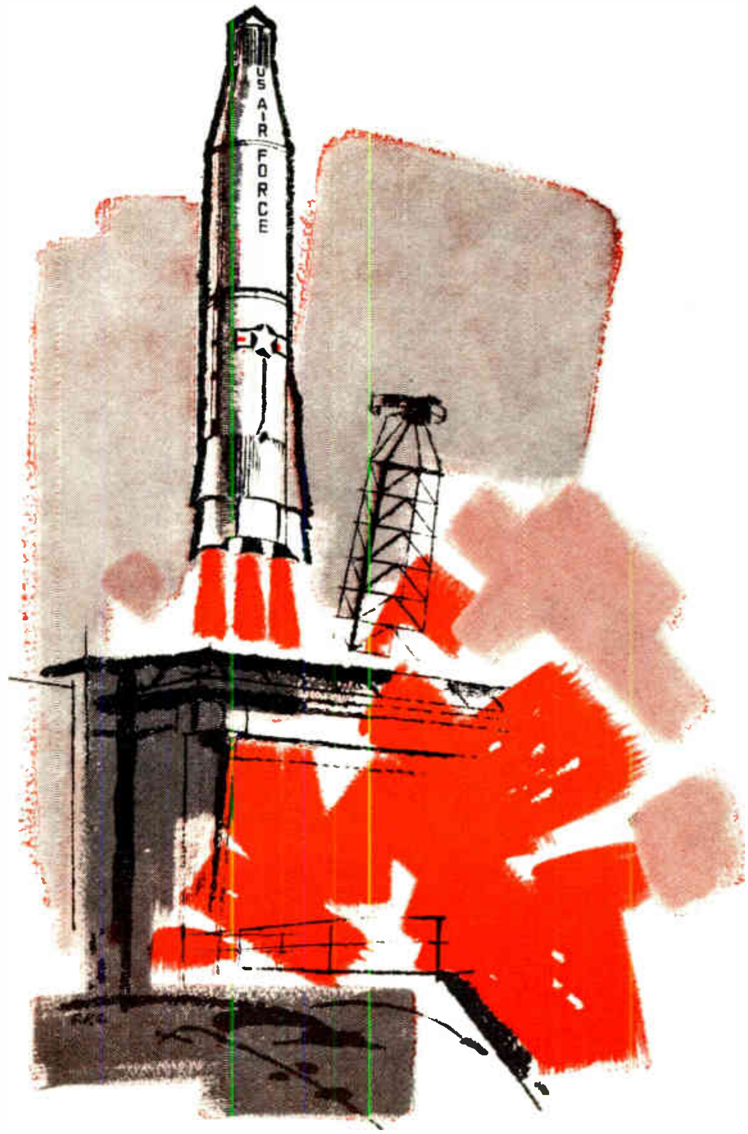
A graduate of Bates College, Lewiston, Me., he received the M.S. degree in engineering science and applied physics from Harvard University Graduate School, Cambridge, Mass. He is a member of the Infrared Information Symposium (IRIS), the Optical Society of America and Phi Beta Kappa. He holds several patents in the field of electronics.

❖

Albert B. Worch (M'58) has become a partner of Halbar Associates manufacturer's representatives, Los Angeles, Calif. Halbar Associates represents companies specializing in the AC and DC Power and Control fields, such as American Electronics, Inc., Precision Power Division, Regulator Engineering & Development Company, Acoustica Associates, Inc., and Electric Regulator Corporation. Prior to joining Halbar, Mr. Worch held the position of Sales Manager at R. S. Electronics Corporation, amplifier-receiver developer and manufacturer, of Palo Alto, Calif.

❖

(Continued on page 68A)



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IRE People

(Continued from page 66A)

Dr. William V. Wright (S'55–M'56) has been named a Vice President of Electro-Optical Systems, Inc., according to a recent announcement. He will continue in his present position as Manager of the company's Solid State Div sion; however, he will also acquire responsibilities for policies concerning proprietary programs, subsidiaries, and expansion activities.

He has been with Electro-Optical Systems since 1957, and as Manager of the Solid State Division has been responsible for directing the company's activities in solid state device and materials research.

Prior to joining EOS, he was Program Director, Semiconductor Materials, for Pacific Semiconductors, Inc., where he was active in all phases of semiconductor technology.

❖

Rainer Zuleeg (A'53–M'55) has been appointed senior staff physicist of Hughes Aircraft Company's semiconductor division in Culver City, Calif., it was announced recently.

In this new position, which is equivalent to a department head within the division, he will conduct advanced research on the theory of parametric amplification and other studies on semiconductor properties. Earlier this year, his work in developing a parametric mode transistor, which may greatly extend the range of radar and communications systems while reducing them in size, was announced to the electronics industry.

In addition to his current transistor studies, he developed a patented forming technique for early point-contact transistors, and was among the first to explore techniques of out-diffusion in semiconductor materials. He has seven patents issued or applied for, and twelve technical publications of his work.

Before joining Hughes in 1959, Mr. Zuleeg was associated with the semiconductor research laboratory of Sprague Electric Company for six years. He attended the Oberreal schule Weissenburg, Hochschule of Bamberg from 1947 to 1949, and the University of Munich from 1949 to 1951. He is a member of the Professional Group on Electron Devices and the American Physical Society.

Professional Group Meetings

Antennas and Propagation

Akron—November 15

"A Three Dimensional Automatic Pattern Analyzer," H. Fulmer, Scientific Atlanta, Inc., Atlanta, Ga.

Boston—November 29

"A Year's Study of the Index of Ionospheric Storminess," R. J. Cormier and R. A. Swirbalus, AF Cambridge Res. Labs., Bedford, Mass.

"A Method for Reducing the Effects of Aperture Blocking in Center-Fed Reflector Antennas," A. R. Stratoti, Sylvania, Waltham, Mass.

Los Angeles—November 10

"Report on 13th URSI Triennial Assembly," J. M. Kelso and R. C. Hansen, Space Technology Labs.

San Francisco—November 9

"The Trapping of Electrons from a Nuclear Detonation in the Earth's Magnetic Field," Dr. R. Dyce, Stanford Research Institute, Menlo Park, Calif.

Antennas and Propagation Microwave Theory and Techniques

Columbus—October 18

"Effective Aperture of Antennas," Dr. Chen To Tai, Ohio State University Antenna Lab., Columbus, Ohio.

Los Angeles/Orange Belt—November 17

"Interaction of Plasma and Electromagnetic Waves," Dr. R. Elliot, UCLA.

Philadelphia—November 16

"Problems in Design of 1000′ Radio Telescope," B. Hagerman, DEECO.

"Feed for 1000′ Radio Telescope." Dr. A. Kay, TRG, Inc.

Washington D. C.—November 11

"Lens Structures and Artificial Dielectrics," K. S. Kelleher, Aero Geo Astro Corp., Alexandria, Va.

Washington, D. C.—November 22

"Principles of Traveling Wave Antennas," Dr. A. A. Oliner, Polytechnic Institute of Brooklyn.

Audio

Baltimore—November 23

"Modern Loudspeakers for Stereo-Design by Listening Approach," L. R. Mills, Recordings Inc., Baltimore, Md.

Chicago—October 19

"An Organ Reverberation Device for High-Fi Stereo," B. Staffen, Hammond Organ Co., Chicago, Ill.

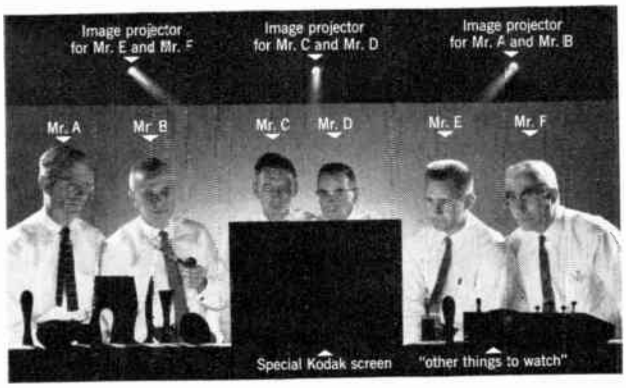
(Continued on page 72A)

Kodak reports on:

aiming information into the eyes . . . a mask for glass, switched by light . . . strange dances in the movies

The final transfer

For an honest purpose these able gentlemen have consented to the indignity of posing a tableau in supercharged staring.



Because they and numerous unpictured coequals are personally involved, they wish to call attention to a certain technical area in which we think we are good. It deals with the art of projecting information into human eyes, the final transfer from machine to man.

The black rectangle labeled "Special Kodak screen" stands for a multitude of possibilities, some which we have already demonstrated and others needing more study. Study, as is well known, costs money. It would be smarter use of the money to do the studying in the context of specific viewing and display requirements—review of vast volumes of reconnaissance photography, for one currently popular example. The composition and design of the screen should not be considered in isolation from the projectors, the eyes, the restrictions on their location, the ambient light, the nature of the visual task, and all the other pertinent factors.

On this broad and subtle subject we have neither off-the-shelf literature nor off-the-shelf products, but we are anxious to be in contact with those whose interest in it is more urgent than academic. Such persons should communicate with Eastman Kodak Company, Apparatus and Optical Division, Rochester 4, N. Y.

An invitation to engrave

Etching is, of course, not the only way to dig into a glass surface. With sufficient patience and skill a grinding wheel yields superb results. If time flits too rapidly for that sort of monkeying around, you coat a resist over your surface, scribe through it, and let the HF go to work.

In case the pattern is intricate, or needs to be repeated, or both, you want a photosensitive resist. Then you can draw up the pattern once, nice and big and black, reduce it photographically onto a *Kodalith* material and use the resulting photograph as a mask which determines where the resist comes off and exposes the naked glass to HF.

Think a moment what you are asking of any photosensitive resist. It must be capable of being switched by a reasonable amount of light from one to the other of two conditions: a) tenacious adherence to the particular material you wish to etch and impenetrability to agents which rapidly attack that material; b) abject submission to attack by agents which do not affect the substrate, or alternatively, full permeability to appropriate etchants for the substrate.

Obviously, *we* have given this matter much more than a moment's thought. Our researches have now brought forth a photosensitive resist for glass and silicate ceramics to join our previously announced *Kodak Photo Resist* ("KPR," for copper, clear anodized aluminum, and high-copper alloys) and *Kodak Metal-Etch Resist* ("KMER," for other metals). We would be justified in trying to recover all that thinking expense by selecting a similar proprietary name to imply the discovery of a new chemical compound but have decided on a cleverer course—

We shall have you buy *Kodak Metal-Etch Resist* and tell you how to convert it to a glass-etch resist by the use of those two arcane compounds, technical-grade aluminum stearate and sulfur-free xylene.

For details, write Eastman Kodak Company, Graphic Reproduction Division, Rochester 4, N. Y. If you don't want to bother stating your problem, just say "photosensitive resists."

Favor for the high-speed congress

Dust Performs for Plant's Pollution-Control Movies, *Chem. Week*, 84:84, 86, May 2, 1959. (The Procter and Gamble Company uses high-speed motion-picture sequences for the qualitative control of in-plant dust.)

The Ign tion of Explosives by Radiation, J. Eggert, *J. Phys. Chem.*, 63:11-15, Jan., 1959; also in *Photochemistry in the Liquid and Solid States*, edited by F. Daniels, J. Wiley, N. Y., 1960, pp. 147-53. (High-speed photography proves that the detonation of nitrogen iodide starts before the light flash ends, showing that only a fraction of the energy is used for the detonation.)

Lathe Check Formation in Douglas-fir Veneer, E. H. Collins, *Forest Products J.*, 10:139-40, March, 1960. (High-speed motion pictures were used by Weyerhaeuser Company to analyze production variables.)

Time after time we have visited a customer proud of some accomplishment with high-speed movies. He is willing to show us—eager, delighted to show us. The projector is started and we watch. We see a collection of strange objects. We don't know for sure what they are. Little seems to be happening. After quite a while, a new object enters the scene from the left. Shortly another new object comes up from the bottom. The two dance around each other, touch, and exit from the top of the frame. All is again static on the screen. After another while the reel comes to its end and we jump to our feet exclaiming hearty congratulations.

He deserves congratulations, probably. If we had lived with the problem as he has, the objects in the picture might have seemed no stranger than the face in the bathroom mirror; the dance might have been the triumphant, forceful, sudden, undisputed clincher to a vexatious problem; the all-purpose enthusiasm of the born salesman might have meant more.

Nevertheless, we need not be ashamed. We help scientists and engineers use high-speed photography by manufacturing a group of films to the stringent mechanical requirements of high-speed cameras. *Kodak Plus-X Reversal Film* we make for reversal processing to a fine-grain positive. *Kodak Tri-X Reversal Film* is four times as fast. *Kodak Double-X Panchromatic Negative Film*, which is a bit faster yet and very sharp, is picked when a quick negative will suffice or when several prints may be wanted later. *Kodak Royal-X Pan Recording Film* is picked only when light is very limited indeed; *Kodak Linagraph Ortho Film*, for accentuated sensitivity to green light; *Kodak High Speed Infrared Film*, for sensitivity to 9000A, with a maximum from 7700A to 8400A; *Kodachrome Film*, for color, with low-cost commercial processing widely available; *Ektachrome ER Film*, for color at exposure index of 160 or higher.

Another thing. A bibliography on high-speed photography. Every item our library knows. Coverage extends into 1960. Got it ready to distribute to the Fifth International Congress on High-Speed Photography in Washington in October. Doomed to a short life, since the Congress promptly generated so many *new* papers on high-speed photography that the abstracts alone run from p. 609 to p. 682 of the September, 1960, issue of the *Journal of the Society of Motion Picture and Television Engineers.*

Eastman Kodak Company, Photorecording Methods Division, Rochester 4, N. Y., would be glad to send the bibliography or answer questions about the above-named films.

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Backscatter or synchronized oblique sounding.

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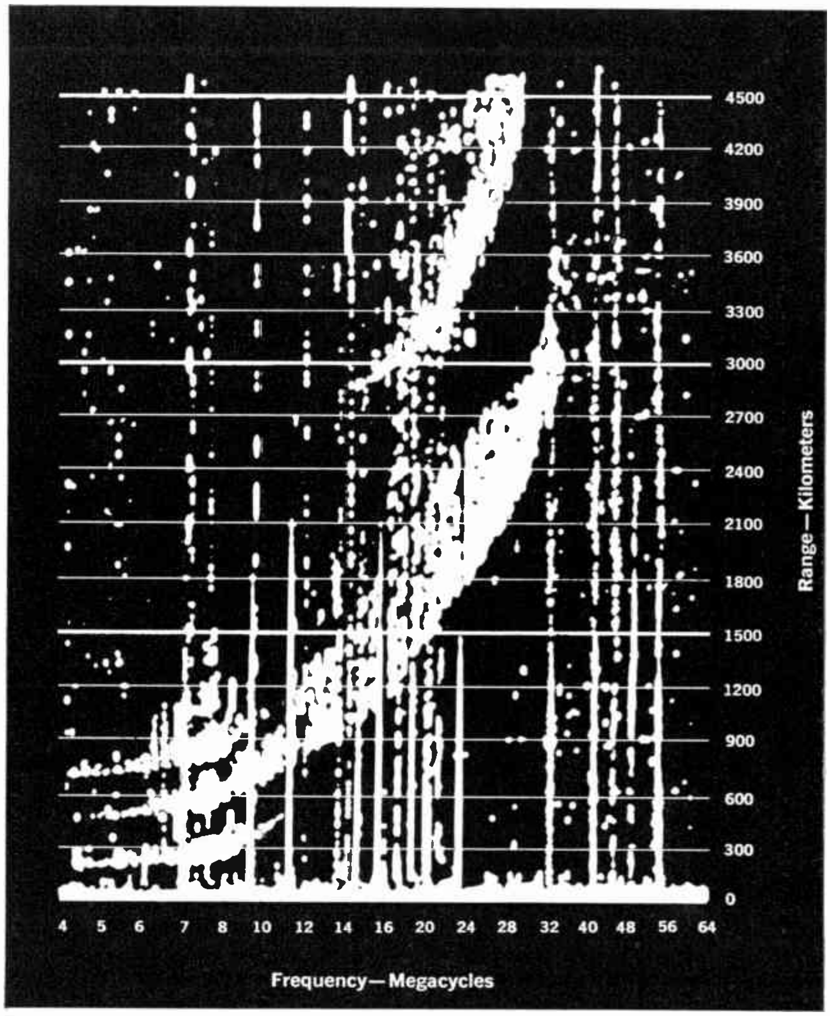
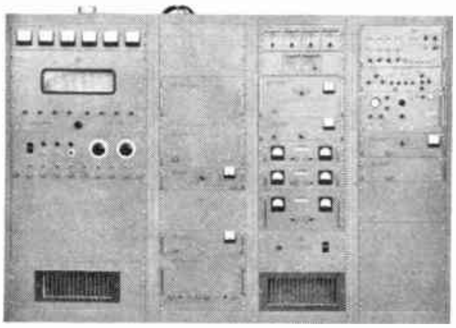
for rapid, precise measurement of ionosphere parameters on a continuous basis.



Typical oscilloscope record shows vertical incidence and backscatter returns using G/A Model 902



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Granger ASSOCIATES

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Professional Group Meetings

(Continued from page 68A)

Chicago—November 16

"Effective Use of the Level Recorder in Acoustic Measurement," M. Basch, General Radio Co., West Concord, Mass.

Philadelphia—October 27

"Linosoid Modulation of Multiplex Techniques," B. Wise, Industrial Transmitters and Antennas, Lansdowne, Pa.

San Francisco—November 2

"Design Considerations of Audio Transformers," S. Hose, Triad Transformer Corp., Venice, Calif.

Automatic Control

Chicago—November 11

"An Active-Circuit Analogue of a Torque-Reflecting Servosystem," E. J. Miller, Motorola, Inc., Chicago, Ill.

Los Angeles—November 8

"Adaptive Autopilot Study for a High Performance Interceptor Missile," J. C. Simmons, Douglas Aircraft Co., Santa Monica, Calif.

Milwaukee—November 15

"Feedback Controls for Inertially Guided Missiles," R. Brown, AC Spark Plug Div. of General Motors, Oak Creek, Wis.

Philadelphia—Oct. 20

"The Model Reference Adaptive Control System," H. P. Whitaker, MIT, Cambridge.

Bio-Medical Electronics

Cleveland—November 16

"Electrochemistry," Dr. R. Weast, Case Inst. of Tech., Cleveland, Ohio.

"Conduction of Electrical Impulses by Living Cells," Dr. T. Hoshiko, Western Reserve School of Medicine, Cleveland, Ohio.

Los Angeles—November 17

"Homeostatic Mechanisms," Dr. D. H. Simmons, VA Hospital, Los Angeles, Calif.

"pH Blood Measurements," E. Valenzuela, Electro-Medical Engineering Inc., Burbank, Calif.

Bio-Medical Electronics
Electronic Computers

Boston—November 2

"Computer Analysis of Brain Waves," Dr. A. Remond, Natl. Center of Scientific Research, Paris, France.

Broadcasting

Omaha-Lincoln—December 2

"The New WOW TV AM FM Office Installation," G. Flynn, Chief Engineer & Bldg. Supt.

First Open House at the New Meredith WOW Building, W. J. Kotera, M. McGown, H. Stutzman, C. Hagerman.

Philadelphia—October 13

"Philosophy of TV Station Automation," F. McNichols, RCA, Camden, N. J.

Circuit Theory

Chicago—November 11

"An Active Circuit Analogue of a Torque-Reflecting Servosystem," E. J. Miller, Motorola Inc., Chicago, Ill.

Component Parts

Los Angeles—November 21

"Trends in Electron Tubes," and Plant Survey, R. J. Zeh, Pioneer Electronics Corp., Los Angeles, Calif.

Philadelphia—October 18

"International Standardization of Component Parts," S. Zwerling, General Electric Co., Philadelphia, Pa.

Washington—October 12

"Thermoelectricity," Dr. C. Zener, Westinghouse Electric Corp., Pittsburgh, Pa.

Electron Devices

Los Angeles—November 7

"Silicon Crystal Growth Technology," Dr. A. Stevenson, Pacific Semiconductor, Culver City, Calif.

Syracuse—November 22

"A Review of Modern Flat Display Devices," Dr. H. B. Law, RCA Labs., Princeton, N. J.

Washington, D. C.—November 21

"Electrostatic Printing Tubes," J. E. Gerling, Litton Industries, San Carlos, Calif.

Electron Devices
Microwave Theory and Techniques

Boston—October 13

"Low Noise Microwave Amplification," Dr. G. Wade, Raytheon (Spencer Lab.) Burlington, Mass.

San Francisco—November 9

"Getting Close to the Ultimate Noise Limit," A. E. Siegman, Stanford University, Stanford, Calif.

Electronic Computers

Boston—November 10

"Combinatorial Techniques for Performing Arithmetic and Logical Operations," H. Fleisher, IBM Corp.

(Continued on page 76A)

Professional Group Meetings

(Continued from page 72A)

Chicago—November 11

"Transistorized Apparatus for Binary to Decimal Conversion," R. D. Carlson, Radiation Counter Labs., Inc., Skokie, Ill.

Los Angeles—October 20

"Super Conductive Computers—The State of the Art," Dr. J. T. Rogers, Space Technology Labs.

San Francisco—October 25

"Self Adaptive Systems," H. J. Bremmerman, Univ. of Calif., Berkeley.

San Francisco—November 15

"Table Lookup and Language Translation," J. Griffith, IBM Mohansic Res. Lab., Yorktown Heights, N. Y.

Twin Cities—November 9

"An Application of a Medium Speed Digital Computer in Real Time," D. Burrows, General Mills, Minneapolis, Minn.

Engineering Management

Washington—September 26

"Advanced Management System for Advanced Weapons Systems," W. R. Fazar, Navy Bureau of Weapons, Washington, D. C.

Washington—November 28

"Motivation for Research," Dr. R. M. Page, Naval Res. Lab., Washington, D. C.

Engineering Writing and Speech

San Francisco—November 15

"Patent Disclosures and Claims," J. F. Lawler, Sylvania MVO, Mountain View, Calif.

Washington D. C.—October 26

Business Meeting.

Instrumentation

Chicago—November 11

"Instrumentation for Large Scale Testing," P. Gottfried, Reliability Engineering Associates, Skokie, Ill.

San Francisco—November 1

"Missile Range Instrumentation," A. Smolen, ITT Labs., Nutley, N. J.

San Francisco—November 22

"State of the Arts, Instrumentation, Magnetic Recording," W. B. Heinz, Ampex, Redwood City.

Washington—December 5

"Radio Collision Avoidance Systems for Aircraft," R. T. Fitzgerald, Diamond Ordnance Fuze Lab., Washington, D. C.

Microwave Theory and Techniques

Boston—November 10

"The State of the Art in Ferrite Devices," Dr. H. Scharfman, Raytheon Co., Waltham, Mass.

Chicago—November 11

"Bridging the Microwave-Infrared Gap," Dr. P. D. Coleman, Univ. of Illinois, Urbana

Los Angeles—October 6

"Magnetically Tunable Microwave Filters," P. S. Carter, Jr., Stanford Research Institute, Menlo Park, Calif.

Omaha-Lincoln—November 9

"Applications of Microwave Relay by the FAA," P. Hertel, Jr. and Robert Millner, Collins Radio Co., Dallas, Texas.

Military Electronics

Boston—October 20

"Military Systems Engineering," A. P. Hill, Mitre Corp., Bedford, Mass.

Detroit—November 16

The group visited the University of Michigan Radio Telescope where they were given a tour and complete description of the facilities.

(Continued on page 80A)

now 100 times greater average power

Now you can obtain traveling-wave tubes capable of 10 to 100 times the average power of conventional helix tubes. These X-band tubes are representative of a wide variety of the first commercially available all metal filter-type structures yielding both high gain and wide bandwidth. Their attractively small size and weight are made possible through application of the latest periodic focusing techniques.

Typical of these recent advances is the pictured 308H. For the first time a power traveling-wave tube is offered with a high-μ grid-controlled gun. This advantage, coupled with 53 db of saturation gain, provides exciting possibilities for the systems designer.

Consult with Hughes Microwave Tube Division if you have exacting design requirements of pulse rise time, phase shift sensitivity, bandwidth or power output. These qualities are yours in a light, compact, yet rugged package of all metal-ceramic construction. These advanced products can make your program a success. Orders are being accepted now for delivery in three or four months.

307H LEFT: 100 kw peak power output (500 watt average), 8.5—9.5 kmc frequency range, 54 db saturation gain, 1% maximum duty cycle, beam voltage = 38 kv, 21 lbs. total weight of tube and magnet.

308H

Control grid μ = 55. 15 kw peak power output (150 watt average), 8.6—9.9 kmc frequency range, 53 db saturation gain, 1% maximum duty cycle, beam voltage = 24 kv, 14 lbs. total weight of tube and magnet.

319H

20 kw peak power output (200 watt average), 8.4—9.6 kmc frequency range, 54 db saturation gain, 1% maximum duty cycle, beam voltage = 24 kv, 17 lbs. total weight of tube and magnet.

For full details on these and other equally outstanding tubes write or wire Hughes Microwave Tube Division, 11105 Anza Avenue, Los Angeles 45, California.

Creating a new world with ELECTRONICS

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MICROWAVE TUBE DIVISION
HUGHES AIRCRAFT COMPANY

Professional Group Meetings

(Continued from page 76A)

Los Angeles—November 16

"STL Able Program which led to Explorer VI and Pioneer V," R. G. Stephenson, STL.

Northwest Florida—November 1

"History and Application of Semiconductors," M. Linhurst, Motorola Inc., Chicago, Ill.

Oak Ridge—November 17

"Recollection of Russian Instrumentation," P. R. Bell, Oak Ridge National Lab.

Philadelphia—October 13

"Role of ARPA," Brig. Gen. A. W. Betts, ARPA, Pentagon.

Rochester—November 10

"The State of the Art of Radar Systems," R. S. Durnell, Stromberg-Carlson, Rochester, N. Y.

San Diego—November 9

"Optical Beacon Design," R. J. Jacobs, Convair-Astronautics, San Diego.

Syracuse—November 9

"The Perceptron," C. E. Young, Cornell Aeronautical Lab., Buffalo, N. Y.

A panel of leading engineers to discuss "The Impact of Adaptive Machines on the Industry."

Nuclear Science

Los Angeles—November 16

"Field Trip," A. Goldstein, B. Ault, and J. L. Shepherd, U. S. Nuclear Corp., Burbank, Calif.

Product Engineering and Production

Boston—October 19

"Europe and Japan Challenge our Production Techniques," R. G. Zens, Electralab Printed Electronics Corp., Natick, Mass.

Philadelphia—October 26

"Influence of the Aircraft and Missile Industry of Electronic Design Progress," Capt. R. Barnaby, Franklin Institute, Philadelphia, Pa.

San Francisco—November 22

"Design and Fabrication of Magnetic Components" J. Biggerstaff, Palo Alto Engineering.

L. Burkurt, Palo Alto Engineering, Plant Tour.

(Continued on page 82A)

Professional Group Meetings

(*Continued from page 80A*)

Reliability and Quality Control

Boston—October 13

"Queing Theory," Dr. N. Hazelwood, Arthur D. Little Co., Cambridge, Mass.

"Human Factors in Reliability," Dr. J. Spiegel, Mitre Corp., Bedford, Mass.

Columbus—October 19

"Stress Analysis Applications in Reliable Pipe Line Design," G. M. McClure, Battelle Memorial Institute, Columbus, Ohio.

Los Angeles—October 17

"Description of Hoffman Semiconductor Division Facility," M. Whitney, Hoffman Electronics Corp., El Monte.

"Review of Application of Solar Products in Space Devices," K. L. Ray, Hoffman Electronics Corp., El Monte.

"Special Measurement Techniques," H. Rauschenbach, Hoffman Electronics Corp., El Monte.

Los Angeles—November 21

"Reliability is Everybody's Business," P. A. Adamson, Hughes Aircraft, Fullerton, Calif.

"Reliability is Principally an Engineering Problem," L. R. Landrey, Burroughs Corp., Paoli, Pa.

"Reliability is Principally a Management Problem," J. M. Wuerth, Corp. Reliability Coord., Autonetics Div., NAA, Downey, Calif.

Philadelphia—May 24

"Reliability as a Profession," C. M. Ryerson, RCA, Camden, N. J.

"Reliability as a Profession," W. T. Sumerlin, Philco Corp., Philadelphia.

"Reliability as a Profession," J. J. Hamrick (presented paper by L. R. Landrey) Burroughs Corp., Paoli.

San Francisco—October 18

"A Reliable Satellite Communications System," G. O. Moore, Philco WDL, Palo Alto, Calif.

Space Electronics and Telemetry

Detroit—November 30

"Space Electronics and Satellite Television Techniques in the USSR," D. Ritchie, Bendix Res. Labs., Southfield, Mich.

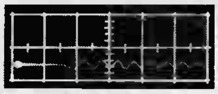
Philadelphia—October 11

"Physical & Aerodynamic Factors affecting Re-entry Telemetry," Dr. W. C. King, General Electric Missile & Space Vehicle Dept., Philadelphia, Pa.

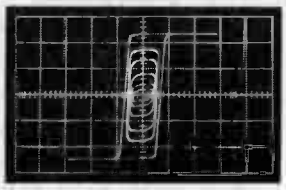
(*Continued on page 84A*)

You can do MORE with the NEW TEKTRONIX C-12 CAMERA than you can with any other Oscilloscope Camera

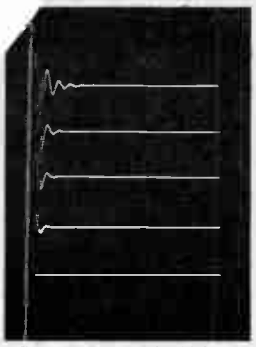
...picture this



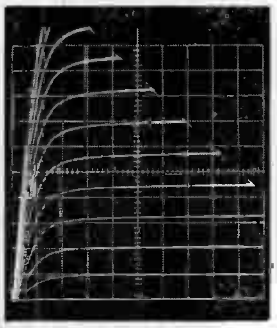
1 KMC damped oscillation (single shot at 2 nsec/cm).



Typical hysteresis loops (multiple exposure of varying amplitudes).

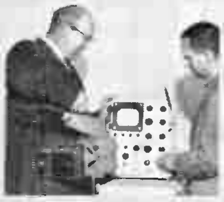


Damped sine wave (multiple exposure using all 5 detent positions).



Family of characteristic curves (for NPN transistor).

...like this



One-hand portability

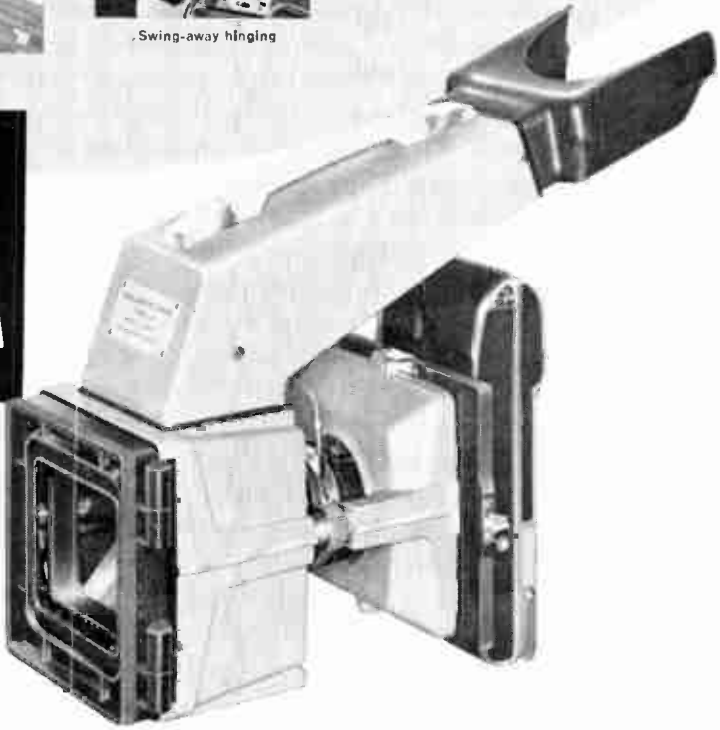
Lift-on mounting

Swing-away hinging

Comfortable viewing—with or without glasses

...only *with this*

C-12 CAMERA



The C-12 Camera combines flexibility with simplicity. It offers you new convenience in undistorted viewing and direct recording of oscilloscope traces.

Here's why:

You can use Polaroid* or any conventional film.

You can use the unique sliding back, adjustable to horizontal or vertical. On this sliding back, you can interchange the parfocal, film-holding backs, lock them securely in 5 detent positions, also rotate them thru 90° increments (with the long axis of the film horizontal or vertical).

You can choose from 8 easily-interchangeable lenses — in varying object-to-image ratios and maximum aperture to f 1.5. The lenses are housed in uniform, pre-focused, calibrated mounts with keyed threads — so the shutter-speed and diaphragm-opening controls always appear at the same accessible position on the camera.

C-12 CAMERA **$500**

Includes: f 1.9 Lens (with 1:0.9 object-to-image ratio) complete with cable release, Focusing Back, Polaroid Back, and Minute Timer.*

*Registered by Polaroid Corporation.

For more information about these and other features of the new C-12 Camera . . . and the many accessories designed for specialized applications . . . call your Tektronix Field Engineer.

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P. O. Box 500 • Beaverton, Oregon

Phone Mitchell 4-0161 • TWX—BEAV 311 • Cable: TEKTRONIX

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Professional Group Meetings

(Continued from page 82A)

Philadelphia—November 8

"Signal Processing in the Tiros Instrumentation System," M. H. Mesner, Astro Electronics Div. of RCA, Princeton, N. J.

San Francisco—September 20

"Physiological Parameters in Biomedical Instrumentation," S. Hall, Lockheed, Palo Alto, Calif.

San Francisco—October 18

"Reliable Communications via Satellite," G. O. Moore, Philco WDL, Palo Alto, Calif.

San Francisco—November 15

"Space Radiation Effects on Telemetry Components," Dr. F. Moser, Lockheed Missiles & Space Division, Palo Alto, Calif.

Los Angeles—October 20

"Instrumentation for Vehicular Radio Service," B. Morris, Marconi Instruments, Englewood, N. J.

Los Angeles—November 15

"Police and Fire Department Communication Centers," A. Brooks, Westrex Corp., Los Angeles, Calif.

Vehicular Communications

Washington—October 18

"Remote Outrage Location by means of VHF Radio Systems," M. Cooper, Motorola C & E Inc., Chicago, Ill.

Washington—November 15

"The Carterfone," T. F. Carter, Carter Electronics Corp., Dallas, Tex.

Vehicular Communications
Communications Systems

Philadelphia—October 18

"Characteristics and Applications of the Tunnel Diode," J. A. Ekiss, Philco Corp., Lansdale, Pa.

Membership

The following transfers and admissions have been approved and are now effective:

Transfer to Senior Member

Allen, G., New Orleans, La.
Beach, O. R., Winston-Salem, N. C.
Bricout, P. A., Silver Spring, Md.
Choate, R. L., Pasadena, Calif.
Coble, R. P., Jr., Greensboro, N. C.
Collmer, R. C., Canoga Park, Calif.
Craiglow, L. H., Rome, N. Y.
Di Franco, J. V., Old Bethpage, L. I., N. Y.
Eaton, C. B., Levittown, N. J.
Estin, A. J., Boulder, Colo.
Marcus, P., Holliswood, L. I., N. Y.
McMillan, J. G., Omaha, Nebr.
Nijenhuis, W., Eindhoven, Holland
Peshel, R. L., San Carlos, Calif.
Robinson, T. B., Winchester, Mass.
Stearns, W. P., Scottsdale, Ariz.
Virmani, B. D., Lynchburg, Va.

Admission to Senior Member

Aust, A. T., Den Haag, Netherlands
Bassett, J. B., Madison, Conn.
Bronder, R. T., Bellevue, Wash.
Davies, P. M., Manhattan Beach, Calif.
Della Giovanna, C., Milan, Italy
Gray, W. F., University, Ala.
Harder, E. L., East Pittsburgh, Pa.
Hollingsworth, L. E., Livermore, Calif.
Holmes, W. C., Saratoga, Calif.
Ittermann, A. S., Winston-Salem, N. C.
Keehner, J. B., Akron, Ohio
Lawless, W. J., Jr., New York, N. Y.
Osgood, R. M., Winchester, Mass.
Roda, C., Milan, Italy
Rosato, D. V., Wahan, Mass.
Tenzer, M., West End, N. J.
Tucker, J. F., Darien, Conn.
Weiss, H. K., Santa Ana, Calif.

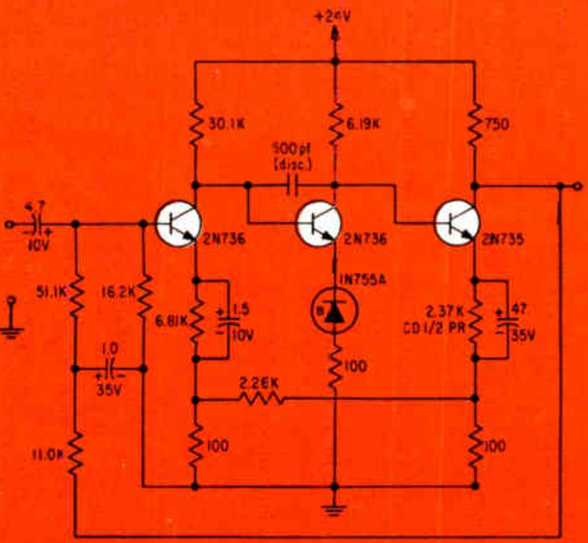
Transfer to Member

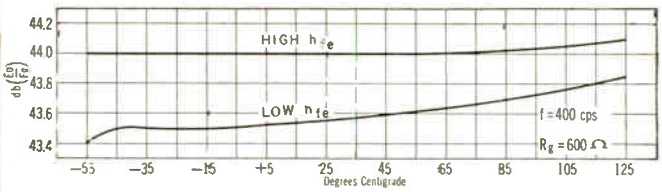
Alde, R. O., Silver Spring, Md.
Allman, D. P., San Diego, Calif.
Amato, W. P., Baltimore, Md.
Annett, M. E., Bristol, Pa.
Aubry, J. D., Saratoga, Calif.
Baldini, B. P., Silver Spring, Md.
Banshak, W. G., Tonawanda, N. Y.
Barlowe, M., Bethpage, L. I., N. Y.
Barse, G. R., Silver Spring, Md.
Bloom, R. A., Bethesda, Md.

(Continued on page 93A)

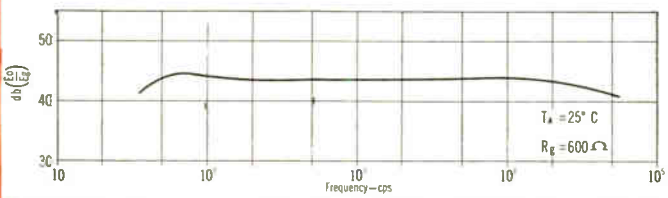
HOW TO GET FLAT FREQUENCY RESPONSE

FROM 37 CYCLES TO 45 KC





Voltage Amplification vs. Ambient Temperature

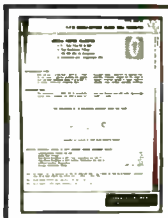


Frequency Response of Typical 2N735-2N736 Transistors

Specify TI 2N734 Series Silicon Transistors for Your Amplifiers!

For your audio/servo amplifiers, power supplies and medium-speed switches, design in TI 2N734 Series Silicon Transistors. Obtain a flat frequency response of ±1.5 db from 37 cycles to 45 kc . . . guaranteed beta at 25°C (1 ma at 1 kc) (5 ma at 1 kc) (5 ma at 30 mc) and at −55°C (5 ma at 1 kc) . . . guaranteed 500-mw free-air dissipation . . . reduced equipment size and weight with TO-18 package.

For even greater power dissipation, investigate the design flexibility of the equivalent TO-5 packaged 600-mw TI 2N1564 Series Silicon Transistors.



Write for complete specifications on 2N734 and 2N1564 Series Today!

SPECIFY TI FOR ALL YOUR SILICON TRANSISTOR REQUIREMENTS • Small Signal • Switchers • Medium Power • Power • Order TODAY from your local TI Sales Office or TI Distributor.

GUARANTEED MINIMUM BETAS

Symbol	Parameter	Test Conditions	2N734	2N735	2N736
h_{fe}	A-C Common-Emitter Forward Current Transfer Ratio	V_{CE} = 5v I_E = 5ma f = 1 kc T_A = 25°C	20	40	80
h_{fe}	A-C Common-Emitter Forward Current Transfer Ratio	V_{CE} = 5v I_E = 1 ma f = 1 kc T_A = 25°C	15	30	60
h_{fe}	A-C Common-Emitter Forward Current Transfer Ratio	V_{CE} = 5v T_A = −55°C I_E = −5 ma f = 1 kc	12	20	40
$\|h_{fe}\|$	A-C Common-Emitter Forward Current Transfer Ratio	V_{CE} = 5v I_E = 5 ma f = 30 mc T_A = 25°C	1	2	2

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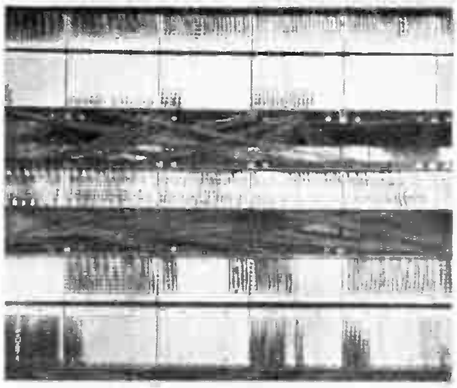
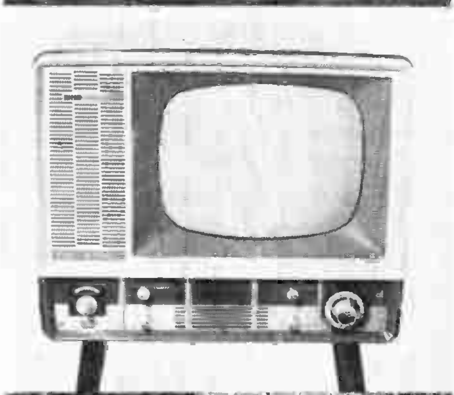
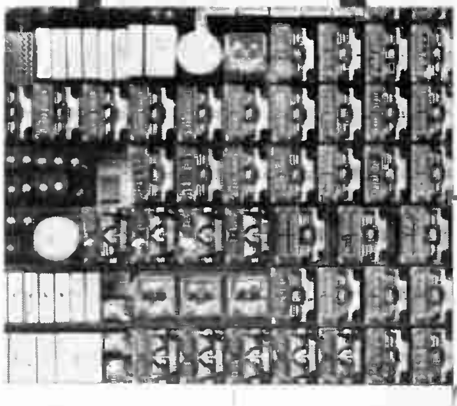
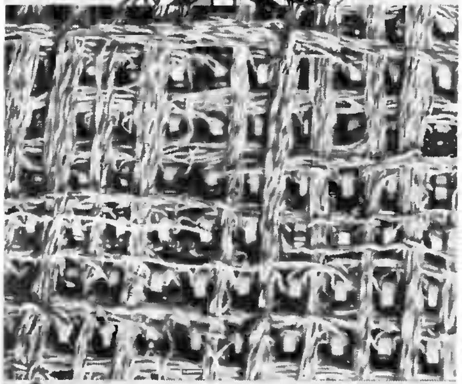
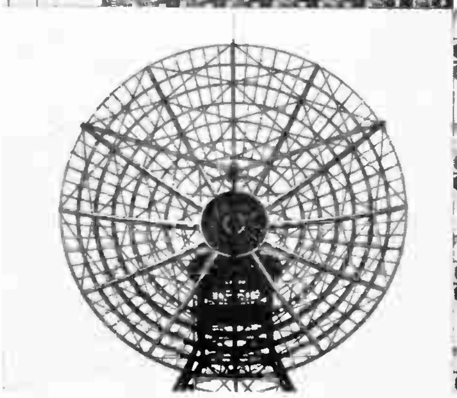
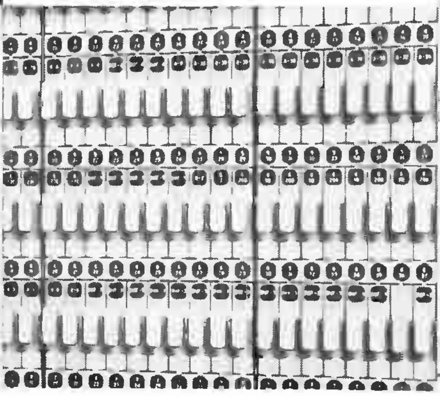
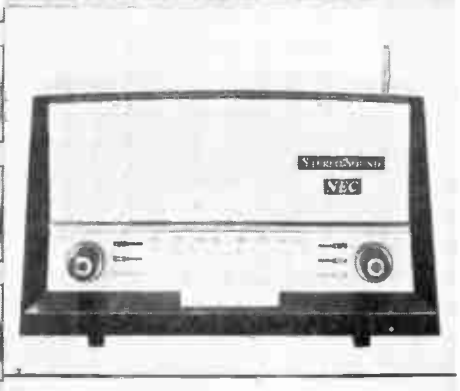
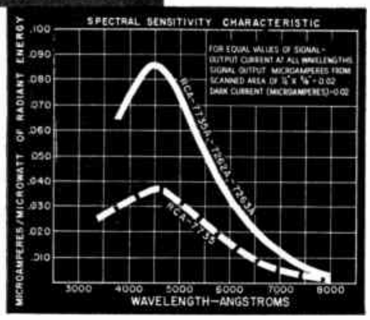






Unequalled Light Sensitivity and Low Lag

New RCA photoconductor utilized in world's most sensitive vidicons



Now, an RCA exceptionally sensitive photosurface ushers in a new family of advanced vidicons unequaled anywhere for their light sensitivity and closely-controlled characteristics.

The RCA-7735-A—double the sensitivity of the 7735—provides good pictures at lower light levels with less lag than previously possible! This sensitivity and the tube's lower "lag" make the 7735-A ideal for industrial use.

Two short-length vidicon types featuring the same sensitivity as the 7735-A are offered for use in compact transistorized cameras. RCA-7262-A is intended for normal outdoor-indoor environments. RCA-7263-A is especially designed for military or other applications involving shock, vibration, humidity, and altitude. Both types operate at 30% less heater power than any other commercially available type of vidicon.

All three of these advanced camera tubes feature a resolution capability of 600 to 900 lines, broad spectral response, and the high tube-to-tube uniformity that has become identified with RCA vidicons.

For complete information about these RCA vidicons—the most sensitive in the world—get in touch with: Marketing Manager, Industrial Tube Products, RCA, Lancaster, Pa. For a technical bulletin on any of these types, write Commercial Engineering, Section B-35-Q, RCA, Harrison, N.J.

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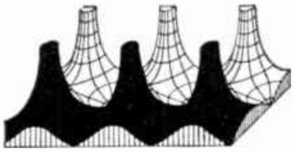
February, 1961
Vol. 49 No. 2
Proceedings of the IRE

Poles and Zeros



The Old and the New. Combining an old, discarded concept or device with the newest available theory and practice frequently produces new and sometimes startling results. Throughout the history of an art or science, ideas, developments, and devices once important and relevant disappear as the new supplants them. The older an art becomes the more difficult it is for those who are entering the field to become familiar with the past and the lessons that it teaches. This has already happened in radio and in electronics despite their comparative youthfulness. There is significance associated with history and danger associated with its neglect.

One aspect of this problem is readily apparent in engineering education. The immense expansion of engineering knowledge poses the problem of what should be taught and what should be deleted. Educators, in their valid concern with the present, their respect for the past, and their dedication to the challenge of the future must face this dilemma. To varying extents they must neglect the detailed steps in the development of their subject, perhaps to an unwarranted degree. The young engineer must accept as a requirement for his own development the importance of scrutinizing the past and familiarizing himself with the background of his art. For this purpose the publications of the IRE are a fertile source for reference as are many volumes to be found in the engineering and science libraries.

One recalls, for example, that Hertz's original experiments used very short waves (as then defined). Marconi, in applying Hertz's experimental evidence of the existence of electromagnetic radiation to communications, found that longer wavelengths gave greater communications range. From that time on, for many years, long waves were considered more useful for communications than short waves. In fact short waves were considered so useless that the annual report of the Federal Radio Commission for the year ending June 30, 1928 indicated that until about 1926–1928 the high frequency band had been supposed to be virtually useless for practical purposes. It cited the erratic behavior of these frequencies and the technical difficulty of the construction of apparatus in this frequency range. The report then noted that intensive study, and experimentation and the development of apparatus capable of transmitting and receiving high frequencies, had demonstrated that they could be used reliably for communications over great distances with comparatively small amounts of power.

This change in understanding of the usefulness of the high frequency portion of the spectrum was a result of combining the disclosures of Hertz with the development, in the early 1900's, of the thermionic tube. The almost forgotten short waves returned to the scene as this new device entered the art and provided a new tool for the generation and detection of short waves. Their early discard as a communications tool was recognized as invalid.

Prior to the advent of the thermionic tube the crystal detector was an accepted and useful part of radio communications systems. The superior performance characteristics of the tube caused the disappearance from the scene (and the textbooks) of the crystal detector. As tube and circuit developments brought higher and higher frequencies into use, the limitations of the thermionic detector were revealed and the resurrection of the crystal detector took place. Again the old and the new combined to advance the art. The crystal detector reappeared both in microwave systems and in the textbooks; will it disappear again?

Frequency modulation as a communications tool provides another example of the disappearing-reappearing phenomenon. It certainly appeared that frequency modulation, as a technique, was laid to rest in 1922. In the ensuing decade developments in higher frequency techniques combined with the genius of Armstrong brought frequency modulation into being as a significant communications system technique.

With these three illustrations (perhaps they are not the best available) as background one is not surprised that the era of space age communications again demonstrates the dependence of progress on the combination of the old and the new. During 1960 long distance telephone conversations were achieved first using the moon, and subsequently the satellite Echo, as passive reflectors. One of the systems accomplishing this communications feat employed a new, extremely sensitive receiving system. This receiving system is an excellent illustration of the importance of history combined with the newest developments. The arrangement comprised an electromagnetic horn antenna commented on in the literature at least as early as 1936, receiver circuitry comprising a special frequency modulation feedback circuit proposed by J. G. Chaffee (PROC. IRE, May, 1939), and combined these with one of the newest developments of the art, the ruby maser.

It is fun to speculate as to which combination of the old with the yet to be discovered will next appear to solve an important new problem.

Iteration. The December, 1960 Poles and Zeros reported the formation of the AIEE-IRE Joint Standards Committee. Lest there be any misunderstanding, this note is to emphasize that the Joint Committee does not replace the individual Standards Committees of the two societies; each of these continues operations as before.—F. H., Jr.

John F. Byrne

Vice President, 1961

John F. Byrne (SM'45–F'50) was born on October 26, 1905, in Cincinnati, Ohio. He attended The Ohio State University, Columbus, receiving the B.S. degree in engineering physics in 1927, and the M.S. degree in electrical engineering in 1928.

After a year with the Bell Telephone Laboratories, he returned to Ohio State as a faculty member in 1929; he was assistant professor of electrical engineering when he left the University in 1937 to join the Collins Radio Company. In 1942 he became associated with the newly-formed Radio Research Laboratory at Harvard University, Cambridge, the first laboratory organization to devote its time exclusively to the development of electronic countermeasures equipment and techniques. He was appointed Associate Director of the Laboratory in January, 1945. During the period 1946–1950, he was Vice President in charge of research and engineering at the Airborne Instruments Laboratory in Mineola, N. Y. Mr. Byrne has been with Motorola since 1950, first as Director of Engineering for the Communications Division, and now as General Manager of the Systems Research Laboratory at Riverside, Calif.

He has served on several government committees since 1942. He was chairman of the Electronic Countermeasures Panel of the Research and Development Board, 1949–1951, and is a past member of the Advisory Council for the Army Electronic Proving Ground at Fort Huachuca.

For his work during World War II, he received the U. S. Navy Certificate of Commendation in February, 1947, and the Presidential Certificate of Merit in September, 1948.

He became a senior member of the IRE in 1945, and received the Fellow Award in 1950, "for his development of a system of polyphase broadcasting and for effective engineering administration in connection with countermeasures during the war." He was a Director of the IRE from 1955–1957, and has served on various IRE committees, including Tellers, 1949, and Awards, 1953–1955. He also served as Chairman of the Professional Group on Military Electronics, Los Angeles Section, 1956–1957.

Mr. Byrne is a member of Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

Scanning the Issue

Circuit Considerations Relating to Microelectronics (Suran, p. 420)—Microelectronics is currently a subject of great interest to the electronics profession. It has also become a topic of some controversy. Many predictions have been made about the revolutionary possibilities which await us, but little has been said about the formidable problems that exist and must be resolved before many of these possibilities can become realities. It is now becoming evident, for example, that the power and heat problems encountered in microsystems will have an important bearing on the maximum allowable packing density of components. It is not enough that components be made smaller. Circuit power dissipation, also, must be reduced. The power problem in microelectronic systems is explored in some detail by the author. In particular, he shows how power requirements are dependent on circuit speed and component tolerances, and discusses the effects of these relationships on packing density. The author's comparison of the packing densities frequently quoted these days with those that are currently achievable is both sobering and enlightening. The analysis provides much needed food for thought for the many persons interested in this new field.

Some Technical Aspects of Microwave Radiation Hazards (Mumford, p. 427)—The microwave equipment available in 1940 was capable of generating only 10 watts average power. By 1965 this figure will probably reach the one-million-watt level. This dramatic increase in power generating capability at microwave frequencies is bringing with it an attendant increase in potential hazards to operating personnel. These hazards have already been the subject of extensive investigation during the last decade. Most of the evidence to date indicates that the chief effect of microwave energy on living tissue is to produce heating. The nature of this heating, its relation to average power, time of exposure and frequency, and the susceptibility of various parts of the body to radiation damage are now broadly understood. Less well understood are the possible effects of peak power, as contrasted with average power, and certain non-thermal effects. Nevertheless, sufficient data has been gathered to permit the Armed Services and leading industrial laboratories to reach general agreement, within the last three years, on maximum safe exposure limits for personnel. This review of microwave radiation studies over the years, and of the findings and recommended safety measures that resulted, provides timely information on a subject of growing interest and importance. Included in the discussion is a recommended method of calculating power densities, a survey of available power density meters, and a nomograph for calculating the shielding effect of wire mesh screen.

A Very Low Frequency (VLF) Synchronizing System (Looney, p. 448)—Last year both the U. S. Navy and the National Bureau of Standards began transmitting standard frequency and time signals in the VLF band. Because of the long range and the remarkable phase stabilities of VLF signals the inauguration of these important services has now made it possible to synchronize local frequency standards in widely scattered locations with an accuracy of better than one part in 10 billion. One result is that these VLF signals are now being used to synchronize the world-wide satellite tracking network with improved accuracy and are paving the way for new and more accurate satellite experiments. This paper describes a receiving system by which VLF signals are used to control the frequency of a local frequency standard. The system is unique in that it does not control the local oscillator frequency directly, but rather controls the output of a time standard by continuously shifting the output phase to correspond with the received VLF reference phase.

Exact Solution of a Time-Varying Capacitance Problem (Macdonald and Edmondson, p. 453)—The analysis of circuits having time-varying elements has been hampered in the past by the lack of methods which would yield exact solutions for the harmonic components. Problems of this type are becoming of quite broad importance, especially with the recent interest in parametric amplifiers. This paper employs the powerful artifice of making a direct Fourier analysis of an intractable integral indefinite, which results in closed form expressions for the coefficients. This enables the authors to obtain an exact solution for the harmonics generated by a sinusoidally varying capacitance in series with a fixed resistor and battery. As the authors show, their new method has many applications of interest to circuit theory and audio engineers, including the analysis of microphones, loudspeakers, vibrating reed electrometers, and frequency multipliers.

Supplement to IRE Standards on Graphical Symbols for Electrical Diagrams (p. 467)—This Standard adds 16 symbols and deletes one from the IRE Standard of the above title which was published in 1954. One addition is a symbol for a gas filled radiation counter tube. The remaining changes concern waveguide components.

Pseudo-Rectification and Detection by Simple Bilateral Nonlinear Resistors (Bridges, p. 469)—Detection and rectification, which we usually associate with unilateral devices such as diodes, can also be achieved with bilateral nonlinear resistors when a pulse-type ac waveform is impressed upon them. This interesting property, which can be found in a number of resistors in common use today, is analyzed in a simple fashion and is shown to be useful not only for developing dc bias but also in noise analysis and the enhancement of signals.

The Amplitude Distribution and False Alarm Rate of Noise After Post-Detection Filtering (Thaler and Meltzer, p. 479)—The passage of noise through a narrow-band filter, and several types of detectors and post-detector filters has been simulated with a digital computer. The numerical results, which are presented in a series of graphs, and the new detectors described will be of fundamental interest to the many readers who are involved with radar systems.

IRE Standards on Radio Transmitters: Definitions of Terms (p. 486)—The IRE Committee on Radio Transmitters has defined some three dozen terms pertaining to transmitters and their use. The definitions have now been approved by the Standards Committee as IRE Standard definitions.

The Quadratic Invariances of Generalized Networks (Pease, p. 488)—This paper is concerned with analyzing systems in terms of relationships that are invariant. Such invariants, when they exist, are significant because they specify many important properties of the system. The invariant approach has application to whole classes of systems and will be of interest to a large number of people concerned with electronic amplification and active circuit theory, including parametric circuits.

Experiments on Magnetic Tape Readout with an Electron Beam (Freundlich, *et al.*, p. 498)—If a beam of electrons is passed over the surface of a magnetic tape and is focused on a fluorescent screen, it is possible to observe on the screen a pattern showing details of the magnetic field of the tape. This novel idea has been explored to determine the feasibility of using electron beams for magnetic tape readout, a method which is inherently capable of reading more densely packed recorded data. The results will not only be of wide interest but should stimulate further work along new and potentially important lines.

Scanning the Transactions appears on page 537.

Circuit Considerations Relating to Microelectronics*

J. J. SURAN†, SENIOR MEMBER, IRE

Summary—**Five basic problems are associated with microscale circuits: power dissipation, thermal generation and its effects on component packing density, compatibility, adjustability and reliability.**

Power dissipation and its relation to circuit functions is discussed in detail and specific circuit problems are used as examples of the principles. Furthermore, it is shown that the power dissipation problems are *fundamental* and relate generally to *all* classes of electronic components. Power dissipation and heat generation are then related to packing density to determine the physical limitations of microelectronic fabrication.

Introduction

In recent months the subject of microminiaturization has become one of the "controversial" topics in the electronics field. However, implicit in the storm, is the general feeling that potentially significant developments are in the making, and heard now and then, in the cacophony of contradictory verbiage, is an appeal for scientific evaluation. There are indeed major problems associated with microelectronics, and the sooner these problems are exposed to objective engineering evaluation, as opposed to the prevailing conjectures of hucksters, the sooner will the fruits of the vast efforts currently underway be realized.

Among the stated objectives of microelectronics are the reduction of size and weight, increase in reliability, reduction in cost, improvement in power utilization and increase in functional capabilities per unit volume of electronic equipment. Size and weight reduction is to be achieved by new fabrication techniques involving advanced material processes such as electrolytic deposition, or evaporation of thin films, etc. Increased reliability is to be achieved by the reduction of solder connections, or their replacement by chemically-bonded material interfaces, and by the increased use of redundancy on the component or circuit level. Cost reduction is to be a by-product of improved mechanized construction and assembly techniques. Furthermore, it is hoped that the vast efforts in material technology will lead to the development of new basic or complex devices which may ultimately supersede the transistor and related conventional devices by improved performance characteristics and more efficient power utilization.[1]

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† Electronics Lab., General Electric Co., Syracuse, N. Y.

[1] Device terminology used in this paper follows the definitions proposed by I. A. Lesk, *et al.*, in the paper, "A categorization of the solid state device aspects of micro-system electronics," Proc. IRE, vol. 48, pp. 1833–1841; November, 1960.

Pervading the hopes and objectives of microelectronics are several fundamental problems which are still unresolved or unsolved. These problems relate to such considerations as power requirements and heat dissipation, microfabrication compatibility, circuit adjustability, cost and reliability. By microfabrication compatibility is meant the ability to microminiaturize enough of a given system to realize on the system level the size, weight, cost and reliability objectives of the material processes. Since most input-output components, including such devices as motors, display panels, ac-dc power converters, antennas and shielded low-noise amplifiers, are not amenable to current methods of microminiaturization, it is questionable whether microelectronics has significant size and weight advantages to offer, in the foreseeable future, to many classes of electronic equipment. The circuit adjustability problem stems from the usual procedures followed in trouble-shooting or aligning complex equipment. Many of the component packing density claims prevalent in the popular technical literature ignore the adjustability problem, or inherently assume a level of circuit design sophistication which is nonexistent. Nevertheless, there is some legitimate expectation that adaptive circuit techniques, once developed, may overcome many of the adjustment problems. Reliability is too complex a consideration to be treated in this discussion, albeit improved reliability is generally accepted as a *de facto* result of microminiaturization on the expectation of a reduced number of solder connections and increased use of redundancy.

The power and heat problems encountered in microsystems have been belatedly recognized as a tempering influence on component packing density objectives (exuberant claims, such as 10,000 components per cubic inch, are beginning to assume more modest proportions). It is the object of the following discussion to explore the power problem in some detail. In particular, it is pertinent to ask why transistors, which are capable of operation at power levels of several hundred microwatts, are used in practical circuits in such a way that power dissipation levels of tens or hundreds of milliwatts are employed. Why are such power levels required? And can power be reduced sufficiently in transistor circuits to allow packing densities on the order of 10^3 to 10^4 components per cubic inch to be achieved practically in microelectronic systems? If the latter question can be answered affirmatively, one major step will have been taken toward realization of the size and weight objectives of microelectronics.

Energy and Speed

In mechanics the power of a force is defined as the time-rate at which the force is doing work. Thus, the energy required to lift a weight against the force of gravity to a given height is fixed and independent of the speed of the process (in a dissipationless medium), but the power required in the prime mover varies directly with the desired speed. This basic principle of mechanics applies to electrical circuits as well. Every circuit is required to perform work, namely, to energize a load. The energy requirement is fixed by the nature of the signal and load, and is independent of the speed or bandwidth of the circuit. However, since information is transmitted through the circuit by varying the electrical potential or charge flow in the load, the circuit power level will vary directly with the rate of information.

The well-known relationship between energy, W, power, P, and time, t, *viz.*,

$$W = Pt, \tag{1}$$

is manifested in electrical circuits by a frequency-power curve of the type illustrated in Fig. 1. A certain minimum amount of power is generally required to overcome the nonlinearities of the circuit components at very low current and voltage levels. At high frequencies, the finite gain-bandwidth product of the active circuit components limits the speed at which useful power gain can be achieved. Consequently, the minimum power requirement and the maximum speed of the circuit are generally determined by component or device properties. But between these two limits, a direct relationship exists between circuit speed and circuit power dissipation; power goes up as circuit speed or bandwidth increases.

A specific illustration of this principle is shown in Fig. 2, where the maximum pulse repetition frequency of a transistor flip-flop is plotted as a function of the power dissipation level of the circuit.[2] Terminal requirements of the circuit have been normalized so that *relative* driving requirements are not a function of power level. For example, at any power level the circuit has been designed to afford a loading factor of unity; *i.e.*, the flip-flop can drive an external load equal to the nominal value of its load resistor. Similarly, fixed current and voltage stability factors have been assumed which are independent of the power level of the circuit. Silicon transistors are used to assure maximum temperature stability of the circuit, even at very low power dissipation levels.

The curve of Fig. 2 is obtained by designing the flip-flop circuit with normalized design parameters, while considering pulse repetition frequency as a variable design parameter. Static power dissipation of each circuit so designed is then plotted as a function of frequency. Circuits operating below the 500-μ watt power level are severely speed-limited because of the drop-off of the current amplification factor α_E, with concomitant decreasing emitter current. Between 500 μw and 50 mw, the flip-flop speed is governed primarily by circuit time constants, such as the coupling network response time and the load resistor-collector capacity time constant. Increase of power levels beyond 50 mw does not allow significantly higher speeds, because of the gain-bandwidth limitation of the transistors. Consequently, between pulse repetition frequencies of approximately 100 kc and 2 Mc, the circuit speed is limited only by the power level of the circuit.

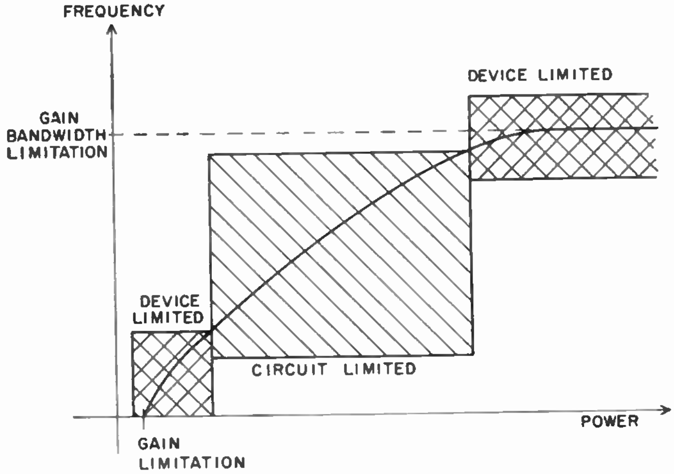


Fig. 1—Speed vs power dissipation.

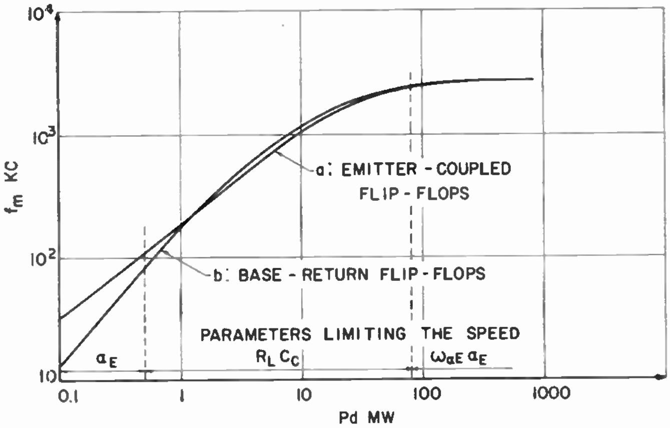
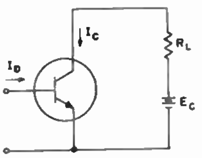


Fig. 2—Flip-flop PRF vs power dissipation.

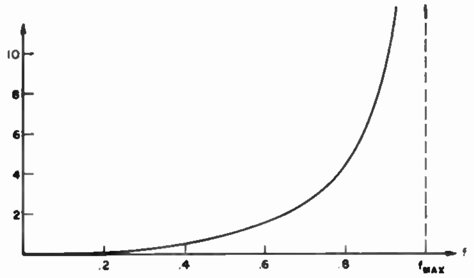
Another example of the fundamental speed-power relationship is illustrated in Fig. 3, for the case of a simple transistor switch. If the transistor is driven into saturation by the drive current I_D, the rise time of the output current I_C, is given approximately by

$$t_R = \frac{G_i(1 + \alpha_{EO}\omega_{\alpha E}R_L C_C)}{\alpha_{EO}\omega_{\alpha E}}, \tag{2}$$

[2] H. Raillard and J. J. Suran, "Speed vs circuit power dissipation in flip-flops," Proc. IRE, vol. 47, pp. 96–97; January, 1959.



(a)



(b)

Fig. 3—Power-speed relation in transistor switch. (a) Transistor switch. (b) Normalized power vs switching frequency.

where G_i is the digital current amplification:

$$G_i = \frac{\dfrac{E_{CC}}{R_L}}{I_D}.$$

In (2), $\alpha_{EO}\omega_{\alpha E}$ is the current gain-bandwidth product of the transistor for constant voltage output (short circuit) and R_LC_C is the load resistor-collector capacity time constant. Neglecting input power and saturation resistance, the average power dissipation of the circuit during the switching interval is

$$\overline{P}_R = \frac{E_C^2}{2R_L}. \tag{3}$$

The power given by (3) is expended primarily in discharging the collector capacity; the energy required for this process is

$$W = \frac{E_C^2C_C}{2}. \tag{4}$$

Combining (2), (3) and (4) results in

$$\overline{P}_R = \left(\frac{G_iW}{t_{\min}}\right)\left[\frac{1}{(t_R/t_{\min}) - 1}\right], \tag{5}$$

where $t_{\min}$ is the minimum rise time given by

$$t_{\min} = \frac{G_i}{(\alpha_{EO}\omega_{\alpha E})}.$$

If the maximum operating frequency of the switch is defined as

$$f_{\max} = \frac{1}{2t_{\min}}, \tag{6}$$

(5) may be written as

$$\overline{P} = (G_iWf_{\max})\left[\frac{(f/f_{\max})}{1 - (f/f_{\max})}\right], \tag{7}$$

where the operating frequency of the switch, f, is

$$f = \frac{1}{2t_R},$$

and where $\overline{P}$ is the circuit power dissipated per cycle of switching. Eq. (7) is plotted in Fig. 3(b), and it is apparent that circuit power dissipation increases almost exponentially as the switching speed is increased beyond 0.5 $f_{\max}$. For speeds below 0.5 $f_{\max}$, the power-speed relationship is almost linear. In this linear range, (7) may be approximated by

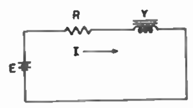
$$\overline{P} = (G_if)W. \tag{8}$$

In (8), (G_if) may be considered as the *circuit* gain-bandwidth product. Hence, for a *given circuit function*, the only way to reduce power dissipation is to reduce the energy term in (8). Since the energy requirement is dictated by the transfer of charge through a potential difference, it is apparent that one way to reduce circuit power dissipation is to operate at low voltages with active devices and loads having minimum reactances. However, low energy circuits are more susceptible to noise and component variations, and, consequently, energy levels cannot be reduced with impunity. The noise problem is particularly severe in transistor circuits operated at low current and low voltage levels because of the relatively high input impedance and low trigger energy characteristics of such circuits; stabilization always entails increasing the circuit power level. A detailed examination of the component tolerance problem will further illustrate the power-price paid for circuit stabilization.

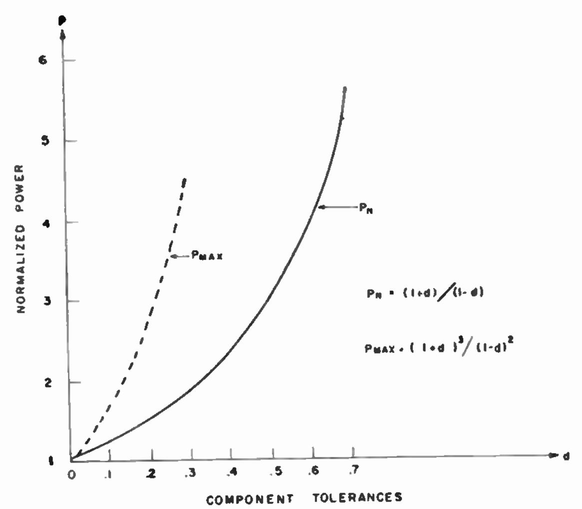
Entropy and Power

Since all components exhibit a "spread" in values around their nominal ratings, the circuit designer never works with ideal components. Allowing greater component tolerances in a circuit design is synonymous to allowing a greater degree of randomness in the selection of circuit components. From a statistical point of view, this may be thought of as equivalent to an increase of entropy in the circuit. However, since the circuit designer cannot permit the greater degree of component randomness to be reflected in the *performance* of the circuit, it is apparent that some measure of increased energy expenditure must be expected if the second law of thermodynamics is to remain inviolate.

A simple example of this principle is illustrated by the circuit of Fig. 4(a), which shows a relay coil Y in series connection with a battery E and a current limiting resistor R. In order for the relay to operate, it is necessary that the current I exceed some threshold value I_T. Now if the battery may vary by $\pm d_E$ per cent, and if the resistor may vary by $\pm d_R$ per cent (assuming, for simplicity, a perfect relay), it is apparent that in order to satisfy the worst-case condition, R must be selected so that



(a)



(b)

Fig. 4—(a) Relay circuit. (b) Power increases vs component tolerances.

$$R \leq \left(\frac{E}{I_T}\right)\frac{(1 - d_E)}{(1 + d_R)}. \tag{9}$$

For maximum circuit efficiency, R is selected so that its *nominal* value is just equal to the right-hand side of inequality (9). This insures that the relay will operate for the minimum expected battery voltage, and for the maximum expected resistance. Hence, if all circuit components (E and R) are at *nominal* values, the steady-state power dissipation of the relay circuit is

$$P_N = EI_T\frac{(1 + d_R)}{(1 - d_E)}. \tag{10}$$

But since the battery voltage may also increase while the resistor may simultaneously decrease, the circuit must be designed to operate with a possible maximum power expenditure given by

$$P_{\max} = EI_T\frac{(1 + d_E)^2(1 + d_R)}{(1 - d_E)(1 - d_R)}. \tag{11}$$

Eqs. (10) and (11) are illustrated in Fig. 4(b), where normalized power (P/EI_T) is plotted as a function of the design tolerances for the assumption that $d_E = d_R = d$. Thus, it is seen that the circuit of Fig. 4(a) will dissipate 50 per cent more power, at *nominal* component values, if the circuit is designed to accommodate ± 20 per cent component variations, than if ideal components (zero tolerance) are used; even more distressing is the fact that, allowing ± 20 per cent component variations, the maximum worst-case power dissipation of the circuit could be as high as 270 per cent above the case for ideal component values, and, consequently, the circuit must be designed to function at this *extreme* in power dissipation. Hence, the burden on the power supply increases very rapidly as increased component tolerances are demanded in the circuit design.

In the simple circuit of Fig. 4(a), an increase in power level because of an increase in design tolerances did not increase the component complexity of the circuit, because of the implicit assumption that the resistor R could be obtained in any power rating. Were the resistor limited to some comparatively moderate finite power dissipation, several would have had to be used in parallel to accommodate the higher power levels of the circuit. It is generally true, that, by demanding higher component tolerances in a given circuit application, a price will be paid not only in increased power dissipation, but also in an increase in the number of components required to perform the specified operation.

Consider, for example, the design of a logic driving flip-flop. The circuit is illustrated in Fig. 5. It is seen that the flip-flop must be capable of driving simultaneously an OR load and an AND load, and that, further, it must be centerpoint triggered through a pulse transmission gate. Thus, the circuit is typical of the kind used in shift registers or counters, where the stages are loaded by diode matrices or by arbitrary diode logic networks.

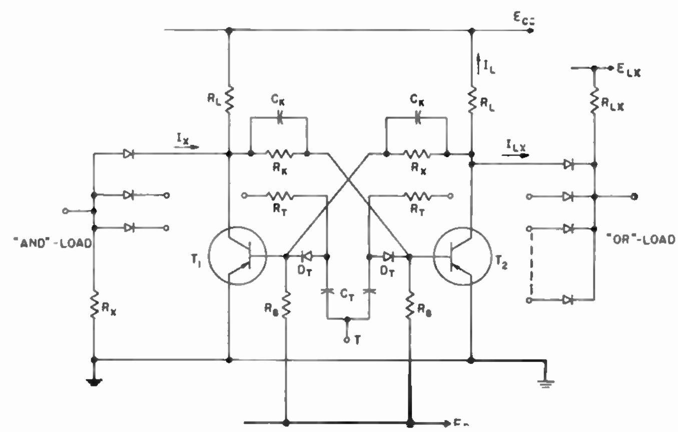


Fig. 5—Logic driving flip-flop.

It is beyond the scope of the present discussion to derive the design procedure for the circuit of Fig. 5.[3,4] However, it is of interest to illustrate how circuit power dissipation and number of components vary as a function of the component tolerances for a specific circuit design. Fig. 6 shows the standby power dissipation (power dissipation of the circuit, but not including power into the load) of the logic driving flip-flop as a function of the resistor tolerances for various values of transistor

[3] V. Mathis, H. Raillard and J. Suran, "Comparative performance of saturating and current-clamped high frequency pulse circuits," *Semiconductor Products*, vol. 3, pp. 35–40; February, 1960.

[4] R. F. Shea, *et al.*, "Transistor Circuit Engineering," John Wiley and Sons, Inc., New York, N. Y.; 1957.

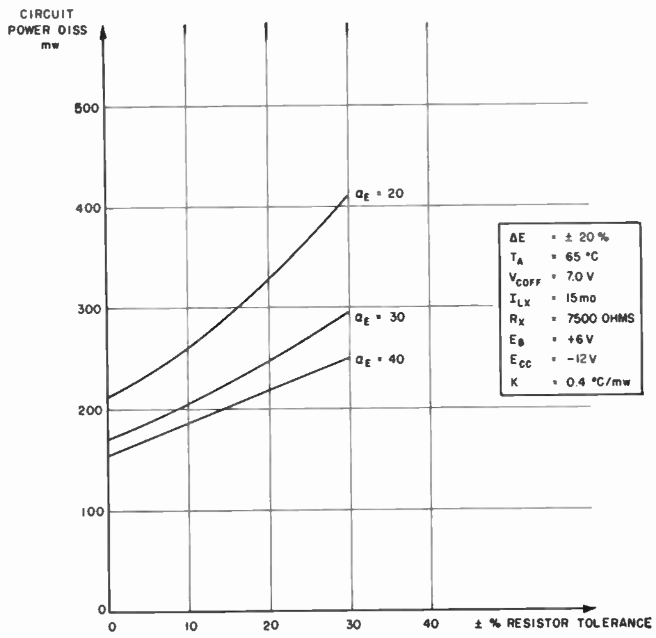


Fig. 6—"Standby" power vs resistor tolerances.

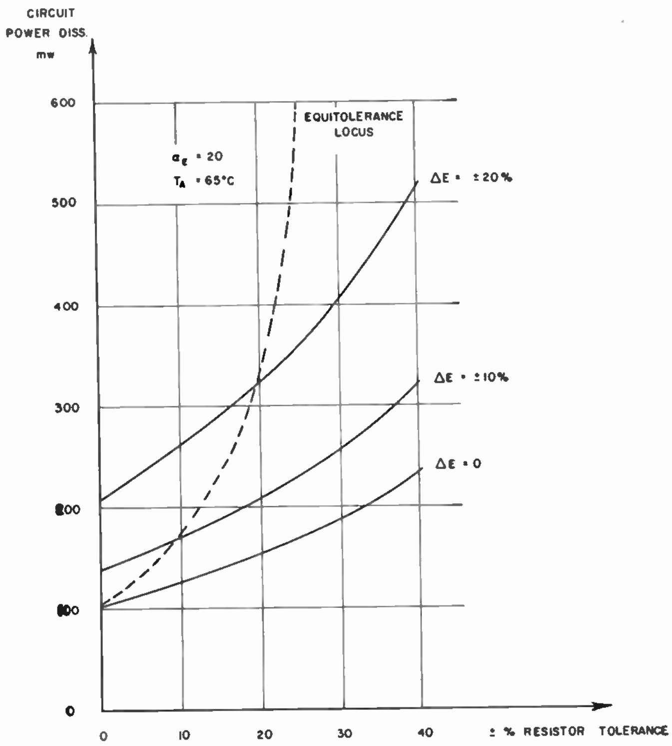


Fig. 7—"Standby" power vs resistor tolerances.

current amplification factors. The specific design values are also shown in Fig. 6, *e.g.*, a maximum ambient temperature of 65°C, an OR load of 15 ma, an AND load of 7 volts into 7500 ohms, etc. Fig. 7 illustrates the relationship between circuit power dissipation and resistor tolerances for different power supply tolerances. The dashed line in Fig. 7, referred to as the "equitolerance locus," is the circuit power dissipation as a function of equal resistor *and* supply voltage tolerances; thus, the circuit dissipates 100 mw for ideal resistors and supplies, but if worst-case deviations of 12 per cent are assumed for both the resistors and supply voltages, the circuit power dissipation is doubled.

As the power level of the flip-flop increases because of increasing parameter tolerances, the energy required to *trigger* the flip-flop also increases. This is illustrated in Fig. 8, where trigger charge (in coulombs) is plotted as a function of resistor tolerances for the circuit example discussed above. Again, referring to the equitolerance locus, it is seen from Fig. 8 that the amount of charge required to trigger the flip-flop at a resistor and supply voltage tolerance level of ±15 per cent is twice what would be required if the tolerances could be held to zero. Thus, an increased energy price is paid both statically and dynamically for increasing the tolerance values in the circuit. Since pulse driver circuits used to trigger flip-flops (*e.g.*, monostable multivibrators or blocking oscillators) are energy-limited by the power dissipation ratings of the transistors, it is apparent that in a large system the number of drivers required may go up considerably, in addition to the power dissipation, as the component tolerances are increased. Consequently, an increasing expenditure in both system complexity and power dissipation is encountered as circuits are required to work with components having wider tolerance margins.

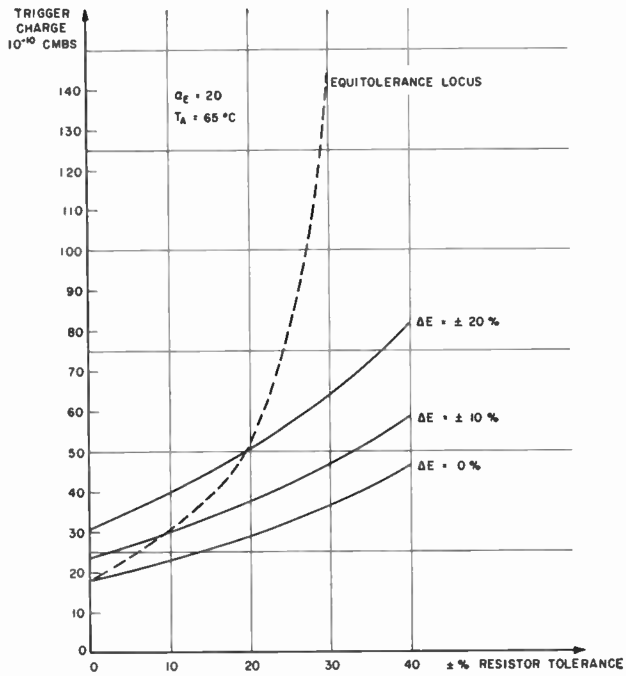


Fig. 8—Maximum trigger charge vs resistor tolerances.

Power Dissipation and Packing Density

Circuit power dissipation may be related to packing density, *viz.*, number of components per unit volume, by solving standard heat-flow equations pertaining to any given structural configuration. For example, if it is assumed that each circuit acts as a heat source of spherical shape, with radius r, that the thermal dissipation is uniform within the sphere, that the sphere is small compared to the total package volume, and that the total package is a sphere of radius R, the following steady-state equation may be used to compute the temperature difference T_D of the hottest point within the package, and the surface of the package:[5]

$$T_D = \frac{q}{8\pi KR}\,(n + \sqrt[3]{n/a}). \qquad (12)$$

In (12), n is the number of circuit elements in the package, q is the heat rate per element, K is the thermal conductivity of the material, and a is the volume utilization given by

$$a = \frac{nr^3}{R^3}. \qquad (13)$$

Curves relating circuit power dissipation and circuit packing density for a one-inch cube of germanium, as obtained from (12) for allowable temperature differences of 20°C, 50°C and 100°C, are illustrated in Fig. 9, where it is assumed that the cube is surrounded on all surfaces by an *infinite heat sink.* Thus, if an internal temperature rise of 20°C is allowable, and if each circuit dissipates 100 mw, a packing density of 7500 *circuits* within the cubic-inch volume may be achieved. If each of the circuits is of the order of complexity of a flip-flop, *e.g.*, approximately 20 components, a packing density of 150,000 components appears to be feasible from a heat-dissipation point of view. However, it should be noted that an *infinite* heat sink has been assumed at the surfaces of the cube, and, needless to say, such a thermal device would be difficult to achieve in practice. Furthermore, if the heat sink could be approximated, its size should be included in the packing density estimate.

A more realistic assessment of the packing density problem is obtained if it is assumed that heat transfer at the package surface takes place by natural convection cooling in air. The heat rate per unit area of surface, q', is then given by

$$q' = h_c(T_S - T_A), \qquad (14)$$

where T_S is the surface temperature, T_A the ambient temperature and h_c a natural convection coefficient which is a function of $(T_S - T_A)$ and of the package size. Curves showing the relationship between circuit power dissipation and circuit packing density for a one-inch cube of germanium in air are illustrated in Fig. 10. The reduction in packing density for this case, relative to the infinite heat sink calculations, is drastic. Taking the previous example for illustration, the packing density is reduced from 7500 to 4.5 circuits which, for

[5] W. Rogowski, "Ergänzung der Erwärmungsvorschriften," *Arch. Elecktrotech.*, vol. 7, p. 41; 1918.

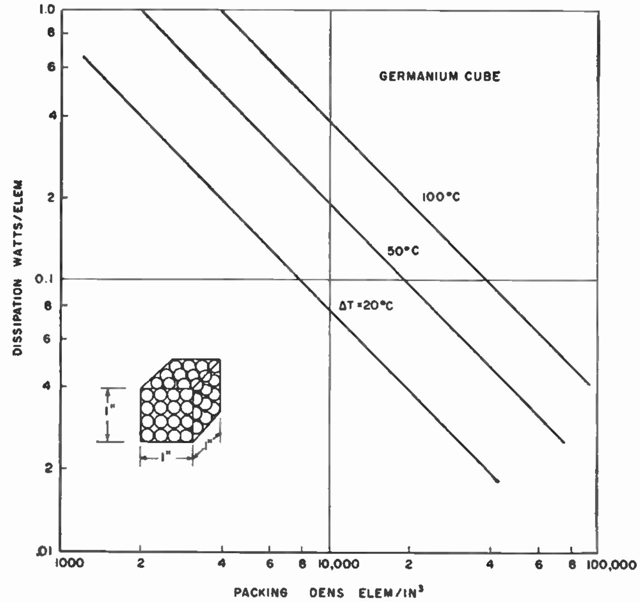


Fig. 9—Dissipation vs packing, density, infinite heat sink.

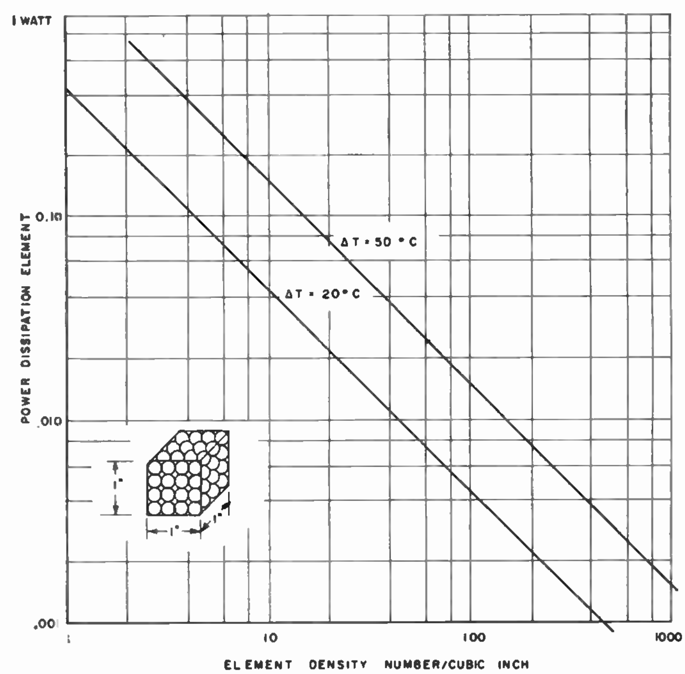


Fig. 10—Packing density vs dissipation for natural convection cooling.

the flip-flop case, corresponds to a reduction in *component* packing density from 150,000 components per cubic inch to 90 components per cubic inch. Furthermore, it can be shown that if cooling fins are mounted on the structure in order to increase the effective surface area, the allowable packing *density* will be decreased, since the air volume around the fins must be included in the calculation of total package volume. It should be noted too, that the packing density can be increased by use of forced air instead of simple convection cooling, but in forced air environments the size and weight of the fan equipment should be considered in the density calculations. Under these circumstances, nothing is gained by forced air cooling.

The packing densities shown in Figs. 9 and 10 are idealized approximations which, for practical purposes, must be significantly reduced to take into account such additional factors as nonuniform heat distribution at the package surface, thermal conductivity inhomogeneities of the material within the cube, and transient thermal effects. Consequently, reduction factors of at least two and as much as ten may have to be applied to the calculated packing densities. In addition, it should be noted that the calculations are based upon a one-inch cube, and that, should the size of the cube be increased, the allowable packing density decreases, because of the additional distance interposed between the innermost circuit and the package surface. For example, if a two-inch cube is assumed, the packing density for a given circuit power dissipation and allowable temperature differential is decreased by a factor of approximately two, relative to the case of a one-inch cube.

Conclusions

It has been shown that circuit power dissipation for any given class of devices is determined by the *functional* gain-bandwidth of the circuit, and by the component tolerances around which the circuit must be designed. Higher operating speeds entail higher circuit power dissipations. Similarly, greater component tolerances require greater power dissipation for circuit stabilization. Since power dissipation is associated with heat generation, thermodynamic considerations determine the practical limits of circuit packing densities in microelectronic equipment.

Conventional solid-state circuits, averaging 50 mw of power and 10 passive components per active device, for example, are at least one to two orders of magnitude too high in power level to achieve packing densities compatible with numbers usually quoted (*e.g.*, 10,000 components per cubic inch) as packaging goals of microelectronics. Attempts to reduce circuit power levels, employing state-of-the-art circuits with state-of-the-art components, encounter fundamental problems associated with circuit-driving capability (gain), circuit speed (or bandwidth) and component tolerances.

Microcircuit fabrication techniques directed to the production of entire circuits or "systems" in "monolithic block" structures must face up to tolerance and power problems on a packing density scale so severe that the problems appear to be beyond solution with components made individually by the most sophisticated present-day technologies. Consequently, it appears reasonable to expect that practical microelectronic equipment, at least within the foreseeable future, will take the form of hybrid structures employing both individually fabricated components, which may be of a basic, complex or integrated nature, and single-process component arrays, all mounted or processed onto a suitable insulating substrate. Determination of which components must be made individually, and which can be fabricated in single-step array form will be made on the basis of cost as well as tolerance control. Furthermore, it is almost certain that microstructures will encompass all known solid-state technologies, drawing not only on semiconductor materials, but also on dielectric, magnetic, electroluminescent, photoconductive, magnetoresistive, cryogenic and other sources which may provide the most economical, efficient and reliable means for realization of a given electrical function.

Considerable progress in the design and fabrication of active, as well as passive components will be required before packing densities of the order of 10,000 components per cubic inch can be realized. Active devices will be required to operate at very low power levels, while at the same time providing high driving capability (gain), high speed (gain-bandwidth product), low reactance, low or medium resistance levels (for high speed and relative noise immunity), temperature insensitivity and directivity (or unilaterality). In addition, such devices will have to be reproducible to tight tolerances, mechanically and chemically rugged, low in cost and compatible with other microcomponents. Although devices having all these properties are nonexistent, considerable encouragement can be taken from the fact that some come fairly close, *e.g.*, tunnel diodes and cryotrons.

Finally, it should be pointed out that the problems which must be solved in microelectronics are not only material or device problems. Quite the contrary, the ultimate problem will have to be solved on a system level where such considerations as over-all reliability, compatibility, and adjustability are encountered. It is rapidly becoming apparent that the success of microelectronics will not only revolutionize the appearance of equipment, but, even more profoundly, will completely change equipment design philosophy. The traditional separation of component, circuit and system design must be expected to yield to a system-oriented design approach operating within constraints set by the state of material chemistry and physics. The "building block" concept of system design will probably be inadequate to realize fully the possibilities of microelectronics, while conventional standardization and "preferred circuit" concepts may be abandoned. In summary, it can be predicted that the success of microelectronics will be predicated upon the design of microminiature systems rather than upon the microminiaturization of conventional system designs.

Acknowledgment

The assistance of Dr. I. W. Wolf in providing solutions to the heat-flow problem is gratefully acknowledged. In addition, stimulating discussions with several colleagues, particularly A. P. Stern, I. A. Lesk and N. Schwartz, helped considerably in crystallizing some of the points of view described in this paper.

Some Technical Aspects of Microwave Radiation Hazards*

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Summary—Man's ability to generate microwave power has been increasing at the rate of about 15 db per decade. Experiments performed by subjecting animals to high power indicate that hazards to personnel could exist if appropriate safety measures are not adopted and observed.

This paper reviews the history of the recognition of this potential hazard and the safety measures adopted by the Bell System and others to protect personnel.

Some typical and pertinent research work is discussed, and it is shown how these results have influenced the establishment of criteria for safe and potentially hazardous environments for human beings.

The currently adopted safety limits of the Bell System and others are reviewed in some detail, and a recommended method of calculating power densities is derived, pointing out the limitations of the approximations used.

Some of the commercially available power density meters are mentioned, and their method of operation is described. Their use in surveying a site is discussed, and the shielding effect of wire mesh fences is presented in a nomograph.

Definition of Microwave Radiation

THE microwave region is considered to extend from the highest radio frequencies down to the ultra-high frequency band between 300 and 3000 Mc, but radiation hazards may exist at any radio frequency capable of being absorbed by the body. Referring to the electromagnetic spectrum chart of Fig. 1, which shows the wavelength from $(10)^7$ to $(10)^{-12}$ cm, corresponding to frequencies from 3000 cps to 3×10^{16} cps, the four main classifications of radiation are indicated at the top.

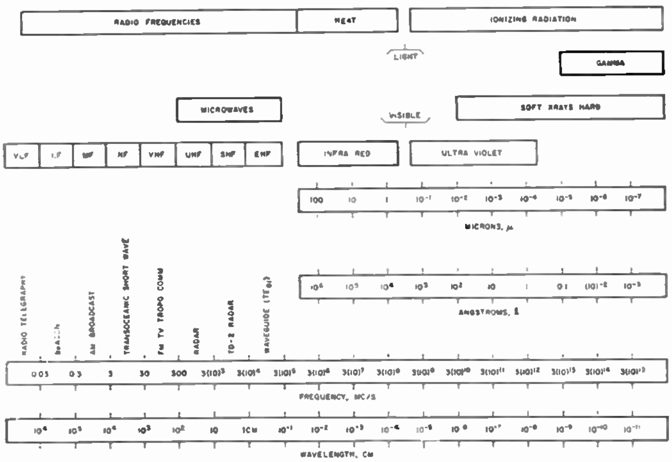

Fig. 1—The electromagnetic spectrum.

These are:

1) Radio frequency waves from 10 kc to $(10)^{12}$ cps,
2) Heat waves or infrared rays from $(10)^{12}$ cps to about $4(10)^{14}$ cps, the latter corresponding to a wavelength of about $0.72\ (10)^{-6}$ meters (0.72 micron),
3) The visible spectrum from a wavelength of 0.72 micron (7200 A) to about 3800 A,
4) Ionizing radiation (such as ultra-violet, X-ray and gamma radiation) at wavelengths less than about 3800 A.

The radio frequencies include eight frequency regions corresponding to the eight decades of wavelength they occupy. These eight bands have been called:

1) Very-low frequencies $(10)^7$ to $(10)^6$ cm
2) Low frequencies $(10)^6$ to $(10)^5$ cm
3) Medium frequencies $(10)^5$ to $(10)^4$ cm
4) High frequencies $(10)^4$ to $(10)^3$ cm
5) Very-high frequencies $(10)^3$ to $(10)^2$ cm
6) Ultra-high frequencies $(10)^2$ to (10) cm
7) Super-high frequencies 10 cm to 1 cm
8) Extra-high frequencies 1 cm to $(10)^{-1}$ cm.

Some of the uses of this portion of the electromagnetic spectrum include radio telegraphy, radio broadcasting, radio communication, radar, and millimeter wave waveguide transmission systems. Diathermy, radio frequency furnaces, and other industrial apparatus based upon the heating effect also use this portion of the spectrum. This heating effect can be hazardous.

Some Uses of Microwave Radiation

As indicated on the chart, the "microwave" band includes the UHF, SHF and EHF bands, and the chief uses of microwaves are radar, tropospheric scatter propagation, and relay links. Satellite communication links will use microwaves too, and these systems will probably be more powerful than the present radar and scatter propagation transmitters.

Some typical microwave antennas are shown in Figs. 2–5.

Fig. 2 shows U. S. Army field site with its associated acquisition and tracking radars. The high power acquisition radars are normally used only when the antenna is scanning, and, hence, the average power absorbed by an object at a fixed point is reduced by the fact that the direct beam is pointed in that direction only a fraction of the time it takes for the antenna to

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complete one whole revolution. Some of these radars lay down such a strong field that if they were not rotating, the power density might be hazardous to a distance of 500 feet or more. In such cases, interlocks may be used to ensure that the transmitter be idle if the antenna does not scan.

The tracking radars usually do not scan, but rather point toward the target or the missile wherever it may be. Hence, they are potentially more hazardous than the acquisition radars even though their power may be less. Safety considerations may dictate that interlocks be provided to prevent a target tracking radar from pointing in certain critical directions.

Fig. 3 shows a TD-2 relay link installation, which is such a familiar sight across the continent. In this system, the total radiated power is so low anywhere in the beam of the antenna that conditions are safe.

Fig. 4 shows the antennas of a White Alice installation. Here the antennas are fixed in space, and the power radiated is high enough so that the power density reaches a potentially hazardous value. The men in the foreground are posing with a portable field strength measuring device which was used to determine where safety fences should be erected. Tests were made at reduced power to be safe.

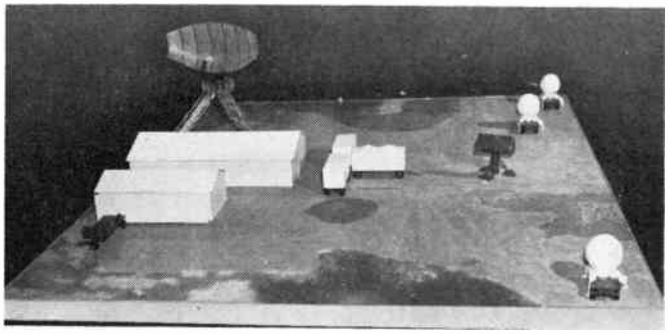

Fig. 2—A model of an Army field site.

Fig. 3—A TD-2 microwave relay link tower (courtesy of Mountain States Telephone and Telegraph Company).

Fig. 5 shows an antenna for a tropospheric scatter propagation communication link. Two of these links are now in operation in the Bell System, and more are being planned. The power of this transmitter is high enough to warrant a field strength survey in front of and around the antenna, and here again two men are shown with the simple laboratory model of a field intensity meter. In this system calculations indicated that a protective fence was needed. After the fence was erected, measurements (at half power) indicated that the power density outside the fence would be less than 1/10 mw/cm² at full power, a value which is 1/10 that which is considered safe for indefinite exposure.

Fig. 4—Antennas for White Alice.

Fig. 5—Tropospheric scatter antennas

Trend of Power Available

Some of the modern equipments are potentially hazardous, but with the trend toward higher and higher powers, our awareness now of the hazards of the future becomes increasingly important. Let us see where we stand today regarding the amounts of microwave power which man has been capable of producing at 3000 Mc, for example. Fig. 6 shows this trend over the past couple of decades, and the dashed line extending into the future represents one estimate. In 1940, the famous British magnetron [14] was capable of delivering 10 kw of peak-pulsed power into a properly matched antenna. This would be an average power of 10 watts if the duty cycle were one in a thousand.[1] By about 1945 [14], the improvements made on this type of magnetron resulted in the capability of delivering 1100 kw of peak-pulsed power, or, with the same duty cycle, 1.1 kw of average power. After World War II, there was, naturally, a slowdown in the advance of this art, but by 1957 [43] a klystron was available which was capable of delivering 8.0 kw of average power. At the present time, it appears that the development of higher powered devices has again been stimulated so that by 1965 [58] at least 1000 kw of average power has been predicted. This trend from 10 watts in 1940 to 1000 kw in 1965 represents a rising capability of 50 db in less than 30 years, or a rate of over 15 db per decade.

Thus, it appears appropriate to examine the situation regarding microwave radiation hazards today, lest we be faced with an intolerable situation in the future.

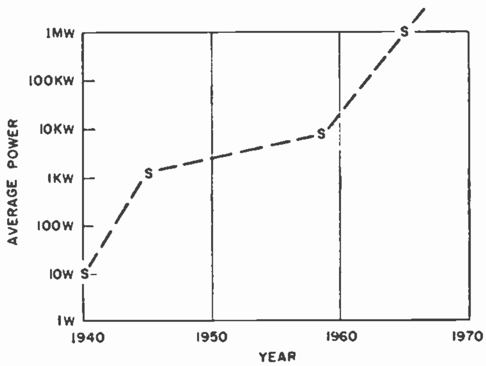

Fig. 6—Trend of power available at 3000 Mc.

Effects of Microwave Radiation

Most of the experimental work to date supports the belief that the chief effect of microwave energy on living tissue is to produce heating. Consequently, exposure to microwave radiation should probably represent no hazard unless overheating is a possibility. Within carefully prescribed limits, the heating effect of radio waves may actually be beneficial; in fact, this is the basis of diathermy, which has long been employed therapeutically. The use of diathermy is so widespread and so well accepted a part of medical procedure that the FCC has assigned seven frequencies, three in the HF or VHF[2] and four in the microwave region,[3] for the operation of medical diathermy equipment.

Heating is a function of the strength of the microwave field, that is, the average power flow per unit area (usually expressed in milliwatts per square centimeter). It is also a function of time. The heating may take place near the surface or deep within the body, the depth of penetration being related to frequency. Frequencies in the region 200 to 900 Mc penetrate deeply, whereas *S*-band (1500–5200 Mc) and *X*-band (5200–11,000 Mc) frequencies used by radars produce heating at or near the surface.

Heating effects, depending on frequency, are 1) a general rise in body temperature, similar to fever, or 2) something more localized, akin to the cooking process in a radar oven, where a steak can be cooked from the inside out. The human body can compensate for a certain amount of heating of the first type through perspiration, if the temperature rise is not too sudden. Consequently, the hazard may be somewhat less in cool weather than on an extremely hot day when the body's cooling mechanism is already working at full capacity. Compensating mechanisms for coping with the second type of heating are less adequate.

Circulating blood acts as a coolant, so that localized heating is least serious in parts such as muscle tissue, which are well equipped with blood vessels. Heating is more of a danger to the brain, the testes and the hollow viscera. The most widespread publicity has related to the effect of microwave radiation on the eyes. The viscous material within the eyeball is affected by heat in much the same manner as the white of an egg, which is transparent at room temperature but becomes opaque white when warmed slightly. In the eyes, as in the egg white, the process is irreversible.

As the surface of the human body is more generously supplied with sensory nerves than the interior, a feeling of warmth may give a warning in case of over-exposure of frequencies which produce surface heating. If the frequencies are such as to cause a general rise in body temperature, the resulting sensation of discomfort may or may not be perceived in time to provide adequate warning. In the case of localized microwave heating deep within the body, it is still less likely that any warning sensations would be noted before damage was done. Hence, it is important to establish limits and to delineate the areas in which a potential health hazard could exist.

[1] Duty cycle is the fraction of the time that the radar is transmitting.

[2] 13.56 Mc
27.12
40.68

[3] 915 Mc
2,450
5,850
18,000

Review of Meetings, Data and Recommendations

Let us review some of the typical, significant and pertinent written publications and conferences on biological effects of microwaves and see how these data and their interpretation have influenced our thinking regarding the adoption of exposure limits and safety regulations. This examination had best be done chronologically in order to interpret the recommendations in their proper perspective.

Much of the material which is about to be presented is published in writing in the proceedings of conferences which have been held recently (*i.e.*, since 1955). Three well documented conferences include:

1) "Symposium on Physiologic and Pathologic Effects of Microwaves," held at the Mayo Foundation House, Rochester, Minn., September 23 and 24, 1955. Holding this meeting at the Mayo Clinic was suggested by Dr. F. G. Hirsch, who was then medical director of Sandia Corporation, Sandia Base, Albuquerque, N. M. It was a particularly appropriate meeting place because microwave diathermy was introduced to the medical profession by the Mayo Clinic. The papers presented at this symposium were subsequently published in the IRE Transactions on Medical Electronics (vol. ME-4; February, 1956). The attendance at this meeting represented the medical profession, governmental agencies, institutions of higher learning, and industry. The data presented yielded sufficient evidence for establishing tolerance levels for experimental animals as a function of exposure time; however, the extrapolation to human beings was still subject to considerable uncertainty. The consensus at that time and the seemingly proper conservative approach was to assume that human beings were just as susceptible as other animals.
2) "The First Annual Tri-Service Conference on Biological Hazards of Microwave Radiation" held at Griffiss Air Force Base, Rome, N. Y., on July 15–16, 1957. This meeting was sponsored by the Air Research and Development Command, Headquarters. It was held "to assemble the A.R.D.C. advisory panel, tri-service representatives and various contractors working in the radiation hazard area to effect an understanding of activities and accomplishments to date." Several interested industrial representatives were also invited. The work to date was described and the program for the future discussed [46], [47]. It was planned to hold periodic meetings to maintain close liaison among the investigators in the field, to expand research programs to include a sampling of frequencies from 200 through 35,000 Mc, to investigate the effects of microwaves combined with X rays, to study methods of microwave field measurement, and to study testicular and ocular damage as a function of frequency and cumulative dosages. The very capable and energetic host was Colonel G. M. Knauf, who presided at the sessions and contributed to the success of the meeting by delivering two papers in addition to his appropriate commentaries and introductions. The Proceedings of this conference were compiled and edited by Dr. E. G. Pattishall, University of Virginia, under project No. 57-13, contract AF 18(600)-1180 with the George Washington University, Washington, D. C.
3) "The Second Annual Tri-Service Conference on Biological Effects of Microwave Energy," held at Griffiss Air Force Base, Rome, N. Y., on July 8, 9, and 10, 1958, sponsored by the Air Research and Development Command, Headquarters, to "bring together key researchers in the bioeffects area so that each department within the Armed Services could discuss on-going and needed research." This meeting was patterned after the first Tri-Service meeting and the Proceedings were compiled and edited by Pattishall and Banghart, Project Directors, University of Virginia, Charlottesville, Va., under contract AF 18(600)-1792, Division of Educational Research, University of Virginia. This document is identified as ARDC-TR-58-54 and ASTIA Document No. AD 131477. Colonel Knauf was again the genial host and chairman in addition to presenting three significant papers.

The three documents which record the presentations at these three meetings contain several comprehensive bibliographies on the subject [41], [42]. Only a few selected references will be quoted here. Each reference was chosen either because it presented a good summary, or represented typical data, or because it appeared to mark a real "milestone" in the advance of knowledge.

Probably the first demonstration of the heating effect of radio frequency energy was that of d'Arsonval in 1890, according to Krusen [30], who gave the address of welcome at the Mayo meeting. He (d'Arsonval) "demstrated that 'high-frequency' electric currents of 10,000 cycles per second produced no muscular contraction in the human being, but causes only heating. Since then physicians have been employing increasingly higher frequencies for heating of living tissues. By 1900, high-frequency currents of 1,000,000 to 3,000,000 cycles per second (long wave diathermy), were in use. . . ."

Krusen continued, " . . . and by 1935, electric currents of still higher frequency, 10,000,000 cycles per second at a wavelength of 30 meters (short-wave diathermy), were being employed." One type of hazard of these machines arose from the fact that the original models were inadequately shielded and improperly operated on communication channels, causing disruption of our overseas short wave circuits.

Brody [28], Medical Liaison Officer, Bureau of Aeronautics, Navy Department, Washington, D. C., speaking at the Mayo meeting, summarized the work conducted between 1935 and 1950 thus: "One heard tales concerning the sterilizing effects of radar beams or other equally serious manifestations. . . . To determine whether or not these suppositions might be based on factual grounds, a clinical study was conducted in 1943 by Daily [6] and in 1945, by Lidman and Cohn [7]. Certainly (in view of our present knowledge), the lack of clinical changes reported resulting from exposures to radar emanations may be attributable to the low power outputs then available. Meanwhile, basic research utilizing animals continued, notably by Clark, Hines, Salisbury and Randall [12], [15]. . . . These studies amply demonstrated the insidious effects of high-power microwaves on certain tissues and pointed out the most significant changes with the use of discrete frequencies."

In 1949, Clark *et al.* [12] at Collins Radio Company, supported by RAND Corporation and the U. S. Air Force developed cataracts in the eyes of rabbits upon exposing them to 3000 mw/cm² for ten minutes at a frequency of 2500 Mc. This corroborated the earlier work of Richardson, Duane and Hines [10] who, in 1947, reported lenticular opacities appeared at about 50°C with intentional overdoses. There still remained, however, the task of gathering enough data to establish a reasonable limit. Meanwhile, in 1948, Imig, Thomson and Hines [11] reported that "testicular degeneration" may occur from microwave heating at a lower temperature than from infrared heating.

Olendorf [13] reported, in 1949, that he had used 2400 Mc radiation to produce lesions in the brains of rabbits.

Thus it was established more than ten years ago that microwaves could be hazardous.

In 1952, Hirsch [19], [23] of Sandia Corporation reported with Parker that lenticular opacities occurred in the eyes of a technician operating a microwave generator. Their report [23] states ". . . In the fall of 1950 he [the technician] set up a microwave test bench and used it from November, 1950, to October, 1951. His equipment included an experimental microwave generator tunable from about 9 to 18 cm wavelength (1660 to 3320 megacycles per second) and having an average power output of 100 watts on a 50 per cent duty cycle. His test line was terminated by a horn antenna which dissipated the power into a room. . . . It has been calculated that the intensity of radiation at a plane coinciding with the rim of the dissipating antenna was about 100 milliwatts per square centimeter.

"A habitual practice of this operator is worthy of mention, since it may well have some bearing on the case. In order to determine whether or not the equipment was generating energy, he made a regular practice of placing his hand in the dissipating antenna . . . and noting the heating effect on his hand. In these circumstances it was necessary for him to look into the antenna in order to place his hand properly. . . .

"On October 11, 1951 he presented himself to one of us [Hirsch] with the chief complaint of an inability to see clearly. . . . This visual disturbance had developed, he said, over a period of about a week or 10 days. He was not aware of any loss of visual acuity prior to that time."

Hirsch continued, as a result of examination, "A diagnosis of bilateral nuclear cataracts with acute chorioretinitis was made" and commented that, "it will be well, therefore, to use this case as a means of recalling the attention of ophthalmologists, industrial physicians, and microwave operators to the potentialities of microwave radiations in order that the use of this form of energy will be accompanied by appropriate respect and precautions."

Hirsch's well documented report attracted widespread attention both in industry and in the Armed Services. On the basis of Hirsch's observation that an estimated 100 mw/cm² was hazardous, organizations began adopting rules and regulations concerning microwave hazards. Some rules specified a hazardous level and others specified a safe level. Naturally, because of the paucity of adequate quantitative data, there was a large spread between the specified hazardous levels and the specified safe levels at that time.

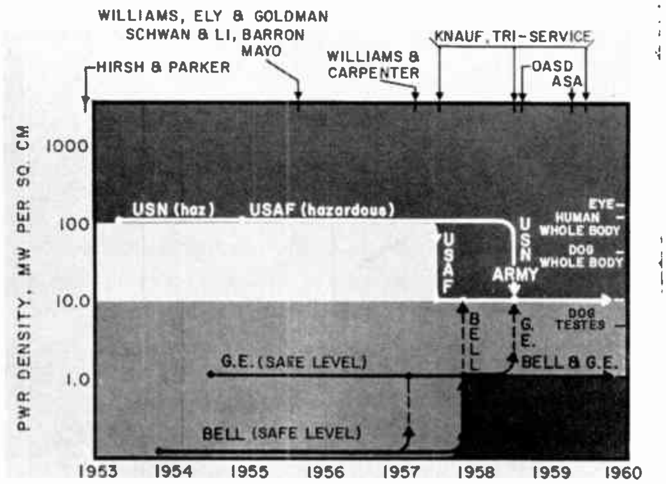

Fig. 7—Recommendations for safe and hazardous levels.

Fig. 7 shows graphically how the various recommendations have progressed from 1953 to date. Power density in mw/cm² is plotted logarithmically on the ordinate, with time in years linearly on the abscissa. At the extreme right the power density levels are indicated at which it has been reported that various effects occur. For example, reports [33], [36], indicated that 155 mw/cm² was the threshold for cataract formation by single exposures to the eye; 100 mw/cm² was estimated to be tolerable under favorable conditions for total immersion of the human body (except for the testes), 40 mw/cm² was lethal to dogs [29], [36], and 5 mw/cm² was considered to be the maximum exposure for no observable change in the testes of dogs [36]. These data will be discussed in more detail later.

With these reported critical levels in mind, let us turn our attention then to the safe and hazardous levels adopted by various organizations through the years following Hirsch's report, and see how the conferences and reports affected these recommendations.

Brody [28] reported at the Mayo meeting that "A density level based on the figure 100 mw/cm² was agreed upon as a damage risk criterion by those attending a symposium at the Naval Medical Research Institute in April, 1953."

The Central Safety Committee of the Bell Telephone Laboratories reviewed the then published material, and on November 16, 1953, one department of the Laboratories issued a bulletin based upon the Safety Committee considerations. Excerpts from this bulletin are quoted. In view of the fact that "about all that has been shown at present is that, at frequencies used in radar and microwave communication, wave fronts impinging on the eye with intensities corresponding to an energy flow of 100 milliwatts per square centimeter may cause injury . . . [therefore] it seems likely that 0.1 milliwatt would be safe." At that time it was believed that a safety factor of 30 db was incorporated in this recommendation. We shall see that subsequent research work reported on dog testes would indicate a safety factor of considerably less than 30 db.

The General Electric Company, Schenectady, N. Y., too, was early in establishing a safe level. Vosburgh [32], consultant, G.E. Health Services, submitted at the Mayo meeting the recorded minutes of a meeting held on June 1, 1954, at Schenectady dealing with "Microwave Radiation; Hazards and Safety Measures" in which recommended safety measures were established. These recommended measures included, among other excellent suggestions, "Appropriate procedures shall be applied to limit the direct and *reflected* intensity to 1 milliwatt per square cm average in all locations to which people require access: a) by the use of shielding or absorbing material, or b) by remote viewing."

The U. S. Air Force was also apparently aware of the potential hazards of microwaves, at least as early as 1954. Major D. B. Williams and Colonel R. S. Fixott [49], in presenting a summary of the SAMUSAF program for research on biomedical aspects of microwave radiation at the first Tri-Service Conference in 1957, stated " . . . only three years ago [1954] the sole data available on microwave tolerance was that an estimated 3000 mw/cm² should be regarded as hazardous for personnel exposure." Apparently, the Hirsch report (1952) had been overlooked.

By November 1, 1954 the U. S. Air Force [24] recognized that " . . . continuous exposure to microwave radiation intensities of 100 milliwatts/cm² can cause damage to human tissues, particularly the eyes." The Hirsch report was apparently accepted.

Colonel Knauf of the U. S. Air Force, [40], [52] speaking at the 106th annual meeting of the American Medical Association in New York, N. Y., on June 7, 1957, said, "About two years ago [1955] . . . a power density of approximately 200 milliwatts per square centimeter seemed to be the point at which a division between effect and no effect could be predicted." Apparently the Hirsch report was being discounted. He continued, "Hence, we published a maximum safe exposure level of 10 milliwatts per square centimeter."

In the meantime, the Mayo meeting had been held September 23 and 24, 1955. It is evident that the excellent papers presented at this meeting had quite an impact on the thinking concerning safe and hazardous levels of microwave radiation. The solid work of Williams *et al.* [33] of the U. S. Air Force in collaboration with Monahan of St. Johns Hospital, London, Eng., on time and power thresholds for the production of opacities was a big contribution. Schwan and Li [31] concluded that 10 mw/cm² should not be applied for more than one hour if total body absorption of radiation is assumed.

While many of the Mayo papers dealt with the production of opacities at levels in the neighborhood of 100 mw and up, one notable contribution, that of Ely and Goldman *et al.* [29] of the Naval Medical Research Institute, National Medical Center, Bethesda, Md., reported *death of animals at less than 50* mw/cm² at a wavelength of 10 cm. In this work they had established that only about 40 per cent of the incident energy had been absorbed. If all the energy had been absorbed, the lethal dose would have been about 20 mw/cm². The estimate of 40 per cent power absorbed seems a reasonable one in light of the work reported at Mayo by Schwan and Li [31] of the Electromedical Group, Moore School of Electrical Engineering and Department of Physical Medicine and Rehabilitation, Schools of Medicine, University of Pennsylvania, Philadelphia, Pa. They had studied the absorption of microwave radiation by the human body as a function of frequency for various skin thicknesses, taking into account the propagation constants of skin, fat and muscle. At frequencies from 150 Mc to 400 Mc, the body could absorb from 30 per cent to 50 per cent of the radiation. Above 900 Mc, they reported that the percentage absorbed fluctuations between 20 and 100 per cent depending on frequency, and thickness of skin and subcutaneous fat.

Also at the Mayo meeting, Barron [26], [27] of Lockheed Aircraft Corporation, Burbank, Calif., reported no serious changes in personnel exposed to fields as high as 13 mw/cm². They had initiated a "comprehensive physical examination program of radar personnel in an effort to determine whether prolonged (years) and short (months) duration exposure to microwave emanations had resulted in transient or permanent biologic damage." Included in the program were 226 personnel with histories of radar contact varying from occasional beam exposure to 4 hours a day and up to 13 years overall. Areas near the transmitter had been zoned, and all personnel were prohibited from entering zone A, where the density exceeded 13 mw/cm², and 88 control subjects were prohibited from entering zones where the density exceeded 1.6 mw/cm². Barron reported, "the original examination showed a significant decrease in the polymorphonuclear cells in 25 per cent of the radar personnel as compared with 12 per cent in the control group. An interesting and disproportionate increase in monocytes above 6 per cent and eosinophiles over 4 per cent was also noted in the radar group." (These were not considered to be serious.) "There was no indication of increased significant pathology among the radar group. Ocular examinations revealed a high incidence of pathology for the radar group. However, in all but a single case of retinal hemorrhage, the etiology was known and determined to be entirely unrelated to radar exposure. . . .

"Re-examination was accomplished on 175 subjects following 6 to 9 months of incidental contact with both 3-cm and 10-cm pulsed radar (power densities less than 13 milliwatts per square centimeter). . . . Preliminary observations revealed a decrease of red blood cells in excess of 10 per cent from the original in 30 per cent of subjects; an increase of white blood cells in 50 per cent; and an increase in lymphocytes in 39 per cent. The blood platelets showed no significant unidirectional change." The observed changes were not considered to be serious. *No* changes in the eyes were observed.

After the Mayo meeting (1955), it was apparent that some estimates of hazardous levels were too high and that the original 1953 Bell Laboratories estimate of safe level (0.1 mw/cm²) was too conservative.

In an urgent action technical order [40] dated June 17, 1957, and also a Rome Air Development Center Regulation [39] dated May 31, 1957, there was established a "hazardous microwave radiation level of 10 milliwatts per square centimeter or greater over the entire microwave spectrum" [50].

Within the Bell Telephone Laboratories, several engineers initiated memoranda, sponsored conferences, and wrote letters to have the safe level limit revised upward. These efforts, in liaison with the engineers at the American Telephone and Telegraph Company, coordinated by Smith, led to the adoption of a potentially hazardous level of 10 mw/cm² on October 24, 1957.

The Navy, the Army, and General Electric Company also reported at the second Tri-Service meeting in July, 1958, that 10 mw/cm² was being considered the limit. Roman, consultant, Bureau of Ships, reported [65], "The Bureau of Medicine has tentatively established a working level of ten milliwatts per square centimeter as the tolerable dosage for constant exposure to microwave radiation."

Lieutenant Colonel L. C. MacMurray [63], U. S. Army Environmental Health Laboratory, reported, "The U. S. Army has adopted the tentative criteria of 10 milliwatts per square centimeter as being the upper limit of safe exposure to microwave radiation." U. S. Army Regulation No. 40-583 [74] states, "Personnel will not be permitted to work in the field of radiation . . . where the . . . power density exceeds 10 milliwatts per square centimeter."

Vosburgh reported [68] that " . . . the health and hygiene service of the G.E. Co. does recommend that the G.E. Co. strike an agreement with the Armed Services on this value . . . and . . . conclude (concur) with the net recommended ceiling tolerance level of 10 milliwatts per square centimeter."

Thus, there appears to be general agreement between all three Armed Services and at least two industrial organizations on the upper limit of 10 mw/cm².

A review of the factors which influenced the establishment of exposure limits would not be complete without mentioning the "California Incident" [37], which received widespread publicity in May, 1957, just prior to the first Tri-Service meeting in Rome. This was the report in which a worker was allegedly killed by radar. Fricker commented at the first Tri-Service meeting "Further investigation revealed that this person died of other causes."[4] Nevertheless, the widespread publicity of this incident did add stimulus to the research work on radiation hazards of microwaves.

The Tri-Service meetings have been helpful indeed in bringing us up to date with the recent advances. In 1957, the reports of Knauf [46], [47], the commentaries of Fricker [45], the analysis of Schwan [48], and the practical engineering report of Dondero [44] were especially worthwhile. At the second Tri-Service meeting in July, 1958, the contributions of Knauf [62], Schwan [66], [67], Hartman [59], Roman [65], MacMurray [63], Herrick [60], Carpenter [57], Michaelson *et al.* [64], Keplinger [61], Susskind and Jacobson [55] are notable. We shall refer to some of these later. The third annual Tri-Service Conference was held at the University of California, Berkeley, August 25, 26 and 27, 1959, and the Proceedings are available [76]. It should be noted that the data presented at this meeting substantiated the previous choice of 10 mw/cm² for the upper limit.

Other meetings have been organized to sponsor the dissemination of information. On August 15, 1958, the office of the Assistant Secretary of Defense, Research and Engineering, held a classified meeting at the Pentagon to review electromagnetic radiation effects on materiel and personnel. In attendance were representatives of the three Armed Services, the OASD, a Bureau of Ships contractor, and one man from Bell Telephone Laboratories.

This meeting paved the way for the organization of a "General Conference on Standardization in the Field of Radio-Frequency Radiation Hazards," held May 4, 1959, at the Headquarters of the American Standards Association in New York, N. Y. At this meeting, it was voted that the American Standards Association be requested to establish a project in regard to hazards aris-

[4] See [45], p. 85.

ing from radio-frequency electromagnetic radiation and that the AIEE and the Department of the Navy, Bureau of Ships, be recommended as cosponsors for the proposed project. The future recommendations of this committee will, no doubt, be very helpful to all.

Detailed Examination of Typical, Significant and Pertinent Data

Let us now look at some of the typical, significant and pertinent data in more detail. Clark [12], [15], Richardson [10], and Hirsch [19], [23] had demonstrated cataract formation prior to the work reported by Williams [33] at the Mayo Clinic, and Williams showed how his more recent results tied in satisfactorily with the results of the previous workers. Not only that, but he presented the data in such a way that one could interpret them in terms of a threshold level vs time of exposure for a 50–50 chance of cataract formation in the eyes of rabbits by single exposures of 2400 Mc radiation. Table I summarizes these data, which are plotted on Fig. 8.

TABLE I

Power-Time Threshold for Production of Opacities in Rabbit Eyes (Single Exposure)

Exposure Time	Threshold Power Density
5 min.	0.6 w/cm²
20 min.	0.4 w/cm²
90 min.	0.3 w/cm²
270 min.	<0.22 w/cm²
270 min.	>0.12 w/cm²

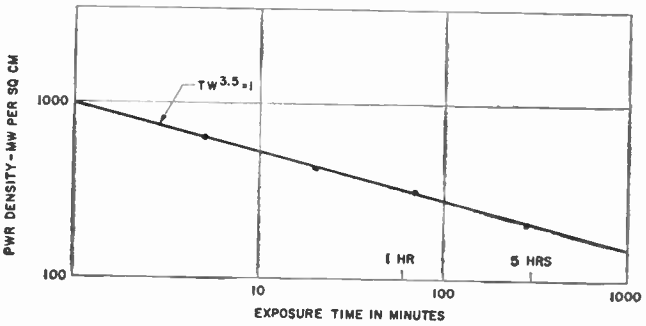

Fig. 8—Threshold for the formation of cataracts in the eyes of rabbits [33].

These observations indicated that the time of exposure for a 50–50 chance of cataract formation varied inversely as the seven-halves power of the power density. Extrapolation of this relationship to an exposure time of one minute would predict an even chance for the formation of cataracts at a power density level of about 1000 mw/cm². Extrapolation to an exposure time of one working day would give about 170 mw/cm².

Conclusions based upon extrapolation both as illustrated above and from animals to man could prove to be misleading. However, until further evidence became available, the more conservative use of the available data seemed to be warranted, and the threshold for producing cataracts in animals has been considered dangerous for human eyes.

Williams also pointed out that the multiple exposures reported by Daily [21] produced opacities at density levels well below the threshold requirements for single exposures. Specifically, ten exposures of 30 minutes each at a level of 150 mw/cm² produced opacities in two of six rabbits. This indicated a possibility that opacities might be produced as a cumulative effect. This cumulative effect was substantiated later by Carpenter [57] at Tufts University, Medford, Mass., and reported at the second Tri-Service meeting in 1958. This is of considerable importance with respect to its connotations in human exposure to low-power levels of radiation over long periods of time.

Since much of the published work dealt with the formation of cataracts, a false impression has been created that the eye is the organ most susceptible to damage by localized exposure to microwave radiation. That the testes were more susceptible than the eye was reported by Ely, Goldman, Hearon, Williams, and Carpenter [36] of the Naval Medical Research Institute in 1957. This may not have been of as much concern since the scrotum is sensitive to the sensation of heat, whereas the eye is not, and the subject would have the warmth as a warning in the scrotum but not in the eye. However, in the exposure of the eye to a fairly uniform wavefront, the sensation of heat in the skin near the eye should give a similar warning. This was observed by Daily *et al.* [21] who reported that "human [eye] exposures of 240 milliwatts per square centimeter were discontinued because of pain or discomfort, yet no other ill effects were observed."

The pioneering work of Ely and Goldman [29], [36] on effects of total immersion of animals in microwave fields had shown that, with a maximum available power density of no more than 100 mw/cm² fatal fevers were induced in rats, rabbits and tranquilized dogs. No animal survived in a treatment which raised the body temperature to 44°C (111.2°F). Continued exposure to 25 mw/cm² maintained a body temperature rise of 1°C, and 50 mw/cm² was lethal to rabbits and dogs.

Subsequently Michaelson, Dondero and Howland [64] of University of Rochester, Rochester, N. Y., who *avoided the use of premedication with tranquilizers*, reported their dogs gave no response to 45 mw/cm² and to 100 mw/cm², but a marked response at 165 mw/cm². A comparison of these data with those of Ely seems to indicate a significant difference. One might think that the use of medication by Ely could be an assignable cause. However, Susskind *et al.* at the University of California [55], present data to show that tranquilized mice withstood more than untreated ones. It was their opinion that the controlling factor was a critical effect of reducing the initial body temperature; considerably more energy was required for the tranquilized mice to reach the critical body temperature. Furthermore, Keplinger [61], [75] at University of Miami, Coral Gables, Fla., working at 24,000 Mc on unmedicated rats, determined that a level of 28 mw/cm² for 139 minutes was lethal and that 24 mw/cm² for 450 minutes was not lethal. At this frequency Schwan and Li have calculated that it is possible to have 100 per cent absorption in the skin, fat and muscle. If, indeed, 100 per cent was absorbed, then Keplinger's data are quite compatible with those of Ely and Goldman; namely, that 20 or 25 mw/cm² absorbed by the fur-bearing animal may be lethal. Tests conducted at M.I.T. Lincoln Laboratories, Lexington, Mass., are in substantial agreement with this conclusion [34], [35], [56].

These data are for fur-bearing animals. Ely, Goldman, Hearon, Williams and Carpenter [36], in their very thorough manner, point out, however, that man has a much better heat dissipating mechanism than fur-bearing animals. They estimate that the human being is capable of dissipating 1 kw of absorbed power. (This is not as much power as would be incident on a man lying prone in the noonday sun at the equator.) Their discussion as to how they arrived at this figure is substantiated by references to earlier work, and their conclusion seems reasonable if the energy is dissipated in the skin, fat and muscle near the surface where heat exchange with the cooling mechanism is good. However, if the energy is dissipated deeper in the body, in the hollow viscera, for example, as it might be for the longer radio waves, the heat exchange mechanism is no longer as effective and, as in the case of the eye, the internal temperature rise might become intolerable. Evidence of this effect was reported by Hines and Randall [22], who observed temperature rises of more than 40°C in the ileum of anesthetized rabbits after 30 minutes of localized irradiation at 12-cm wavelength, during which time the rectal temperature rose less than 1°C.

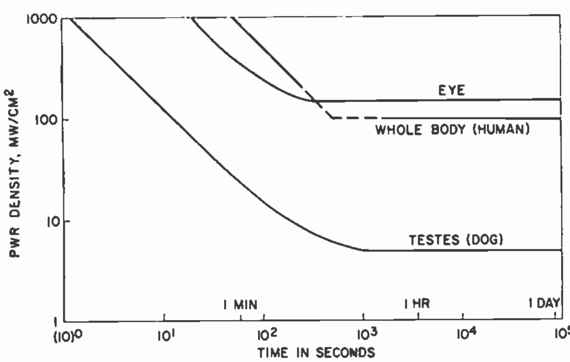

Fig. 9—Threshold levels vs time for three sensitive structures [36].

With the foregoing discussion in mind, let us look at the conclusions of Ely *et al.* [36] in more detail. Fig. 9 shows their preliminary estimates of threshold power densities according to time of exposure for three sensitive structures: the eye, the whole body, and the testes. Power density, in mw/cm² is plotted logarithmically on the ordinate, and time of exposure is plotted also logarithmically on the abscissa. The downward slope at short exposure time is based on their estimates of the thermal time constant of the structure. The terminal power density for long exposure is determined by the heat exchange characteristics of the structure and an estimated tolerable terminal temperature which was assumed to be 102.2°F for the whole body, 113.0°F for the eye, and 98.6°F for the testes. The power densities required to maintain these "tolerable" temperatures are 100 mw/cm² for the whole body, 155 mw/cm² for the eye, and 5 mw/cm² for the testes.

Note that the only threshold power density level which is lower than the currently adopted potentially hazardous level of 10 mw/cm² is that for the testes, 5 mw/cm². The question naturally arises, just how tolerable or how hazardous is a temperature of 98.6°F in the testes? Ely discussed this point briefly at a panel meeting held in New York, N. Y., on October 7, 1959, which was cosponsored by the New York Section of the IRE, the local chapter of the IRE Professional Group on Medical Electronics and the New York Section of the Communications Division of the AIEE. At this meeting Ely presented the graph which indicated that his estimated threshold for testicular exposure was 5 mw/cm². In answer to a question from the floor regarding the criterion for this limit, he went on to say that it is well known that undescended testes are sterile; that is, when testes are kept at a temperature of 98.6°F continuously, they become sterile. Furthermore, that in the experiments with dogs this temperature (98.6°F) could be maintained in the testes with a power density of 5 mw/cm².

Ely *et al.*[5] comment further on this point thus, "Although the criterion of hazard used in this study [of testicular changes] has been the least demonstrable damage, other factors should be considered in the overall viewpoint. The minimal testicular damage is almost certainly completely reversible. Even considerably more severe testicular insult will probably be reversible, with the only finding being a temporary sterility. An even greater injury can result in permanent sterility, which result would evoke varying reaction."

In regard to the varying reaction, there is the story of a radar station which had chronic failure every Friday night. Investigation disclosed that a technician had cut a large hole in the side wall of the transmitting waveguide and closed it up again with a flap. When this flap was kept closed the radar operated normally.

Each Friday night, however, before his buddies left for weekend leaves, he opened the flap and allowed the microwave power to escape into the room. He charged a fee for an exposure to this microwave power which, he claimed, would render the subject temporarily sterile for the weekend without jeopardizing any other capa-

[5] See [36], p. 91.

bility. Whether or not a temperature of 98.6°F in the testes is tolerable, the technician did a "land office" business for a while.

The possibility that nonthermal effects may exist was suggested by Schwan [48], [66] at both Tri-Service Conferences. Herrick [60] also suggests the possible existence of nonthermal effects in her report on pearl-chain formation. Schowalter of Western Electric Company points out, however, the similarity of this to the Reauleaux effect, which is induced by heat, the difference being only that in Herrick's observations, the pearl-chain formations were aligned (perhaps parallel to the electric vector), whereas in the Reauleaux effect the direction of chain formation is random.

More recently other investigators have reported effects which may be nonthermal. Heller and Teixeira-Pinto at the New England Institute for Medical Research, Ridgefield, Conn. [72] report on a new physical method of creating chromosomal aberrations. They also observed that the application of radio frequency energy to a medium in which paramecia were swimming in random directions caused these micro-organisms to swim parallel to the electric field. A change in the frequency caused them to turn and swim at right angles to the electric field. This work was done with pulsed power in the frequency range of 27 Mc. The power density at which these presumably nonthermal effects were observed was not given in the report.

Heller described his observations and showed some most amazing moving pictures (photo micrographs) of swimming paramecia at the twelfth Annual Conference on Electrical Techniques in Medicine and Biology held in Philadelphia, Pa., November 10, 11, 12, 1959. This meeting was sponsored by the IRE, the AIEE, and the Instrument Society of America. In answer to a question from the floor, he stated that the estimated electric field strength of the radio frequency pulses required to produce the observed nonthermal effects was of the order of 100 to 1000 volts/cm. If we knew the impedance of his medium, the corresponding peak power density could be calculated. For example, if the medium were water, with a relative dielectric constant of 80, the peak power density would be of the order of 3 to 300 watts/cm². In a typical radar with a duty cycle of one in a thousand, these peak, pulse power densities could be achieved in free space with an average power density of 3 to 300 mw/cm². The lower limit of the estimate is in the range of the currently adopted potentially hazardous level.

Carpenter [57] at Tufts University, has interpreted some of his data in terms of a possible nonthermal effect. Specifically, he found that rabbits' eyes exposed to 140 mw/cm² with *continuous waves* for 20 minutes *did not* develop opacities, but that rabbits' eyes exposed to 140 mw/cm² average *pulsed power* for 20 minutes *did* develop opacities. If the thermal time constant of the lens were long compared with the pulse length, then the cataractogenic nature of pulsed power densities could indeed be explained in terms of a nonthermal effect.

With the foregoing examples in mind, the implication is inescapable that we must not ignore completely the possibility of harmful nonthermal effects. We must hence keep our minds open for consideration of the adoption in the future of a limit for peak pulse power in addition to the limit already adopted for average power. Perhaps, too, someday we may have enough information on the effects at various frequencies to specify a limit that varies with frequency. More quantitative data are needed, however, before these effects may be taken into account properly.

In the meantime, pending further information, the limit has been based upon average power density. It is understandable that there was some question regarding the safety factor of the adopted limit of 10 mw/cm². Some organizations felt that it would be desirable to recommend that the power density in living quarters be kept somewhat below the upper limit. For example, the U. S. Army [63] had made an exhaustive study of the published literature and had made measurements at Army training installations. As a result of this study, they had arrived at some generalized criteria that they believed should be applied to training situations. One of these criteria was that rest areas should be provided for trainees at distances where the power densities were less than 1 mw/cm². This conclusion was compatible with Schwan's opinion [31] that 10 mw/cm² should not be allowed for more than one hour.

The General Electric Company [68] expressed the opinion that it would become necessary generally to monitor at a 1 mw-mean value in order to make the necessary allowance for harmonics and spurious waves.

Bell System Recommendations in Detail

The viewpoints of the Army, Schwan, and General Electric Company were shared by the American Telephone and Telegraph Company and the Bell Telephone Laboratories. Table II presents the summary of the Bell System recommendations.

Calculation of Power Densities

After having established the tentative exposure limits, attention was directed toward the problem of recommending simple transmission formulas for calculating the average power density in the beam of an antenna.

The field in front of the usual parabolic antenna can be characterized by referring to two separate regions:

1) The "near-field," or Fresnel region, where the radiation is substantially confined within a cylindrical pattern.
2) The "far-field," or Fraunhofer region, beyond the Fresnel region in free space, where the radiation is essentially confined to a conical pattern and

the power density along the beam axis falls off inversely with the square of the distance.

To compute the value of the power density in the near-field, assuming a circular "dish" antenna, use

$$W = \frac{16P}{\pi D^2} = \frac{4P}{A}, \tag{1}$$

where

P = average power output, *not* peak power,
D = diameter of antenna,
A = area of antenna.

If this computation reveals a power density which is less than the limit, then there is no need to proceed with further calculations, since (1) gives the *maximum* power density that can exist on the axis of the beam of a properly focussed antenna. (A defocussed antenna could give more, but that condition is not usual.)

If the computation from (1) reveals a power density greater than the limit, then one assumes that this value may exist any place in the near-field region, and attention is directed to the far-field region.

In the far-field region, the free-space power density on the beam axis may be computed from

$$W = \frac{GP}{4\pi r^2} = \frac{AP}{\lambda^2 r^2}, \tag{2}$$

where λ = wavelength.

The distance from the antenna to the intersection of the near-field (1) with the far-field (2), is given by

$$r_1 = \frac{\pi D^2}{8\lambda} = \frac{A}{2\lambda}. \tag{3}$$

These formulas do not include the effect of ground reflection which could cause a value of power density which is four times the free-space value.

Setting the power density equal to the potentially hazardous level of 10 mw/cm², one may calculate the distance to the boundary of the potentially hazardous zone. This has been done for some of the radars which have round antenna apertures. More involved calculations are necessary when the antenna shape and illumination are more complex. Other calculations have been reported for several types of radars [53], and these are included in Table III.

TABLE II

SUMMARY OF BELL SYSTEM RECOMMENDATIONS

1. For the time being, microwave exposure limits may be classified as follows:

Average Power Density mw/cm²	Classification
Above 10	Potentially hazardous
Between 1 and 10	Safe for incidental or occasional exposure
Below 1	Safe for indefinitely prolonged exposure or permanent assignment

2. Employees are cautioned to abide by the following rules:

a) Never enter an area posted for microwave radiation hazard without verifying that all transmitters have been turned off and will not be turned on again without ample notice.
b) Never look into an open waveguide which is connected to energize transmitters.
c) Never climb poles, towers or other structures into a region of possible high radar field without verifying that all transmitters have been turned off.

TABLE III

NIKE AND OTHER COMMON RADARS—DISTANCE IN FEET FROM RADAR ANTENNA TO BOUNDARY OF POTENTIALLY HAZARDOUS ZONE*

(Arranged in descending order of distances)

Radar Type	Distance in Feet for 0.01 watt/cm²
AN/FPS-16	
Sig C Mod.	1020
Standard Mod.	590
AN/FPS-6	560
Herc. Imp. Acq. HIPAR (Fixed)	550†·‡
AN/MPS-23	530
AN/MPS-14	472
Herc. Imp. TTR	400
AN/TPQ-5	350
AN/FPS-20	338
AN/MPQ-21 (10′)	300
Herc. MTR (Ajax)	270
Ajax Acq. (Fixed)	260†
AN/CPS-9	260
AN/MPQ-21 (7′)	210
AN/MPS-4	205
AN/FPS-8 (40′×14′)	205
Ajax MTR	205
AN/CPS-6B	200
AN/FPS-10	200
AN/MPS-22	185
AN/FPS-18	178
AN/MPS-12	175
AN/MPQ-18	175
AN/FPS-3	172
AN/MPS-7	172
AN/MPQ-21	165
AN/TPS-1G (40′×11′)	150
AN/FPS-36	150
Ajax TTR	132
Herc. Imp. Acq. (Fixed)	130†·‡
Herc. Acq. (Fixed)	130†·‡
AN/FPS-14	109
AN/FPS-4 (narrow pulse)	106
AN/MPS-8 (narrow pulse)	106
AN/TPS-10D (narrow pulse)	106
AN/MPS-10 (C)	105
AN/FPS-8	101
AN/MPS-11	101
Scr 584	70
AN/MPQ-10 (S)	50
AN/TPS-1-D	50
AN/FPS-25	40
AN/FPS-31	27.5
Herc. Imp. Acq. HIPAR (Rot.)	25
Ajax Acq. (Rot.)	8
AN/PPS-4	2.5

* Based on the following assumptions:

1) Free space transmission.
2) No ground reflections. These could double the distances shown.
3) Calculations apply to the axis of the beam, *i.e.*, where the power density is greatest.
4) The beam is considered to be fixed in space, *i.e.*, not scanning.

† Not normally used with fixed antenna.
‡ Interlocks provide assurance that transmitter is idle unless antenna is rotating.

Let us see how those approximate formulas, (1), (2), and (3), were derived and just how accurate they are.

Assume that we have a plane electromagnetic wave impinging on a totally absorbing body. If the projected area of the body is A, then the power absorbed will be the power density times the area or

$$P = WA. \tag{4}$$

If the field strength of the incident wave is E volts per meter in free space, then the power density in watts per square meter will be

$$W = \frac{E^2 \text{ volts per meter}}{377 \text{ ohms}}. \tag{5}$$

Ordinarily it is not necessary to specify the field strength in volts per meter, but only to specify the power density in watts per unit area if the impedance of the medium is known.

For an isotropic radiator in free space, radiating a total average power P in all directions equally, the power density on the surface of a concentric sphere of radius r will be simply the total radiated power divided by the area of that sphere, or

$$W = \frac{P}{A} = \frac{P}{4\pi r^2}. \tag{6}$$

Now if the radiator is not isotropic and radiates with a directivity gain G in a given direction, then the power density at a distance r would be, in the far-field region,

$$W = \frac{GP}{4\pi r^2}. \tag{7}$$

Allowing for 100 per cent ground reflection, which doubles the electric field strength (and hence quadruples the power density), we have

$$W = \frac{GP}{\pi r^2}. \tag{8}$$

If we express the gain G in terms of the antenna area thus

$$G = \frac{4\pi A}{\lambda^2} \tag{9}$$

and insert this in (7) and (8), we have the alternative expressions in terms of antenna area (3).

$$W = \frac{AP}{\lambda^2 r^2} \tag{10}$$

for the free-space power density, and

$$W = \frac{4AP}{\lambda^2 r^2}. \tag{11}$$

which assumes 100 per cent reflection from the ground.

It will be convenient to express the far-field free-space power density (10) in terms of the power density at the antenna aperture, $W_0 = P/A$, and hence $P = W_0 A$. Substituting this expression for P in (10) and dividing both sides by W_0, we have,

$$\frac{W}{W_0} = \left(\frac{A}{\lambda r}\right)^2. \tag{12}$$

For convenience, this may be rewritten thus

$$\frac{W}{W_0} = 4\left(\frac{A}{2\lambda r}\right)^2. \tag{13}$$

This expression applies in the far-field region, but a simple modification makes it applicable to the near-field region for a uniformly illuminated round aperture, thus

$$\frac{W}{W_0} = 4\sin^2\left(\frac{A}{2\lambda r}\right). \tag{14}$$

This more exact expression is plotted in Fig. 10 along with the lines representing the approximate formulas, (1) and (2).

In this graph and in some subsequent graphs, the relative power density is plotted in decibels on the ordinate and the distance, or more specifically $\lambda r/A$, is plotted logarithmically on the abscissa.

Note the alternate maxima and minima in the near field. The maxima all are 6 db above (4 times) the power density at the aperture. To be realistic, in the near-field region we must assume that a man could be standing at such a distance that he could be exposed to a maximum. Hence the near-field approximation

$$W = 4W_0 = \frac{4P}{A}. \tag{15}$$

This agrees with (1).

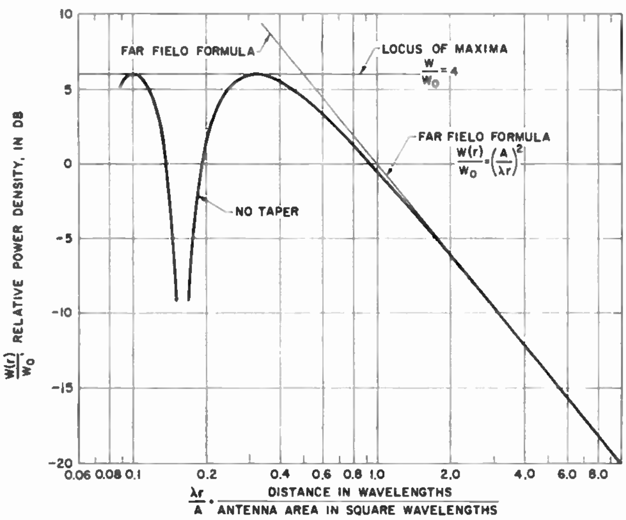

Fig. 10—Power density vs distance for uniformly illuminated round aperture.

Note that the near-field approximation intersects the far-field approximation when

$$r_1 = \frac{A}{2\lambda} = \frac{\pi D^2}{8\lambda} \tag{16}$$

as given in (3).

Also notice that the largest deviation of the approximations from the more exact expression occurs at this distance and this deviation amounts to only 1.5 db which, in many cases, is of little consequence. For all other distances the error will be less than this 1.5 db for the conditions assumed above.

Thus, it appears that these approximations are fairly reliable. They do, however, neglect the loss of power due to "spill over" at the antenna and the effectiveness of the antenna area and hence, in general, will give conservative estimates. Likewise, transmitter powers are often rated as power available *at the generator* so that any transmission line losses between the generator and the antenna would make the estimates even more conservative.

So far, we have assumed uniform illumination of a round aperture, whereas most antennas have the illumination tapered so as to reduce the sidelobes. Often a square-law taper to 10 db down at the edge of the aperture is used. Let us see what this does to *the power density when the total radiated power is kept the same for different tapers.*

Fig. 11 is a plot of the relative power density at the aperture of the antenna as a function of the radial distance from the center of the aperture for different tapers. For uniform illumination the relative power density is, of course, unity all over the aperture. However, for a linear electric-field taper, the power density is more than three times this value at the center, tapering to about 0.3 at the edge. For a square-law taper of the electric field, the power density at the center of the antenna is over twice that for the uniform case and tapers to about 0.2 at the edge.

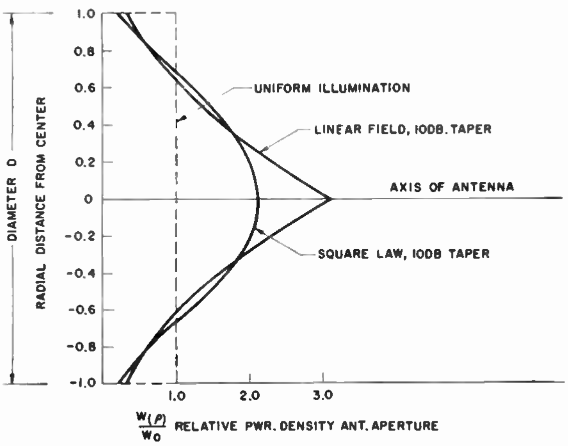

Fig. 11—Comparison of power densities at the aperture for different tapers of illumination.

Judging from those curves, one *might* think that the power density maxima along the axis in the near field and the power density in the far field might be quite different for these three tapers. Such, however, is not the case as will be shown in the following analysis.

We assume a circular aperture of diameter $2a$ with an illumination taper function $A(\rho)$. The field strength at a distance r along the axis of the antenna is given by the well-known relation

$$E_{(r)} = \frac{i\beta E_0}{r} \exp(-i\beta r) \int_0^a A(\rho) \exp(-i\beta(r'-r))\rho d\rho \tag{17}$$

where E_0 is the electric field at the center of the aperture,

$$\beta = \frac{2\pi}{\lambda}$$

$$A(\rho) = \text{normalized amplitude taper}$$

$$\exp(x) = e^x.$$

If we now neglect the angular dependence of the Huygens' sources, then, for distances more than a few diameters away, the following approximation may be used:

$$(r'-r) = \frac{\rho^2}{2r}$$

$$\left(\text{for } \frac{a}{r} \ll 1\right). \tag{18}$$

Thus, when the distance is much greater than the radius, a good approximation is given by

$$E_{(r)} = \frac{i\beta E_0}{r} \exp(-i\beta r) \int_0^a A(\rho) \exp\left(\frac{-i\beta\rho^2}{2r}\right)\rho d\rho. \tag{19}$$

For uniform illumination, $A(\rho)=1$ and the formula quoted previously is derived.

For a linear taper

$$A(\rho) = 1 - (1-\alpha)\frac{\rho}{a}, \tag{20}$$

where

$$\alpha = \frac{E(\rho=a)}{E(\rho=0)}.$$

Performing the quadrature results in the following expression:

$$\frac{W(r)}{W_0} = \frac{6}{1+2\alpha+3\alpha^2}\left|1 - \alpha\cos\frac{\pi}{2}\mu_0^2 - \frac{1-\alpha}{\mu_0}C(\mu_0) + i\left(\alpha\sin\frac{\pi}{2}\mu_0^2 + \frac{1-\alpha}{\mu_0}S(\mu_0)\right)\right|^2 \tag{21}$$

where

$$W_0 = \frac{P}{\pi a^2}$$

$$\mu_0^2 = \frac{2a^2}{\lambda r}.$$

C and S are the Fresnel integrals

$$C(\mu) = \int_0^\mu \cos\frac{\pi}{2} v^2 dv$$

$$S(\mu) = \int_0^\mu \sin\frac{\pi}{2} v^2 dv$$

as tabulated by Jahnke and Emde.[6]

This expression is plotted in Fig. 12 where it is seen that the major effect of the taper has been to fill in the deep nulls which existed in the near field for the uniformly illuminated case. The difference between this calculated curve and the approximate far-field formula (2) is less than $\frac{1}{2}$ db in the far field and again about 1.5 db at the intersection where $r = A/2\lambda$.

In the near field, the maxima are within about $\frac{1}{2}$ db of the $W = 4W_0$ line.

For the case of a square-law taper of the electric field, the amplitude function becomes

$$A(\rho) = 1 - (1-\alpha)\left(\frac{\rho}{a}\right)^2. \tag{22}$$

Performing the indicated quadrature results in the following expression

$$\frac{W(r)}{W_0} = \frac{3}{1+\alpha+\alpha^2}\left|1 - \alpha\cos\frac{\pi}{2}\mu_0^2 - (1-\alpha)\frac{\sin\frac{\pi}{2}\mu_0^2}{\frac{\pi}{2}\mu_0^2} + i\left[\alpha\sin\frac{\pi}{2}\mu_0^2 + (1-\alpha)\frac{1-\cos\frac{\pi}{2}\mu_0^2}{\frac{\pi}{2}\mu_0^2}\right]\right|^2 \tag{23}$$

where again

$$W_0 = \frac{P}{A},$$

and

$$\mu_0^2 = \frac{2a^2}{\lambda r}.$$

A plot of this expression is given in Fig. 13. One curve and two straight lines are plotted. The straight lines are the approximations of (1) and (12). The curve is for a square-law taper of 10 db. It is seen that the departure from the approximate formula is not of much consequence.

Thus, it is seen that tapering the illumination of a round aperture has not affected the power density on the axis enough to cause concern in either the far field or the near field.

What about other shapes of antennas, for example a square antenna? The formula for a square antenna with uniform illumination is given by the relation

$$\frac{W(r)}{W_0} = 4\left[C^2\left(\frac{d}{\sqrt{2\lambda r}}\right) + S^2\left(\frac{d}{\sqrt{2\lambda r}}\right)\right], \tag{24}$$

where

$$W_0 = \frac{P}{A} = \frac{P}{d^2}.$$

[6] E. Jahnke and F. Emde, "Table of Functions," Dover Publications, Inc., New York, N. Y., p. 34; 1943.

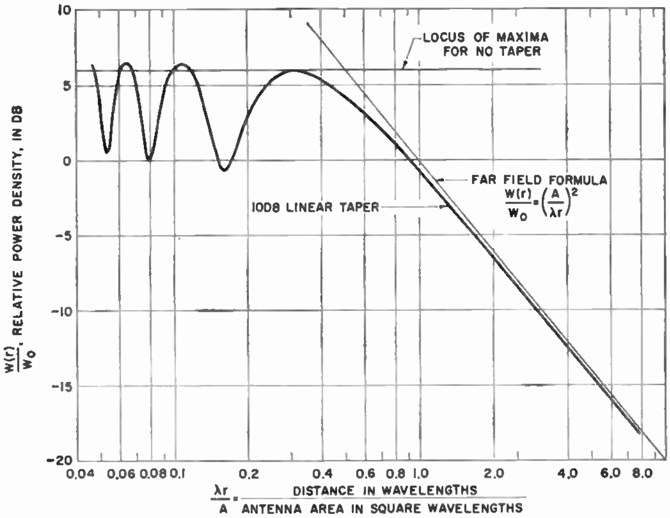

Fig. 12—Power density vs distance for round aperture with linear taper (10 db).

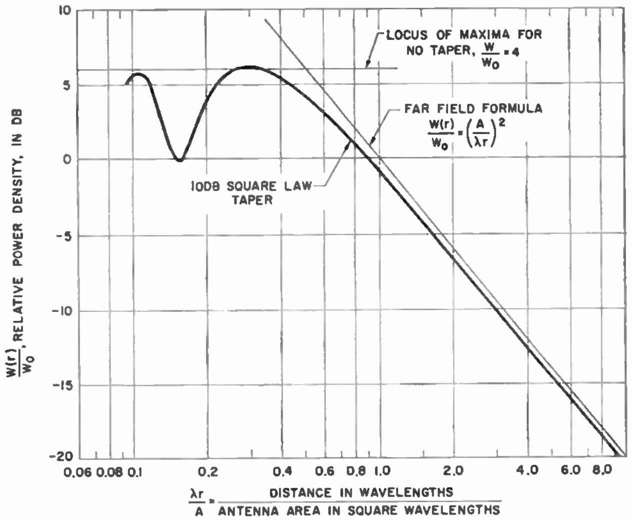

Fig. 13—Power density vs distance for round aperture with square-law taper (10 db).

This is plotted in Fig. 14, together with the approximations based upon *equal area antennas*. For the square antenna the near-field maxima are lower than the approximation, but the one at the greatest distance is only about one db low. In the far field, the asymptotic agreement is excellent.

To summarize these observations, Fig. 15 is presented. This shows the near- and far-field approximations along with the maxima for the round antennas with linear and square-law 10-db tapers and for the square antenna with uniform illumination.

The abscissa is plotted on $\lambda r/A$ which can be applied to either round or square antennas. This shows that the approximate formulas (1), (2), and (3) can be applied to square antennas by choosing a diameter which makes the areas equal.

Thus, the approximate formulas are seen to apply quite well to round and square antennas. For more exact results at the region of crossover between near and far fields, the expression containing $\sin^2$ may be used (14).

For long rectangular antennas, an approximate formula derived by Engelbrecht is

$$\frac{W}{W_0} = \left(\frac{A}{\lambda r}\right)^2 \text{ beyond the distance } r \geq \frac{d_1^2}{2\lambda}, \qquad (25)$$

and

$$\frac{W}{W_0} = \frac{2d_2}{\lambda r} \text{ within the distance } r < \frac{d_1^2}{2\lambda} \qquad (26)$$

d_1 = wide dimension

d_2 = narrow dimension.

This near-field formula was derived from a more complete analysis for distance such that

$$\frac{d_1}{r} < 1$$

and

$$\frac{d_2^2}{\lambda r} < 1.$$

Reviewing the foregoing discussion, it is seen that a simple calculation of the maximum power density in the near field according to (1) reveals whether or not a hazard is involved in the beam of the antenna. If this power density exceeds 10 mw/cm², then a simple calculation based upon the far field formula (2) reveals the potentially hazardous distance in free space. These formulas apply to uniform illumination of square, round or rectangular apertures or to illumination which is tapered in amplitude for round apertures. For other shapes and tapers a more complicated analysis would be necessary.

To calculate accurately the power density off the axis of the main beam requires the solution of a more difficult mathematical problem. One approach [53] reveals that the collimated beam in the near field falls off approximately 12 db per radius.

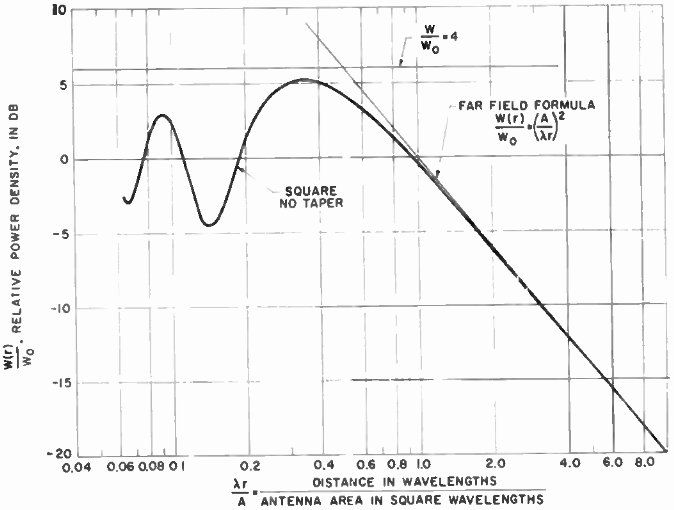

Fig. 14—Power density vs distance for square aperture uniformly illuminated.

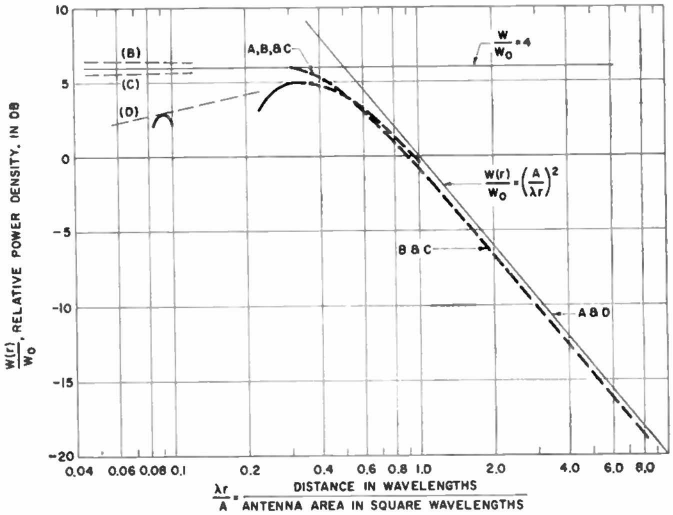

Fig. 15—Comparison of power densities for different shapes and tapers: A. Round—no taper; B. Round—10-db linear taper; C. Round—10-db square-law taper; D. Square—no taper.

Many of our radars do not have the simple shapes nor simple illumination tapers that were treated above. In such cases the approximate formulas will not apply directly and a more complete analysis is indicated.

Scanning Antennas

The specified limits for safe and potentially hazardous power densities have been based upon average power. In the case of the scanning antenna, the average power absorbed by a fixed subject will be reduced by the ratio of the effective beamwidth to the scanned angle, if the thermal time constant of the part of the exposed individual is long compared with the scanning period. Accordingly, the potentially hazardous distance is reduced by the square root of this ratio.

The effective beamwidth in the far field will, in general, be somewhat greater than the 3-db beamwidth, and somewhat less than the width to the first null. The exact value depends, of course, upon the form factor of the radiation pattern.

In the near field, the effective angle of the beamwidth will vary with distance since the field is collimated. Here the average power density of the scanning antenna is given approximately by the relation

$$W \doteq W_0 \frac{D}{2\pi r} \times \frac{360}{\theta}, \tag{27}$$

for

$$\theta \geq \frac{D}{2\pi r} \times 360,$$

where θ is the scanned angle in degrees. For

$$\theta < \frac{D}{2\pi r} \times 360 \text{ degrees}$$

$$W \doteq W_0. \tag{28}$$

Since

$$W_0 = \frac{4P}{\pi D^2} \text{ (circular aperture)} \tag{29}$$

$$W \doteq \frac{4P}{\pi D^2} \times \frac{D}{2\pi r} \times \frac{360}{\theta}.$$

for

$$\theta \geq \frac{D}{2\pi r} \times 360.$$

And

$$W = \frac{4P}{\pi D^2}, \tag{30}$$

for

$$\theta < \frac{D}{2\pi r} \times 360.$$

Setting $W = 10$ mw/cm², the potentially hazardous distance in the near-field is

$$r_{\text{hazardous}} \doteq \frac{1}{5\pi^2} \cdot \frac{P}{D} \times \frac{360}{\theta} \text{ cm} \tag{31}$$

for

$$\theta \geq \frac{D}{2\pi r} \times 360 \quad \text{and} \quad r \leq \frac{D^2}{4\lambda}.$$

where P is in mw, D is in cm.

For

$$\theta < \frac{D}{2\pi r} \times 360,$$

the previous discussion is applicable.

The potentially hazardous level of 10 mw/cm² was based on average power measurements. In fact, many of the experiments on animals were made with continuous-wave oscillators and the deleterious effects appeared to be due solely to the production of heat and the accompanying rise in temperature. If the duty cycle of the radar is one in a thousand, the peak-power density associated with this level would be 10 w/cm². It may be reasonable to assume that this peak-power density may not cause damage by breaking down of tissues. However, if the radar duty cycle were say $(10)^{-7}$ the peak-power density would be 100 kw/cm², approaching the level which causes breakdown in air. At this level, it would be reasonable to assume that there might be a possibility of breakdown of tissue. Somewhere between the two examples just cited, danger due to breakdown of tissue from effects other than heating may be possible. A similar situation exists for very narrow-beamed sweeping antennas, whose exposure cycle is very small. The observation of Heller and Pinto [72] suggests effects other than thermal and if, in fact, these nonthermal effects are due to peak power, rather than average power, a more elaborate specification of safe and hazardous levels will be necessary.

Measurement of Microwave Power Density

The measurement of the power density depends basically upon the determination of the power absorbed in a given area. Knowing the power absorbed by the given area, one computes the power density by dividing the absorbed power by the area.

Devices for measuring microwave power have been available for many years and are described in standard text books. A brief description is given by Southworth,[7] and a comparison of power-measuring techniques is discussed by Sucher of the Microwave Research Institute, Polytechnic Institute of Brooklyn [25]. General details are given by Green, Fisher, and Ferguson [8]. Montgomery devotes one whole chapter to the discussion of microwave power measurements [9].

There are four basic methods of measuring power. These are: calorimetry, bolometry, voltage (and resistance) measuring systems, and measurement in terms of radiation pressure on a reflecting surface [20]. Calorimetric methods are based upon the transfer of the electromagnetic energy into heat, and the power is determined solely by the measurement of temperature, mass, and time. With a knowledge of thermal capacity, only temperature and time need be measured.

Bolometric measurements are based upon the absorption of the power in a temperature-sensitive resistive element. The change in resistance is then used to indicate the power. This method is the one most widely used in commercially available power meters, and the most popular sensing element is the thermistor, de-

[7] See [16], p. 655.

scribed by Southworth: [16]

"Certain mixtures of semiconductors have especially high temperature coefficients of resistivity. Resistor units made up of such materials are therefore thermally sensitive . . . [and] the measurement of power is based upon this resistance change. . . . Power measurements are based on calibration made at [dc or] low frequencies.

"To measure the change of resistance as power is dissipated, the thermistor is often placed in one arm of a dc bridge. By noting the dc power necessary for balance with and without RF power applied to the thermistor, the difference is determined. In a modification of this method, the bridge is balanced before RF power is applied, and the measurement consists merely of noting the meter deflection (due to the bridge unbalance) when RF power is added. It is possible to make such a power meter direct reading. . . ."

There are many commercially available thermistor bridges on the market. However, the electronic circuits of some of these are not shielded adequately to be unaffected by the high-power microwave-pulsed energy of a radar. The peak pulse power may be 1000 or more times the average power, and since in the application to radio-frequency hazards we are currently interested chiefly in average powers from 1 to 100 mw; the peak pulse power may be from 1 to 100 watts. Without adequate shielding, the vacuum tubes associated with some of these power meters may be susceptible to the microwave power interference and readings may be obtained even when the pick-up antenna is disconnected.

There are, however, some commercially available thermistor bridges designed specifically to work in an environment of pulsed microwave radiation. These, when used in conjunction with an appropriately calibrated antenna and a matched thermistor in an RF head, are suitable for power density measurements.

If the radiated power is continuous, rather than pulsed, a much simpler relative power meter using a readily available diode rectifier and dc meter in conjunction with a calibrated antenna may be used. After a calibration in the laboratory, this type of unit is convenient, because of its light weight and small size, to carry around in the vicinity of the antenna for survey purposes or for monitoring. These are not, however, suitable for pulsed-power applications since the high-peak-pulse power may burn out or damage the delicate point contact of the microwave diode.

All of these power measuring devices are used with a suitable antenna whose effective area is known. "Standard" waveguide horn antennas, loops and dipoles are available commercially whose effective areas are known. For example, the effective area of a lossless loop or small dipole (doublet) is $\frac{3}{8}$ of the wavelength squared divided by π [16], [18], and has a gain, over an isotropic antenna (whose effective area is the square of the wavelength divided by 4π) of 1.76 db. A half-wavelength antenna has an absolute gain (over an isotropic radiator) of 2.15 db and an effective area of 0.131 times the square of the wavelength.

Waveguide horns, either circular or rectangular, are amenable to calculation [18], and their effective areas range from about one half the physical area for optimum horns, to about 80 per cent for very long horns. An optimum horn is one whose open-end area is scaled to give the maximum gain for that particular length of the flare in the waveguide. Specifically, optimum circular horns have an effective area equal to 51 per cent of the actual area, while optimum rectangular horns have 49 per cent effectiveness. An infinitely long circular horn has a relative effectiveness of 84 per cent while infinitely long rectangular horns have 81 per cent effectiveness. Such "standard" antennas are used to pick up the energy which is then transmitted by a well-shielded transmission line to the thermistor-sensing element whose resistance has been adjusted so as to match the impedance of the transmission line. All of the power is dissipated in the thermistor, causing its temperature to rise, its resistance to change, which unbalances the dc bridge and the unbalance current indicates the power, from which the power density is calculated.

Recently there have appeared on the market at least four equipments designed specifically for microwave power density measurement. There are:

1) "Broadband Power Density Meter," Model NF-157, Empire Devices, Amsterdam, N. Y., covering the frequency range from 200 Mc to 10,000 Mc with three different RF pick-up probes. The power density for midscale reading is 1 mw/cm² to 1 w/cm². The claimed accuracy is ±1 db at midscale. The weight is 11 to 13 pounds.
2) "Electromagnetic Radiation Detector," Microline Model 646, Sperry Microwave Electronics Company, Clearwater, Fla. Three units are available to cover three bands (2700 to 3300 Mc, 5400 to 5900 Mc, and 8200 to 12,000 Mc). The range of power density is 1 to 20 mw/cm² with 10 mw/cm² at midscale. The weight is about 8 pounds.
3) "Densiometer," Model 1200, Radar Measurements Corporation, Hicksville, N. Y., covers 5 bands with 4 different antennas: VHF (200 Mc–225 Mc); UHF (400 Mc–450 Mc); S band (2600 Mc–3300 Mc); C band (5000 Mc–5900 Mc), and X band (8500 Mc–10,000 Mc). The power density, for midscale reading is 10 mw/cm². The weight is about 2 pounds. It comes with a handy "exposure meter" type of carrying case.
4) Radiation Monitor, Model B86B1, Sperry Microwave Electronics Company, Clearwater, Fla., covers from 200 to 10,000 Mc with one antenna. The power density for midscale reading is 10 mw/cm². It receives all polarizations simultaneously. The claimed accuracy is ±2 db at any frequency within the operating range. The weight is 2 pounds.

The devices which rely upon antenna pick-up probes all have two limitations in common: they are frequency sensitive, and they are directive. These limitations are no handicap if the power density meter is being used to survey a particular transmitter site where the frequency, direction and polarization of the radiation are all known. However, for use as a detection device by the telephone serviceman who approaches an area which is merely posted as being hazardous without specifying the frequency and direction of the microwave radiation, they are of rather limited usefulness.

What is needed here is something that responds to all radio frequencies, regardless of polarization and direction of arrival. Perhaps some day we shall have such a device. It will probably be based either on the calorimetric method, using an absorber which acts as a black body, or on the radiation pressure principle. This problem has been and is being given considerable thought in various research centers.

Some people have suggested using the glow discharge in a neon lamp as an indicator. These lamps are known to respond to the peak-pulse power, rather than the average power, since "the ionization and deionization times of neon are presumably much less than the width of the pulse modulation envelope."[8] These lamps are therefore unsuited for indicating average power density.

Further information on the determination of power density at microwave frequencies has been given [24], [44].

Regarding the procedure in making a survey of a particular site in the field, if such a survey seems to be warranted on the basis of a calculated estimate, Dondero [44] points out that, after selecting the appropriate measuring equipment, "it is always good practice to perform initial measurements at some practical (safe) distance from the radiator and gradually work in toward the source of radiation." This provides some safeguard for personnel performing the measurement. This procedure differs from that which has been used in making field strength surveys patterned after the style set in the early 1920's by Bown [1], [2], who measured the coverage of a radio broadcasting station, in that the survey is always made by *approaching* the transmitter. Also, it may be advisable to perform the survey at a known reduction of power and interpret the results in terms of power density for full transmitter power.

Reflection may complicate the field strength pattern. It has been known for many years [3] that obstacles such as trees, buildings, telephone lines, etc., reflect radio waves to a certain extent and produce standing waves in the field strength pattern. It is essential to keep this in mind while making the survey and to move the pick-up antenna around, this way and that, to establish whether the reading is being affected by reflections. If it is, the maximum reading should be recorded.

If a directive antenna is used, the horn should be directed not only toward the transmitting antenna but also around in other directions toward any object nearby which might act as a reflector. If a significant reading is obtained from such a reflecting object, then the total power should be figured, not by adding the two powers, but by figuring the square of the sum of the square roots of the powers. In other words, since the direct and reflected waves are coherent, coming from the same source, their electric field strengths will add vectorially to give a maximum when they are in phase. The corresponding maximum power is proportional to the square of the sum of the field strengths. Since the field strengths are proportional to the square roots of the powers, we have, therefore, the square of the sum of the square roots of the powers.

However, in a complex situation where there are no reflections but several different transmitters, for example, as would exist at the operational field-radar site, then the power densities from the radars on different frequencies should be added directly and not the field strengths, since, in general the radar frequencies will not be identical.

Having once established the boundary of the potentially hazardous zone, where power densities of 10 mw/cm^2 may occur, appropriate measures must be adopted to keep personnel from entering that zone. Barricades or shielding fences may be installed. If gates must be provided for occasional access to the area by personnel, interlocks should be provided to ensure that the transmitter be idle whenever personnel open the gate and enter the area.

In some cases it may be necessary that the transmitter be kept operative while personnel enter into an area where the field is potentially hazardous. In this event, some kind of a shielding arrangement is indicated. Whether this be of the fence or shielded room type or the mobile-portable shielding-garment type, some knowledge of the effectiveness of the shield is necessary in order to be sure that the power density within the shielded area is low enough to be safe. Of course, completely enclosing the area in a water-tight copper container would most surely be adequate, but neither economical nor practical. A wire mesh might also be adequate and much cheaper and more practical from the standpoint of the circulation of air.

Protective Mesh—Transmission Through a Grid of Wires

How effective is a wire screen? It is apparent that the more wide open the mesh, the less the shielding. In order to answer this question, some early published work [5] was reviewed. Laboratory tests were made both at the Bell Telephone Laboratories in Whippany,

[8] See [9], p. 220.

N. J., and at Wheeler Laboratories in Great Neck, N. Y., in an attempt to check the earlier work. It was found that *good* agreement was obtained at the *same frequency* as that of the earlier experiment, but that quite *poor* agreement was obtained at *three times that frequency.*

Looking further into the published literature, the formulas of Schelkunoff and Sharpless [4] and that of Marcuvitz [17] were tried. A somewhat better agreement with the data was obtained, but still neither formula was satisfactory. However, by combining the two formulas properly, since one neglected one thing and the other neglected something else, an empirical formula was derived which checked all of the available experimental data quite well. This empirical formula was then revised so as to make it suitable for use in nomographic form and the nomograph was constructed, as seen in Fig. 16. It applies to normal incidence on a grid of parallel wires of diameter $2r$ having a spacing a between centers. The electric vector is assumed to be parallel to the wires. To use the nomograph, a straight edge is aligned with one point at the left corresponding to the spacing in wavelengths and another point at the right corresponding to the ratio of the spacing to the radius of the wire. At the point where the straight edge crosses the line in the middle labeled "Transmitted Power (db)," the shielding effectiveness of the grid of wires is expressed in decibels.

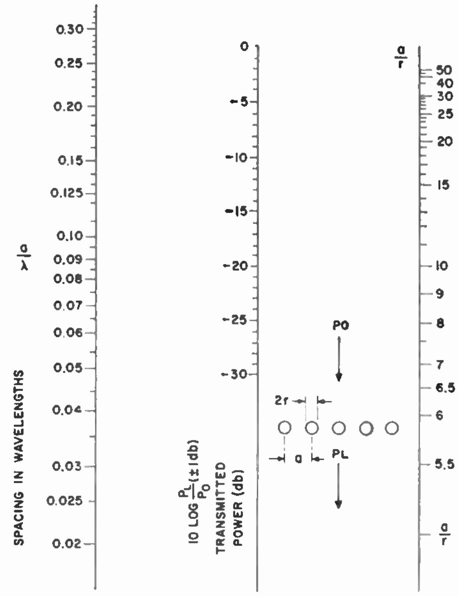

Fig. 16—Transmission through a grid of wires of radius r and spacing a.

The modified empirical formula upon which this nomograph was based is the following

$$\frac{P_0}{P_L} = \frac{B^2}{4}, \tag{32}$$

where

$$|B| = \frac{\lambda}{a} \frac{1}{\ln\left[\dfrac{\left(.83 \exp \dfrac{2\pi r}{a}\right)}{\exp\left(\dfrac{2\pi r}{a}\right) - 1}\right]}$$

P_0 = Incident power,

P_L = Transmitted power.

The accuracy of the nomograph appears to be slightly better than ± 1 db, judging from a comparison with the available measured data.

The results are applicable equally well to a screen of perpendicular wires by ignoring one or the other set of parallel wires forming the mesh.

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A Very Low Frequency (VLF) Synchronizing System*

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Summary—**A receiving system for the use of standard-frequency Navy and NBS VLF transmissions to phase-control a local frequency standard has been designed, built, and tested. Frequency synchronization accuracies of better than one part in 10,000,000,000 are practical, and the electronic instrumentation of the receiving system is not complex. Techniques for time synchronization to similar accuracies are not yet available; however, a time standard may be maintained in time accuracy to within 10 μsec once a synchronization is performed.**

Introduction

One of the most demanding requirements in tracking satellites is the time accuracy necessary to be compatible with the angular precision of the interferometer used in the Minitrack system of the National Aeronautics and Space Administration. The present status of over-all tracking system accuracy requires time knowledge to no worse than perhaps two milliseconds. This timing requirement may not be apparent, and the following comments may help to clarify the statement. The basic problem in orbital computation is to obtain sufficient data points around the orbit to define the orbital parameters. A single station can obtain, at most, two points on the orbits, and will get data only twice a day. Two stations, properly chosen for a given orbit, may give as many as four data points at six hour intervals; however, the locations of these stations may not be optimum for all possible satellite orbits, and a station may occasionally miss a data point. The Minitrack stations were chosen to provide adequate coverage in both angular dispersion around the orbit and minimum time between observations. The Minitrack stations are located in or near Fairbanks, Alaska; East Grand Forks, Minn.; Newfoundland; Blossom Point, Md.; Fort Myers, Fla.; Antigua, West Indies Federation; Quito, Ecuador; Lima, Peru; Antofagasta, Chile; Santiago, Chile; Johannesburg, South Africa; Woomera, Australia; San Diego, Calif.; and England. A large percentage of these stations are along the 75th meridian to intercept satellite orbits with inclinations to the equator of less than 45°, and the four northernmost stations provide coverage for the more nearly polar orbit satellites. The accuracy of any individual station is maintained at less than 20 seconds of arc by regular calibration flights of an airplane-borne antenna, transmitter, and flashing light. Through calibration of light images against stars as photographed by an astrographic camera, and correlation against simultaneous electronic records, the electronic calibration is performed to an accuracy of four seconds of arc. If each station's angular measure may depended on to an accuracy of 20 seconds of arc, then the time tag on each data point must not impair this accuracy, and one station's time relative to any other station's time must be known to within the same accuracy. For a satellite in a circular, 300-mile, orbit, the time required to cover 20 seconds of arc is six milliseconds. If time knowledge is no worse than one to two milliseconds, the angular accuracy will not be materially affected. Since time is proportional to electrical phase angle, the long-term accuracy requirement on the frequency source for the time standard is infinite. The system in use at the Minitrack stations requires daily adjustments of phase angle and frequency to cope with the oscillator stability of one part in 100,000,000 per day, and still achieve one to two millisecond timing errors referenced to Washington, D. C., time. The signals from the National Bureau of Standards, radio station WWV are used to synchronize the world-wide tracking net, but the propagation time of high frequency (HF) signals is not predictable to better than one to two milliseconds.[1,2] VLF signals can now be used to maintain a local frequency standard to a frequency accuracy of one part in 10,000,000,000 with relatively simple instrumentation as described below; however, the problem of time synchronization to better than one-half millisecond has not yet been satisfactorily solved.

Satellite experiments have been proposed which will require increased tracking accuracy in order to make more precise geodetic measurements. Proposed experiments to delve more deeply into the relativistic theory may require time accuracy beyond the present capability of HF systems. Present satellite magnetometer experiments require tracking and computation accuracies barely within the timing accuracy of one to two milliseconds.

For a number of years, Professor Pierce of Harvard University, Cambridge, Mass., has been measuring phase and frequency stabilities of VLF transmissions.[3] His work indicates that potential phase stabilities of a

* Received by the IRE, July 20, 1960; revised manuscript received, November 14, 1960.

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[1] M. Boella and C. Egidi, "Measurements of the time of propagation of time standard signals," *Alta Frequenza*, vol. 24, pp. 309–338; August–October, 1955.

[2] A. H. Morgan, "Precise Time Synchronization of Widely Separated Clocks," Natl. Bur. Standards, Washington, D. C., NBS Tech. Note 22; July, 1959.

[3] J. A. Pierce, "Intercontinental frequency comparison by very low frequency radio transmission," Proc. IRE, vol. 45, pp. 794–803; June, 1957.

few microseconds are definitely possible. Recently, the U. S. Navy began transmissions on 18 kc from Panama for timing purposes. Their transmitted frequency is monitored at the Naval Research Laboratory, Washington, D. C., against an Atomichron reference, and is adjusted to be 150 parts in 10^{10} low in order to approximate time as specified UT_2.[4] The Navy station, call sign NBA, has a radiated power such as to give good reception over North and South America and as far away as Australia. NBA will be transmitting 24 hours a day, seven days a week by about July 1, 1960, with a keyed carrier, on 300 milliseconds, starting each second except the 29th, and the 56th through 59th. This pattern is broken every 15 minutes to identify the station. Still more recently, the National Bureau of Standards initiated transmissions on 20 kc from Boulder, Colo. This station, call sign WWVL, is presently broadcasting, five days a week, a continuous signal broken only at twenty-minute intervals for station identification. The radiated power of WWVL is not as large as that of NBA, but the signals are usable throughout the continental United States, and there are plans for increasing the power at some future time.

While VLF signals possess remarkable phase stabilities and do permit synchronizing frequencies to better than one part in 10,000,000,000, there is not yet a dependable method for synchronizing time to the accuracy necessary to realize the full capability of VLF propagation. In the frequency range of interest (18–20 kc), a carrier cycle is 50 microseconds long. In order to resolve the ambiguity inherent in phase angle measurement, a timing mark must be obtained which is known to an accuracy of one-half a cycle or approximately 25 microseconds. The pulsed transmission of a VLF station having an efficient antenna is limited to from 10 to 15 milliseconds rise time. Assuming quiet atmospheric conditions and an excellent signal-to-noise ratio, the start time of the VLF pulse might be determined to 100 microseconds, but a more practical value, allowing for atmospherics and operator error, would be 500 microseconds. Work is in progress in several laboratories to develop correlation techniques in an attempt to improve this 500 microsecond figure, but probable eventual accuracy will not be less than 50 microseconds. The Bureau of Standards is investigating approaches utilizing multiple-frequency transmission; however, these techniques may require more complex receivers than the single-frequency system, although timing marks could be obtained with more than sufficient accuracy to resolve the ambiguities. The problem of propagation time would also require solution, but transportation of a precise clock by air to the remote site, and return, could determine the propagation time sufficiently. This physical transportation of time could also be used to effect the time synchronization of the remote clock, but would then have to be repeated at every occasion of equipment malfunction at the remote site. A more acceptable approach from the standpoint of simplicity of field equipment and operations will be one utilizing radio transmission of time.

[4] U. S. Naval Observatory, Washington, D. C., Time Service Notice No. 4; April 15, 1959.

VLF Receiving System

The VLF synchronizing system in use at the Goddard Space Flight Center, Greenbelt, Md., is a phase controlled servo, as shown in block diagram form in Fig. 1. The approach taken in this system is to avoid manipulation of the master oscillator frequency controls directly, but to add in a continuous phase-angle correction to create a frequency-controlled time standard. By

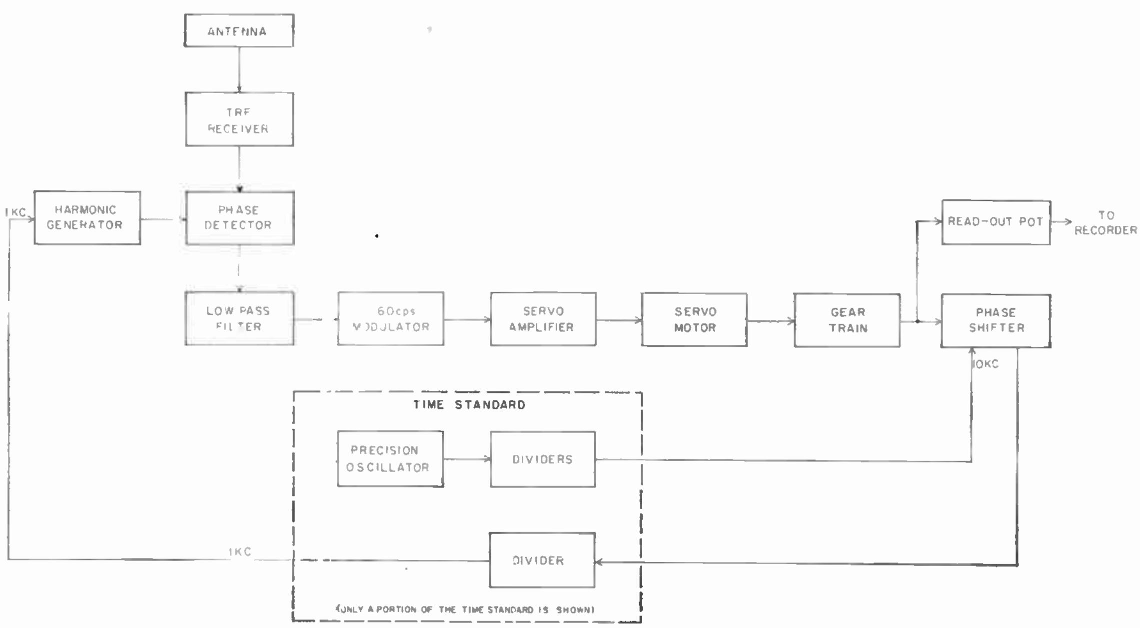


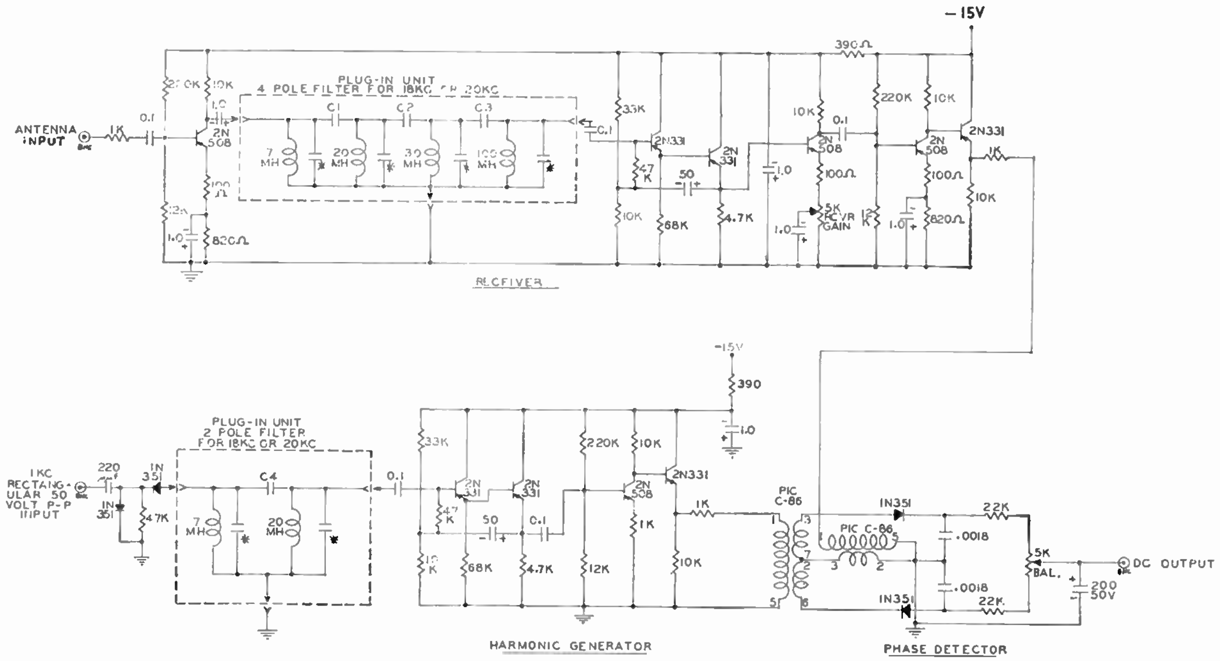
Fig. 1—Block diagram of VLF synchronizing system.

recording phase shift as a function of time, data is obtained permitting manual adjustment of the master oscillator frequency to minimize necessary phase-angle corrections. A 500-foot antenna supplies a signal to the TRF receiver for amplification and frequency selection. The harmonic generator provides a signal of the same frequency as the received signal, and the phase detector develops the servo error signal. The low-pass filter is necessary to reduce noise saturation effects in the servo-amplifier and thus reduce heating. The phase shifter varies the phase angle of the frequency source for the local time standard. The output frequencies of the local time standard are therefore proportional to the master oscillator frequency with its phase angle modified by the servo-controlled phase shifter; the phase angle being shifted continuously, if necessary, to track the phase angle of the incoming reference signal. Consequently, the combination of the master oscillator, phase shifter, gear-box, motor and amplifier constitute an electromechanical voltage-controlled oscillator with a frequency deviation proportional to motor shaft speed.

Receiver

The receiver design is based on a requirement of maximum phase stability. The frequency selection is performed by a four-pole filter designed in accordance with procedures in "Reference Data for Radio Engineers"[5] to obtain a 3-db bandwidth of 430 cps. The two transistors connected to the filter output (Fig. 2) are cascaded emitter-followers to minimize filter loading. Although the receiver selectivity is 430 cps, the system bandwidth is far narrower. For example, the signal from NPM, Hawaii, 19.8 kc, is 15 to 20 db higher than the WWVL signal, but because it lacks phase coherence with the reference, it is filtered out by the low-pass filter and the phase-locking action of the system until it is hardly discernible. The over-all system bandwidth is approximately 0.003 cps, and until an undesired signal comes within a small fraction of a cycle per second of the reference, or has an amplitude so large as to cause severe distortion, no appreciable interference will result. The receiver maximum gain from antenna to phase-detector emitter follower is approximately 100,000. Noise level at the present location of the Time Measurement Branch of Goddard Space Flight Center is the limiting factor on receiver gain; however, sufficient output is provided for synchronizing to WWVL with a noise of approximately 0.1 μsec rms. The noise is quoted in microseconds because the system output is a graph with a calibration of 100 μsec full scale. To quote the signal-to-noise ratio in volts is meaningless since the system output is not a voltage. The purpose of the system is to synchronize a local oscillator with an incoming reference signal as accurately as possible, and since the noise is a measure of the synchronization accuracy, it is expressed in units of time. NBA provides

[5] "Reference Data for Radio Engineers," 4th Ed., ITT, Nutley, N. J., pp. 187–228; 1956.



	c1	c2	c3	c4
18 kc	330	100	33	150 μμf
20 kc	270	68	22	120 μμf

* Tune for 20 kc or 18 kc resonance.
All resistors are 1/2 watt unless noted.
All capacitors are microfarads unless noted.

Fig. 2—Schematic diagram of VLF receiver.

an antenna signal of approximately one millivolt and drives the receiver to saturation at maximum gain. Transistors used in the receiver are not critical; any transistor with a current gain in excess of 50 should provide adequate performance. The first transistor should be a moderately low-noise type, but transistor noise is generally lower than interfering signals and noise.

Harmonic Generator

The harmonic generator uses a harmonic-rich, 1-kc, rectangular wave from the local time standard to drive circuitry strongly resembling the receiver. The two-pole filter is designed from the same handbook as that for the receiver, and the transistor stages are identical to those in the receiver. The amount of modulation on the resultant 18- or 20-kc signal is less than 10 per cent in both magnitude and phase.

Phase Detector

The phase detector is the ratio type and has proved to be quite stable. The only adjustment necessary is the balance potentiometer used to obtain minimum "creep" in the absence of a signal from the antenna. The "gain" of the phase detector is approximately unity; the dc voltage for 0° or 180° phase angle is approximately equal to the input to the base of the emitter-follower in rms volts. Fig. 2 is a schematic of the receiver, harmonic generator and phase detector.

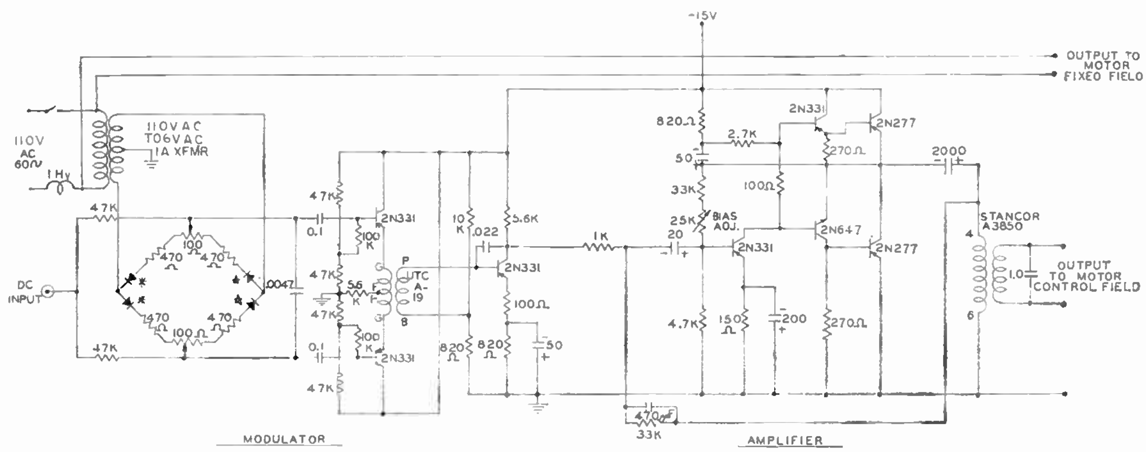
Servo-Amplifier and Modulator

The servo-amplifier and modulator are shown in Fig. 3. The modulator is a balanced type with two adjusting potentiometers used to minimize 60-cps amplifier output with the input shorted. The servo-amplifier is based on a circuit developed by H. C. Lin,[6] and is capable of supplying 110 volts rms to a Norden-Ketay size 15, 60-cps, servomotor. The over-all gain of the servo-amplifier and modulator is 170. In this application, the fixed field of the servomotor is operated at the reduced voltage of 80 volts to minimize heating of the motor. This reduced excitation causes maximum motor speed to be lowered; a no-load speed of 2400 rpm is used in system calculations. The bias-adjusting potentiometer is set for equal voltages across the output transistors. A single 15-volt power supply is used for the entire system; surge currents will reach 1.5 amp, but normal operating current will be less than 0.2 amp.

System Calculations

The last item to be determined is the gear ratio between servomotor and phase shifter. A maximum limit may be established by the accuracy of the master oscillator of the local time standard. The poorest oscillator to be used at the Minitrack stations will be no worse than one part in 10^8 oscillator-frequency drift per day. This oscillator drift is corrected by daily adjustment to hold oscillator-frequency accuracy to within one part in 10^8, corresponding to a time accuracy of 1100 μsec per day. The phase shifter operates at the 10-kc frequency, and shifts time 100 μsec for each shaft revolution. Therefore, the resolver might have to be moved eleven revolutions per day or approximately 0.008 rpm. The maximum gear ratio is servomotor no-load speed divided by maximum resolver speed, or 2400/0.008;

[6] H. C. Lin, "Quasi-complementary transistor amplifier," *Electronics*, vol. 29, pp. 173–175; September, 1956.



* Matched diodes 1N351's
★ Matched diodes 1N351's
All resistors are 1/2 watt unless noted.
All capacitors are microfarads unless noted.
MOTOR-NORDEN-KETAY—105 EIC—SIZE 15—60 cps.

Fig. 3—Schematic diagram of servo-amplifier.

300,000 to 1 is the maximum gear ratio that could be used with the above parameters.

The minimum gear ratio can be determined from the desired system bandwidth. A paper by McAleer[7] provides the basis for the following discussion. System bandwidth is given by the product of voltage-controlled oscillator coefficient and phase-detector efficient. The voltage-controlled oscillator coefficient is K_1:

$$K_1 = K_a K_m K_r \frac{1}{n};$$

where

$$K_a = 170 \frac{\text{volts}}{\text{volt}} \text{ (servo-amplifier)},$$

$$K_m = 0.33 \frac{\text{rps}}{\text{volt}} \text{ (servomotor)},$$

$$K_r = 4\pi \frac{\text{radians}}{\text{revolution}} \text{ (resolver)},$$

$$n = \text{gear ratio}.$$

The phase detector coefficient is K_2 and has a nominal value of two volts per radian.

$$\omega_c = K_1 K_2 = (170)(0.33)(4\pi)(2)(1/n);$$

$$\omega_c = 1400/n.$$

The system bandwidth should be small to minimize noise, but large enough to avoid unnecessary lags in achieving synchronization. An integration time ($T_c = 1/\omega_c$) of 45 seconds was chosen as a reasonable compromise between the two extremes.

$$\omega_c = \frac{1}{45} = \frac{1400}{n};$$

therefore, $n = 63{,}000$, and system bandwidth is approximately 0.003 cps. The actual gear ratio used was 61,500. The time constant of the low-pass filter will cause negligible tendency towards system oscillation if its value is less than one-fourth the system integration time of 45 seconds. The value for the low-pass filter was chosen to be 2.7 seconds as a convenient maximum; the capacitor was 200 μf, and the equivalent series resistance to the capacitor was 13,500 ohms.

[7] H. T. McAleer, "A new look at the phase locked oscillator," Proc. IRE, vol. 47, pp. 1137–1143; June, 1959.

The choice of the various gain coefficients to obtain a desired system bandwidth should be made to require a large gear ratio. A large gear ratio will in turn require large amplifier gains and create a "tight" system—a system which will have a minimum hysteresis and dead zone.

Results

The described system has been used with success on both NBA and WWVL. The noise as indicated by a record of phase-shifter shaft position when used with WWVL is approximately 0.1 μsec rms, and somewhat less for NBA. The time required to recover 63 per cent of a 5-μsec offset from NBA was approximately 35 seconds, indicating close agreement with the calculated 45 seconds.

Additional features will be added to this system to provide automatic shut-down in the absence of a carrier, and switch tuning to selected stations instead of plug-in filters; and a concerted effort will be directed towards some method of obtaining time, as well as frequency, synchronization. The phase shifter will be changed to signal frequency and removed from the time standard in order to avoid forcing the time standard through the diurnal phase variations which may be several tens of microseconds. This will permit several standard-frequency receivers to operate from one time standard in order to monitor simultaneously WWVL, NBA, and GBR.

Conclusions

The use of available standard-frequency VLF transmissions makes possible the synchronization of field stations throughout the world to a frequency accuracy of one part in 10,000,000,000. Known phase shifts occurring day-to-night can be used to correct the phase of the VLF carrier to an accuracy of a few microseconds. Considerable work remains to be done in developing a time synchronization system that will take full advantage of VLF phase angle stabilities.

Acknowledgment

The author is indebted to A. J. Rolinski and C. A. Schroeder for assistance in the development of the receiving system herein described, and to J. O. Williams for constructing the electronic chassis and preparing the final drawings. The project would not have been successful without the assistance and suggestions of many others, specifically the members of the Time Measurement Branch of the Goddard Space Flight Center.

Exact Solution of a Time-Varying Capacitance Problem*

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Summary—By means of a new method, a closed-form solution is obtained for the harmonics generated by a sinusoidally varying capacitance in series with a fixed resistor and battery. The solution describes the behavior of the condenser microphone, the vibrating-reed electrometer, a vibrating plate contact potential measuring apparatus, and a special loudspeaker improvement. With only minor modifications the solution can also apply to the case of a sinusoidally varying resistance in series with a fixed inductance and battery; thus, it may, in addition, be usedl to caculate the response of a carbon microphone. The present large-signal solution, which applies for any finite values of the modulation index and frequency, is compared with previous small-signal approximate results, and the dependence on modulation index and frequency is investigated for such quantities as output waveform, total harmonic distortion, harmonic amplitude and phase, and average input and output power. A very distorted waveshape is obtained for low relative frequencies and values of the modulation index near and including unity.

INTRODUCTION

FEW time-varying circuit problems have been solved to yield exact expressions for the harmonic components and thus, to allow their large-signal behavior to be investigated. With the current general interest in parametric amplifiers, such problems are becoming of more importance. Parametric amplifiers generally involve time-varying components, such as capacitors, in circuits which involve both inductive and capacitive energy storage. Exact large-signal analysis of such systems is very difficult and is not attempted herein. Instead, we shall be concerned only with the simpler problem of capacitive energy storage and shall show that here, at least, it is possible to give an exact solution in closed form.

Fig. 1 shows a circuit in which the center plate of a double capacitance can be moved by an outside force. We shall be concerned only with the case in which the equilibrium position of the center plate is such that $(C_1)_0=(C_2)_0\equiv C_0$, where the zero subscripts denote equilibrium. In addition, we shall take $R_1=R_2\equiv R$ and $C_3=C_4$. In the resulting antisymmetrical push-pull circuit there is no interaction between the top and bottom circuit halves, and initial attention can therefore be restricted to the top, or single-ended, half alone. Finally, it will be assumed that the restoring force acting on the center electrode when it is displaced from

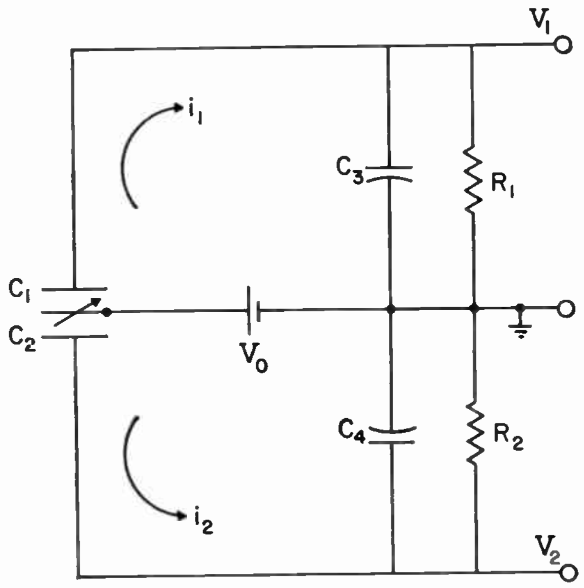

Fig. 1—Circuit diagram for time-varying double capacitor.

equilibrium is proportional to the displacement, so that the system is mechanically linear. For the present analysis we shall focus attention on the electrical part of the problem, as shown in Fig. 1, and shall not be concerned with mechanical impedances and the details of electromechanical coupling between the movable plate and the outside world.

When the movable capacitor plate is driven sinusoidally, the resulting time-varying current which flows in the circuit of Fig. 1 will not generally be sinusoidal but will contain harmonics of the driving signal. Such harmonic generation, while similar to that which arises in a nonlinear circuit, occurs here in a linear time-varying system which obeys a linear differential equation and satisfies the principle of superposition. Harmonics are produced here because of the time-varying capacitance and not principally because of the inverse dependence of capacitance on electrode spacing.

The circuit of Fig. 1 can represent a variety of devices of physical interest. First, it can be used as a representation of a single-ended or push-pull condenser microphone.[1] It can also be used to analyze the behavior

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[1] F. V. Hunt, "Electroacoustics," John Wiley & Sons, Inc., New York, N. Y., p. 170; 1954.
H. F. Olson, "Acoustical Engineering," D. Van Nostrand Co., Inc., Princeton, N. J., p. 253; 1957.

of a capacitance type of displacement probe.[2] In addition, it applies to the vibrating plate method of contact potential measurement[3,4] and to the vibrating reed electrometer.[5] As shown in Appendix VI, the present analysis, with relatively few changes, can also be used for a treatment of the carbon microphone where a time-varying resistance is in series with a time-independent inductance. Finally, the treatment applies as well to the special loudspeaker discussed below.

The magnetic loudspeaker is one of the weakest links in the high-quality reproduction of sound. Its performance has sometimes been somewhat improved by negative feedback derived from an auxiliary voice-coil winding and applied around the driving amplifier. This approach is only partially successful, especially for heavy-coned, low-frequency loudspeakers, because the voice-coil current has only partial control over cone motion and is not, therefore, a true measure of the output sound. More ideal control of cone motion can be obtained by metallizing the cone and making it the center electrode moving between two fixed metal-screen electrodes in front of and behind the cone. If bias is applied as shown in Fig. 1, the motion of the cone will generate a push-pull output signal between electrodes 1 and 2 which can be used for negative feedback. This signal will be a better measure of average cone motion and sound output than any that could be derived from the voice coil. Using it for negative feedback will result in flatter frequency response, lower nonlinear distortion, and possibly even some improvement, because of averaging, in the deleterious effects of cone breakup when it occurs. Note that the above arrangement is, in some sense, the inverse of the usual push-pull electrostatic loudspeaker where electric forces are used to move the center membrane instead of the magnetic forces of the present system. Although the same electrostatic forces exist in the present situation, they are negligible compared to the magnetic driving forces. After the above speaker improvement system was thought of by one of the present authors, a patent describing a single-ended version of the device was discovered.[6] It will be shown later that the push-pull system without feedback can exhibit much less nonlinear distortion generation than the single-ended system.

In the present analysis of the circuit of Fig. 1, we shall be concerned with the simplest case, that of sinusoidal driving force, such as that occurring when a condenser microphone is exposed to a single-frequency sound source. There have not been many treatments of the present problem, and none has been carried to such a stage that it is practical to calculate the high-order harmonics which are of importance at low relative frequencies and high values of the modulation index, m. Wente[7] analyzed the condenser microphone in 1917 and gave results valid for the fundamental response at low m and high relative frequencies only. Since then, the most ambitious treatment of the problem seems to have been that of Anderson and Alexander.[4] They have dealt with the cases where there is no parallel fixed capacitance C_3 across R and where C_3 is nonzero, but their analysis of the latter situation is incorrect. As we shall show later, such capacitance can usually be made negligible in practice, and it will be neglected in much of the present work because it considerably complicates the analysis.

Analysis

Consider the top half of Fig. 1 only, with $i_1 \equiv i$, $C_1 \equiv C$, and $R_1 \equiv R$. The basic equation to be solved is then

$$\frac{dq}{dt} = \frac{V_1}{R} + C_3 \frac{dV_1}{dt}, \tag{1}$$

where q is the instantaneous charge on C and $V_1 = V_0 - (q/C)$. Eq. (1) can be manipulated to yield

$$\frac{dq}{dt} + \frac{q}{R(C+C_3)}\left[1 - \frac{C_3 R}{C}\left(\frac{dC}{dt}\right)\right] = \left(\frac{C}{C+C_3}\right)\frac{V_0}{R}. \tag{1'}$$

The quantity we wish to calculate is the steady-state value of i/i_0, where $i_0 \equiv V_0/R$. This quantity can be written from (1′) as

$$(i/i_0) = \left(\frac{C}{C+C_3}\right) - \frac{q}{V_0(C+C_3)}\left[1 - \frac{RC_3}{C}\left(\frac{dC}{dt}\right)\right], \tag{1''}$$

which can be calculated when $C(t)$ and $q(t)$ are known.

Eq. (1′) may be formally integrated by means of an integrating factor when $C(t)$ is specified. The result involves rather unwieldy integrals however, and further analysis will be carried out here only for the simpler case, for which $C_3 = 0$. Then, a steady-state solution for i/i_0 is of the form

$$\left(\frac{i}{i_0}\right) = 1 - \frac{1}{RC}\exp\left[-\frac{1}{R}\int\frac{dt}{C}\right] \cdot \int \exp\left[\frac{1}{R}\int\frac{dt}{C}\right]dt. \tag{2}$$

[2] R. D. Shattuck, "Capacitance-type displacement probe," *J. Acoust. Soc. Am.*, vol. 31, pp. 1297–1299; October, 1959.
[3] W. A. Zisman, "A new method of measuring contact potential differences in metals," *Rev. Sci. Instr.*, vol. 3, pp. 367–370; July, 1932.
[4] J. R. Anderson and A. E. Alexander, "Theory of the vibrating condenser converter and application to contact potential measurement," *Australian J. Appl. Sci.*, vol. 3, pp. 201–209; September, 1952.
[5] H. Palevsky, R. K. Swank, and R. Grenchik, "Design of dynamic condenser electrometer," *Rev. Sci. Instr.*, vol. 18, pp. 298–314; May, 1947.
[6] G. H. Brodie, U. S. Patent No. 2,857,461; October 21, 1958.
[7] E. C. Wente, "A condenser transmitter as a uniformly sensitive instrument for the absolute measurement of sound intensity," *Phys. Rev.*, vol. 10, pp. 39–63; July, 1917.

Further progress requires knowledge of the time variation of C. We shall assume that the spacing between the plane-parallel plates of C is given by $d=d_0(1+m \sin \omega t)$ for an input driving frequency of $(\omega/2\pi)$. Here m is a modulation factor usually satisfying $0 \leq m \leq 1$. Then, neglecting fringing effects, taking rigid capacitor plates, and assuming that the driving frequency is sufficiently low that Maxwell's equations need not be invoked, one may write

$$C = C_0/(1 + m \sin \omega t). \tag{3}$$

For simplicity, let us now introduce the new variables $\phi \equiv \omega t$, $\beta \equiv 1/RC_0$, $z \equiv \omega/\beta$, $y \equiv 1/z$, and $x \equiv my = m/z$. Also let $M \equiv (1+m \sin \phi) \equiv C_0/C$. Note that z is a normalized frequency variable. Eq. (2) may now be simplified with the help of (3) to yield

$$\left(\frac{i}{i_0}\right) = 1 - yMe^{-y\phi + x \cos \phi} \int e^{y\phi - x \cos \phi} d\phi. \tag{4}$$

The integral in (4) cannot be carried out explicitly to yield i/i_0 in closed form. It will be shown, however, that closed expressions for the fundamental and harmonic components of i/i_0 can be obtained.

When $x \ll 1$, one can expand the exponentials involving $x \cos \phi$ in (4) in a simple power series. The integration can then be carried out and the result simplified to yield the fundamental and harmonic current components. When this procedure is applied in general, it is found that the harmonics, far from appearing in closed form, must be calculated from the product of two double series. In Appendix I, the results of this approach are given to the second order in m and up to second harmonic terms only.

Another method of handling (4) is to use the expansion[8]

$$e^{\pm x \cos \phi} = \sum_{s=0}^{\infty} \epsilon_s (\pm 1)^s I_s(x) \cos (s\phi), \tag{5}$$

where $\epsilon_0 = 1$, $\epsilon_s = 2 (s > 0)$, and $I_s(x)$ is a modified Bessel function of the first kind. When (5) is used, i/i_0 may be expressed as the product of two series or as a double series.[9] Finally, each harmonic current component can be expressed as a single infinite series of modified Bessel functions. Such reduction is very laborious, and the resulting series are only rapidly convergent for small x. The zero-order harmonic component of i/i_0 turns out to be

$$\left(\frac{i}{i_0}\right) = 1 - \sum_{s=0}^{\infty} (-1)^s \epsilon_s I_s^2(x) \qquad (0 \leq x < \infty). \tag{6}$$

[8] W. J. Cunningham, "Introduction to Nonlinear Analysis," McGraw-Hill Book Co., Inc., New York, N. Y., p. 248; 1958.

[9] Since the present analysis was completed, the treatment of Anderson and Alexander[1] has been discovered. It makes use of (5), but a double series is formally avoided since the authors Fourier analyze their single-series results separately to obtain harmonic components.

Since the sum of the series may be shown to be unity, there is no static component of current, which is in agreement with the fact that a direct current cannot flow through a capacitance, even when it is varying with time as in the present case.

Another approach, and the one we shall follow in detail here, is to Fourier analyze the steady-state part of (4) directly in order to obtain closed expressions for the harmonic current components. Before Fourier analysis can be applied, the steady-state current must be expressed in terms of a definite rather than an indefinite integral. Such transformation is carried out in Appendix II with the result

$$\left(\frac{i}{i_0}\right) = 1 - \frac{yMe^{x \cos \phi}}{(e^{2\pi y} - 1)} \int_0^{2\pi} e^{y\mu - x \cos(\phi + \mu)} d\mu. \tag{7}$$

Next, we wish to express i/i_0 in the complex Fourier series

$$\left(\frac{i}{i_0}\right) = \sum_{n=-\infty}^{\infty} c_n e^{in\phi} = \frac{a_0}{2} + \sum_{s=1}^{\infty} \{a_s \cos s\phi + b_s \sin s\phi\}, \tag{8}$$

where $c_n \equiv (a_n - ib_n)/2$.

The complicated calculation of the complex coefficients c_n is carried out in Appendix III. The final closed-form results are

$$c_0 = 0, \tag{9}$$

$$c_n = \frac{in\pi}{\sinh \pi y} I_{iy}(x) \cdot I_{n-iy}(x). \qquad (n > 0). \tag{10}$$

Eq. (10) is difficult to use directly for numerical calculations because of the imaginary and complex orders of the modified Bessel functions appearing in it. As shown in Appendix IV, however, recursion relations may be established between the complex and real harmonic coefficients of different orders. These relations allow the coefficients for any harmonic order to be calculated provided those for the two adjacent orders are known. One simple way of obtaining such initial starting coefficients is to calculate them directly from the power series expansion of (10). The necessary results are developed in Appendix V. Once a_1, a_2, b_1, and b_2 are calculated, the recursion relations of Appendix IV allow coefficients of higher orders to be obtained quite simply. Although the calculation of the initial a's and b's requires series evaluation, the series are far simpler than those obtained by the other methods of solution discussed briefly above, and the convergence of the present series is such that they are useful for much higher x values than could be treated practically by other methods.

The quantity $z = y^{-1} = RC_0\omega$ is a normalized frequency variable proportional to the ratio of the time constant of the undisturbed system to the period of the driving force. In addition, the following symbols will be used in the next section. Each harmonic component of i/i_0 appears in the form

$$h_n(\phi) \equiv a_n \cos n\phi + b_n \sin n\phi \qquad (11)$$
$$= \alpha_n \sin(n\phi + \chi_n),$$

where

$$\alpha_n \equiv (a_n^2 + b_n^2)^{1/2}, \qquad (12)$$

and

$$\chi_n \equiv \sin^{-1}(a_n/\alpha_n). \qquad (13)$$

In addition to the harmonic amplitude α_n, we shall also be interested in the normalized amplitude $\gamma_n \equiv (\alpha_n/\alpha_1)$. The total harmonic distortion (THD) given by

$$\mathrm{THD} \equiv \left[\sum_{r=2}^{\infty} \alpha_r^2\right]^{1/2} \Bigg/ \left[\sum_{r=1}^{\infty} \alpha_r^2\right]^{1/2} \qquad (14)$$

is likewise a quantity of interest. When the entire circuit of Fig. 1 is operated in the push-pull mode with $C_1 = C_2$ and $R_1 = R_2$, the symmetry of the arrangement is such that no even-order harmonics appear between the 1–2 terminals. In this case, it is pertinent to define the modified total harmonic distortion factor (MTHD) by

$$\mathrm{MTHD} \equiv \left[\sum_{r=2}^{\infty} \alpha_{2r-1}^2\right]^{1/2} \Bigg/ \left[\sum_{r=1}^{\infty} \alpha_{2r-1}^2\right]^{1/2}, \qquad (15)$$

an expression which involves odd harmonics only.

Using an IBM 650 digital computer, (49)–(52), and (59) in Appendixes V and VI have been summed for values of z and m that are of interest. In such summation, additional terms of the series are calculated until a term is reached which is sufficiently small to cause no change, within the eight-figure precision of the computer, in the partial sum to that point. This procedure, which yields sums of maximum computer accuracy, is necessary because the recursion relations (42) and (43) require starting values as correct as possible to allow accurate higher order harmonic components to be calculated.

Discussion of Results

Since the series for the harmonic components are convergent for any finite value of x, they can be used for very large x values, which can correspond to high values of m and low values of z, the normalized frequency. Although $m=1$ is not usually a useful value for the physical devices discussed in the Introduction, it is found that there is a smooth transition from $m=0.99$ to $m=1$, and it is therefore convenient to consider this limiting case. The pertinent series converge, in fact, for $m>1$; so the limitation $m<1$, when pertinent, is physical, not mathematical. It would be possible to use an analog computer to represent the circuit of Fig. 1 in such a way that negative capacitances were realized. In this case, m could exceed unity, and the present mathematical results would still apply. However, since the physical devices which the mathematical results describe are limited to $m<1$ or $m \leq 1$, the numerical calculations leading to the results of the present section have been also limited to the range $0<m\leq 1$.

Fig. 2 shows how the distortion factors depend on frequency for various values of m. It will be noted that for $z \ll 1$ both THD and MTHD approach limiting values which, in the case of THD, are very nearly equal to m. Thus, for example, no matter how low the frequency, the maximum total harmonic distortion for $m=0.01$ is one per cent. For $z \gg 1$, both THD and MTHD decrease as the frequency increases with limiting slopes of -1 and -2, respectively. As expected, MTHD is always less than THD even at very low frequencies.

Fig. 3 presents the distortion factors as a function of m with z the parameter. These graphs show clearly that only for high values of m near unity can decreasing z below 0.1 make any very appreciable difference in THD and MTHD. Such decrease, however, can change the harmonic constitution considerably. The limiting slopes in Fig. 3(a) are unity, while those in 3(b) are equal to two. The dotted lines show the linear extrapolations of the curves.

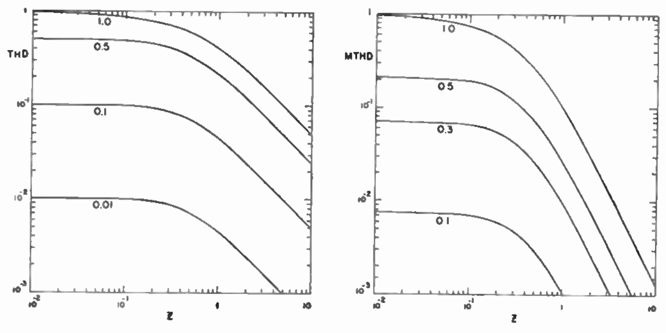


Fig. 2—Harmonic distortion factors, THD and MTHD, as functions of normalized frequency z for various values of the modulation index m.

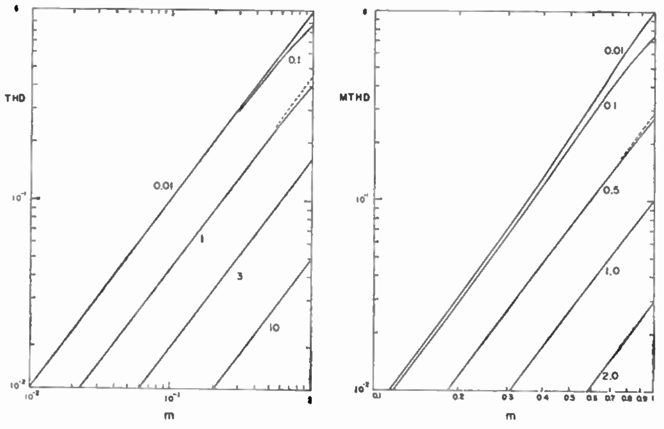


Fig. 3—THD and MTHD vs m for various values of z.

For most of the practical devices to which the present analysis applies, it is desirable to operate under conditions which minimize harmonic distortion. Fig. 4 is drawn for THD and MTHD values of one per cent and shows how m and z must be interrelated to maintain these values. To the right of each curve the distortion will be less than one per cent. Clearly, for a given z, m may be much higher for a total push-pull harmonic distortion of one per cent than for a single-ended total harmonic distortion of the same value. The limiting slopes in this figure are both two.

Another quantity like THD or MTHD which is determined by the entire spectrum of harmonics is the rms relative wave amplitude. We shall actually plot the amplitude

$$A = \left(\sum_{n=1}^{\infty} \alpha_n^2 \right)^{1/2}$$

which is $\sqrt{2}$ times the rms amplitude. The quantity A reduces to the zero-to-peak amplitude of the wave only when a single sinusoidal component is present. Thus, for large z, it approaches α_1 which, in turn, approaches m. For push-pull operation we shall take A as

$$\left(\sum_{r=1} \alpha_{2r-1}^2 \right)^{1/2}$$

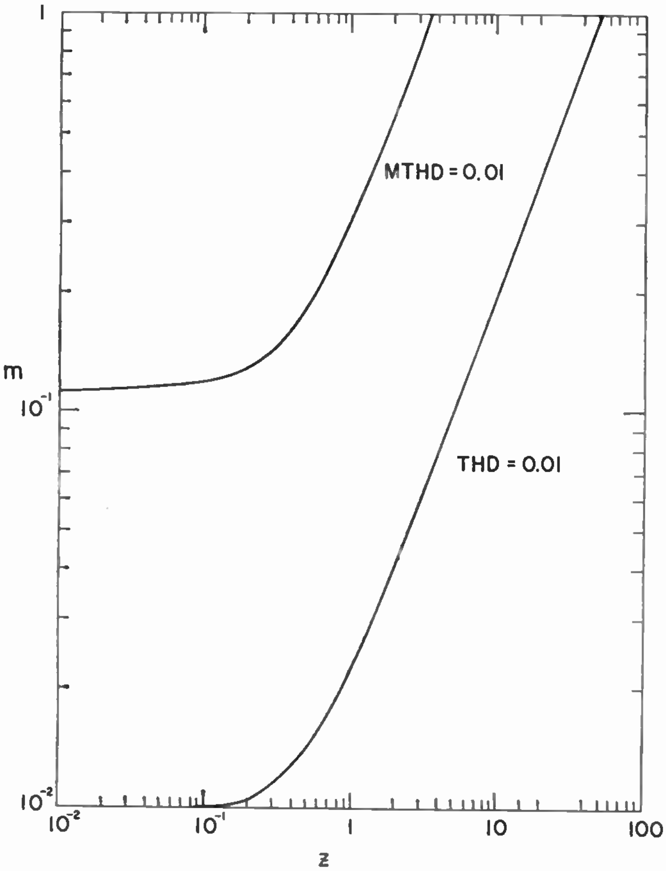


Fig. 4—Interrelation between m and z necessary for MTHD and THD to remain constant and equal to 0.01.

for convenient comparison with that for single-ended operation. Fig. 5 shows how A depends on z for $m = 0.5$ and 1. The limiting slopes for the $m = 0.5$ curves are unity, the usual 6 db/octave slope to be expected for a capacitative reactance. Note that when the single-ended and push-pull curves are very close together, only the fundamental is of importance.

The equations for all the above quantities which depend on sums of harmonics have been written with an infinite upper limit. In practice, as the harmonic index n increases, one eventually reaches a region where higher harmonic amplitudes are decreasing so rapidly that further harmonics add nothing appreciable to the series. In the machine calculations, summation of the series are always carried to this point even when n values as high as 25 are required.

Fig. 6 shows how the normalized harmonic amplitudes depend on the order of the harmonic for various m and z values. We have connected the calculated

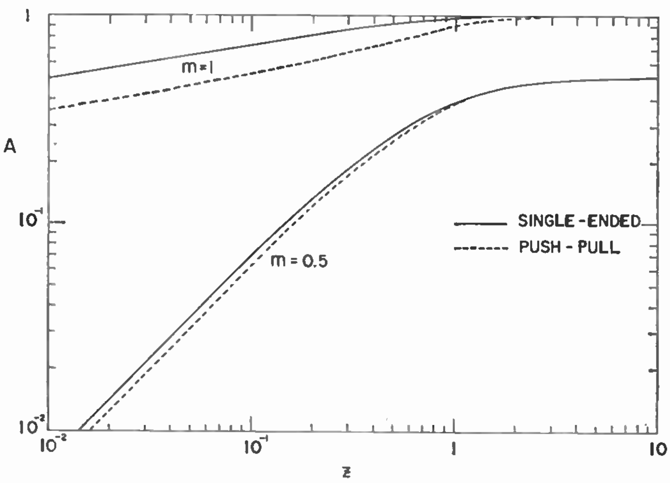


Fig. 5—Dependence on normalized frequency of the single-ended and push-pull amplitudes, A, for $m = 0.5$ and 1.

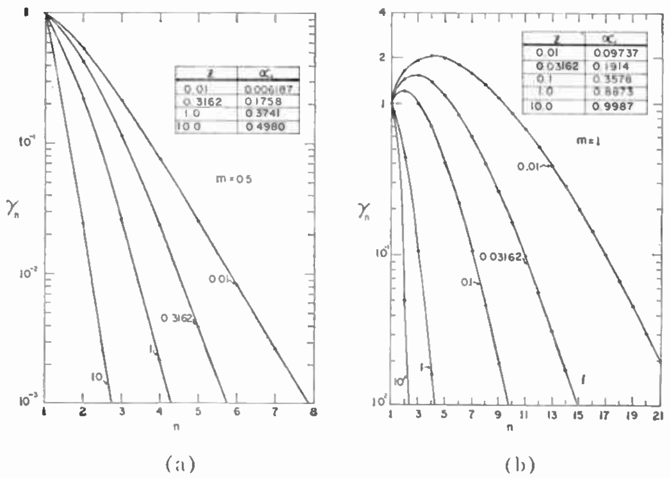


Fig. 6—Dependence of normalized harmonic amplitudes, $\gamma_n = \alpha_n / \alpha_1$, on harmonic index, n, for $m = 0.5$ and 1 and various z values.

points with light lines for convenience, but only the dots themselves are significant. Also shown in the tables are the fundamental amplitude values, α_1, for the various z values considered. For $z \gg 1$, α_1 approaches m. As expected, the harmonic amplitudes decrease very rapidly when z is unity or greater. When $z=1$, $\omega RC_0=1$; so $z=1$ is a natural dividing point. When $z \gg 1$, the period of the driving force is much smaller than the natural time constant RC_0. Under these conditions, the charge on the variable capacitance cannot change appreciably within a period, and the instantaneous voltage across the capacitor will be proportional to $1/C$ and will thus involve the fundamental component only. In the limit of high frequencies, the variable capacitor charge q will remain virtually constant and there will be no harmonic generation.

For $n \geq 3$, the harmonics in Fig. 6 have been calculated using the recursion relations of Appendix IV. These relations eventually involve small differences between large numbers and, as n increases, harmonic coefficient accuracy will eventually become impaired. With the eight significant figures available on the 650 machine, this point is reached when γ_n has decreased somewhat below 0.01. The value of γ_n which is still accurate is still more than sufficiently small so that the sums involving $\alpha_n{}^2$ converge excellently.

Anderson and Alexander[4] have been able to apply their technique for solving the present problem to m values as high as 0.667 and to z values as small as 0.222 ($x=3$). In this case, they obtained γ_n values of 63, 26, 9, and 1 per cent for $n=2$, 3, 4, 5, respectively.[10] For the same input, the present analysis yields 63.3, 26.7, 8.8, 2.4, and 0.58 per cent for n from 2 to 6. This is relatively good agreement and affords a check of both methods of solution.

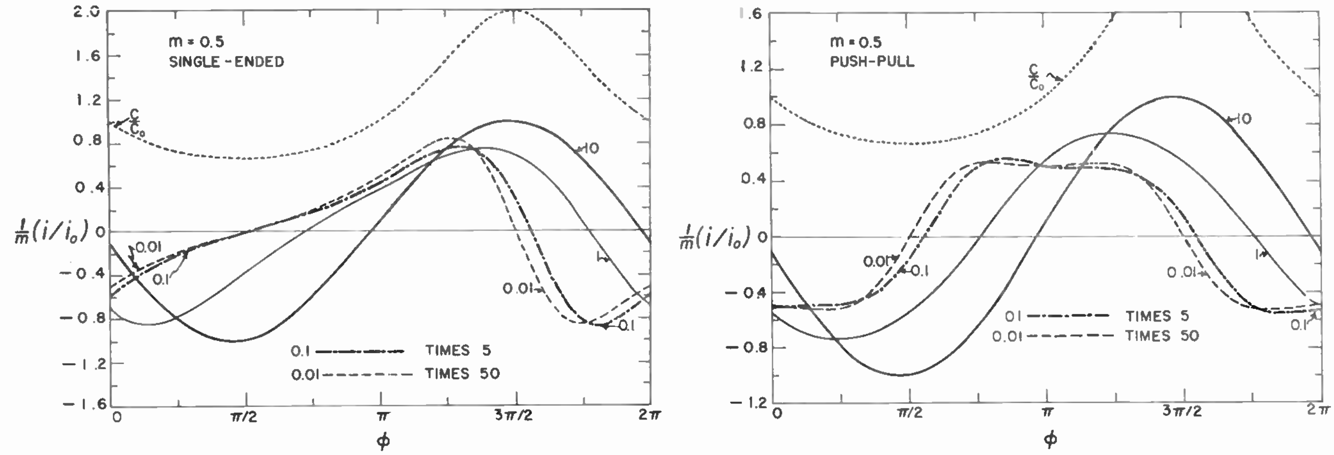
An interesting feature of Fig. 6(b) is the rise of some of the higher harmonic amplitudes above the amplitude of the fundamental. This behavior occurs to a smaller degree as well for m values of 0.9 but has disappeared by $m=0.7$. Curves for $m=1$ and z, considerably less than 0.01, could not be obtained with the present 650 calculation program because it was limited to a maximum of 100 terms in each of the series of Appendixes V and VI. Some idea of how many terms in these series were required is given by the following data: for $m=1$, the following z values: 10, 1, 0.3162, 0.1, 0.03162, and 0.01 required a maximum of 3, 7, 11, 21, 40, and 82 terms, respectively; smaller values of m of course needed fewer terms.

The harmonic coefficients a_n and b_n can be recombined when known to yield the Fourier series of (8) which allows i/i_0 to be plotted as a function of ϕ. The resulting waveshapes for various values of z are shown in Fig. 7 (next page) for $m=0.5$ and in Fig. 8 for $m=1$. We have actually plotted $(1/m(i/i_0))$ rather than (i/i_0) in order to facilitate comparison between the two figures. The $(1/m)$ factor causes the fundamental signal components to have the same amplitude at high frequencies (*eg.*, $z \geq 10$) independently of the value of m. Also shown in these figures are dotted curves of (C/C_0) or $(C/10C_0)$ which indicate how the normalized capacitance varies through a cycle.

Fig. 7 shows that at high relative frequencies the current is in phase with the capacitance, and, even at $z=10$, there is little distortion of the waveshape and very small phase shift. The situation is considerably changed as z decreases, however, and the harmonic components shown in Fig. 6 begin to play an important role. Note that the decrease in amplitude shown in Fig. 5 has been partly compensated in the curves for $z=0.1$ and 0.01 by multiplying the amplitudes by the factors shown. As z decreases, the most striking alteration is that the current changes from having a maximum at $\phi=3\pi/2$, the point where the capacitance is maximum, to going through zero at this point. In the low-frequency limit, the current curve thus tends to be proportional to the derivative of the capacitance.

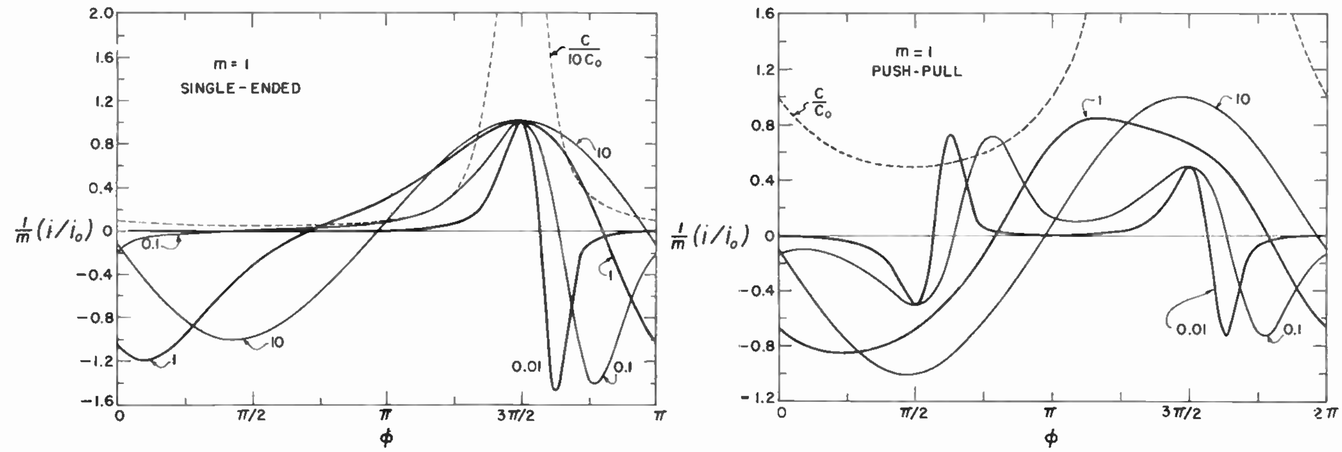
Somewhat similar results are shown in Fig. 8. For $m=1$, however, Fig. 5 shows that there is not a very appreciable decrease in the rms current amplitude as z decreases; thus, none of the curve amplitudes has been changed here. For $m=1$, the capacitance reaches infinity at $\phi=3\pi/2$. The equations show, however, that at this point there is no voltage across the capacitor and no charge on it. Hence, it is merely a short circuit and, at this value of ϕ, the current is limited only by the series resistance and must therefore be equal to i_0. This requirement is independent of the value of z. As mentioned in Appendix VII, the force between the capacitor plates never becomes infinite even for $m=1$. Except at very low relative frequencies, the force with $m=1$ will not, in fact, vary much over a cycle. The stiffness of the suspension of the moveable plate need only be great enough to balance the static electrical attractive force and give the desired spacing, d_0, when $V_1=0$ and V_0 is equal to the applied value. Note that near $\phi=3\pi/2$ the capacitance somewhat approximates a delta function and the current approximates a doublet impulse function, the derivative of the delta function. Because of the requirement that $i=i_0$ at $\phi=3\pi/2$, the doublet cannot be equal to zero at $\phi=3\pi/2$, as in the $m=0.5$ case, except in the low-frequency limit. The short-circuit condition and the resulting waveshape near $\phi=3\pi/2$ are responsible for the slow decrease of the rms amplitude of i/i_0 for $m=1$ as compared to that for $m=0.5$. Note that Fourier analysis shows that the average value of (C/C_0) is $(1-m^2)^{-1/2}$. For $m=1$, this quantity reaches infinity, unlike the average value of a delta function which is finite. Little need be said about

[10] It should be noted that Anderson and Alexander have denoted by fundamental, first harmonic, second harmonic, etc. quantities which are usually (and in the present treatment) termed fundamental or first harmonic ($n=1$), second harmonic ($n=2$), etc.



(a) (b)

Fig. 7—Dependence of $(1/m)(i/i_0)$ on $\phi = \omega t$ for $m = 0.5$ and various z values. The dotted curve shows (C/C_0) vs ϕ.



(a) (b)

Fig. 8—Dependence of $(1/m)(i/i_0)$ on ϕ for $m = 1$ and various z values. The dotted curve shows $(C/10C_0)$ in (a) and (C/C_0) in (b).

the push-pull curves; their symmetry arises from the absence of all even-order harmonics.

The question of how well the simplest approximate expressions for the harmonic coefficients given in Appendix I represent the actual behavior of the system is of some interest. For $z = 0.01$ and 1, Figs. 9 and 10 show how the ratio of exact to approximate coefficients, (α_n/α_n^0), depends on z for $n = 1$, 2, 3. For the $z = 0.01$ case, the simple solution is only a good approximation for $m < 0.3$. Also, for $z = 0.01$ the higher the harmonic order the worse the approximation, while for $z = 1$ the reverse is true.

The results of Appendix I may also be used to compare approximate and exact phase predictions. In Fig. 11, the quantity $-\Delta\chi_n = \chi_n^0 - \chi_n$ is plotted vs m for $n = 1$ and 2 and three z values. In Fig. 12, the phase results are plotted vs z in different forms. In these graphs, solid lines denote positive and dotted lines negative quantities and, for convenience in plotting, all χ_n values have been diminished by 180°. First, the accurate values of χ_1 and χ_2 in degrees are plotted. In addition, the percentage deviation of the accurate values from the approximate values are shown. Note that very high deviations occur for χ_2 when $m = 1$. The open breaks in the $(100\Delta\chi_2/\chi_2^0)$ curves near $z = 0.6$ appear because in this region the signs of the approximate and accurate second harmonic phases are different.

Finally, Fig. 13 (page 461) shows how the zero frequency or dc harmonic amplitude in the carbon microphone case (Appendix VI) depends on modulation for various frequency values. One sees that in this case, where a dc component is allowed, the dc part of i can greatly exceed $i_0 = V_0/R_0$ when m is near unity and z is small. This is an interesting case of rectification without nonlinearity.

For low harmonic distortion yet appreciable m, z must be unity or greater. It is of interest to inquire what value of R is necessary to ensure that $z = 1$ at $f = 20$ cps in the modified loudspeaker discussed in the Introduction. Since the capacitance modification will be of most value for large, low-frequency speakers, we may consider a typical cone area of 1300 cm². If the fixed screens are 0.25 inch in front of and behind the cone, the single-ended equilibrium capacitance is about 181

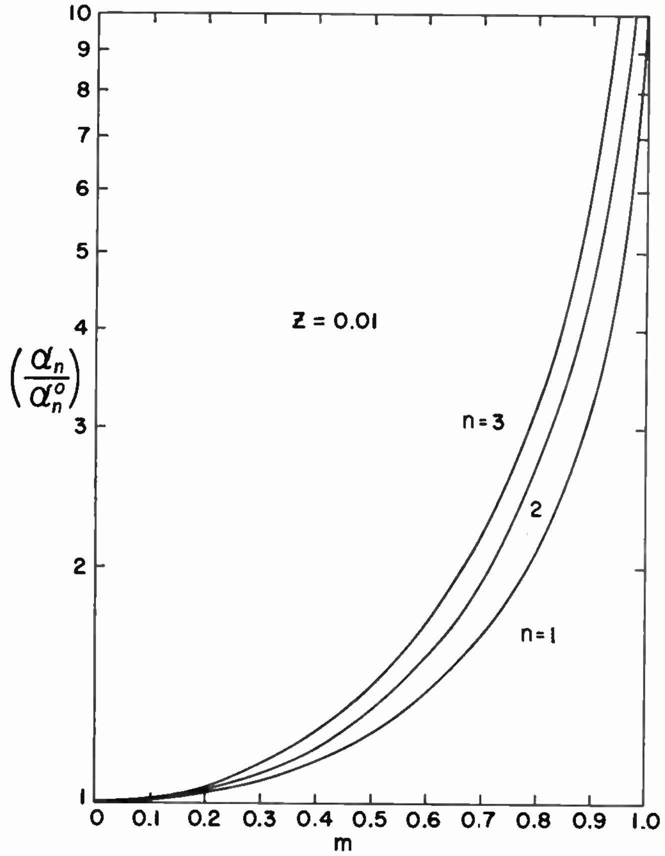


Fig. 9—Comparison of exact and approximate harmonic amplitudes as functions of m for $z=0.01$.

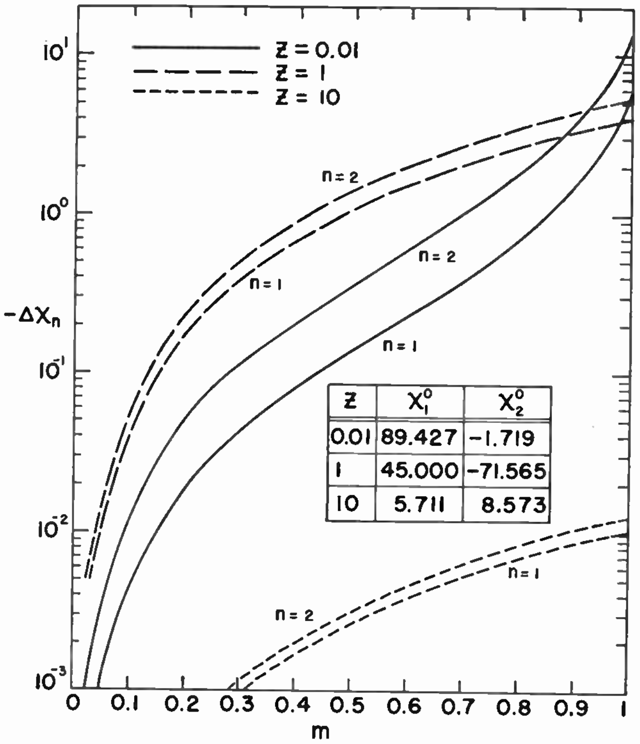


Fig. 11—Dependence of phase difference $-\Delta\chi_n=\chi_n^0-\chi_n$ on m for $n=1$ and 2 and various z values.

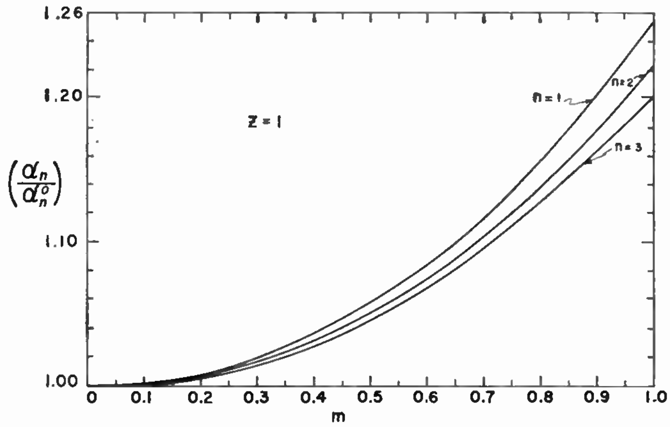
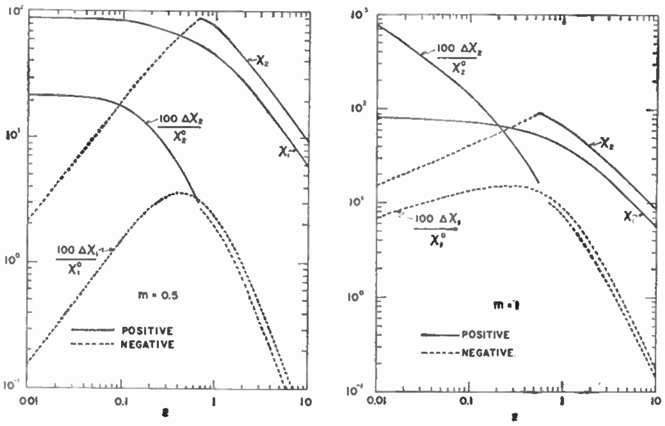


Fig. 10—Comparison of exact and approximate harmonic amplitudes as functions of m for $z=1$.



(a) (b)

Fig. 12—Dependence of χ_1 and χ_2 in degrees on z for $m=0.5$ and 1, and dependence on z of percentage differences between accurate and approximate phases.

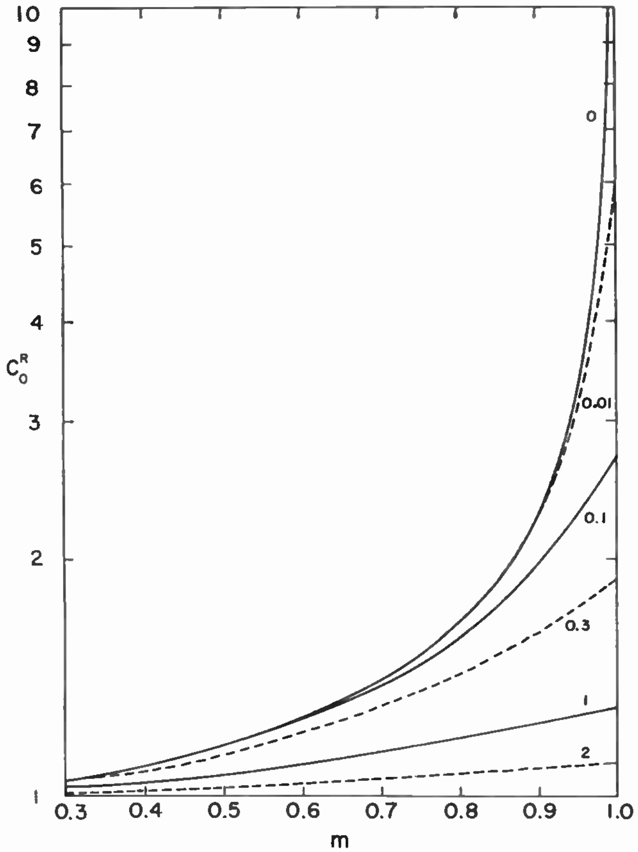


Fig. 13—Dependence of dc component of (i/i_0) in the carbon microphone case on m for several z values.

p.f. This leads to an R value of 44 megohms, an easily realized magnitude. Note that the present analysis applies only when R is much less than the leakage resistance of C. It was mentioned earlier that the capacitance C_3 in parallel with R would be neglected. Exact conditions which must be met by C_3 to justify such neglect have not been derived. It is clear, however, that one necessary condition is that $C_3 \ll C_0$ for any m. Further, if the load is to remain primarily resistive, it is essential for all frequencies of interest that the reactance of C_3 appreciably exceed the magnitude of R. If we require a capacitive reactance of 100 megohms at $f=1000$ cps, the parallel capacitance must be less than 0.016 p.f. This is a reasonably stringent requirement, but it can be met by fairly well-known feedback techniques[11,12] which make it possible to achieve an amplifier input impedance made up of a resistive component exceeding 10^9 ohms and virtually zero input capacitance over the audio frequency range.

In Appendix VII, expressions are derived for the instantaneous values of the vibrating capacitor charge, voltage, stored energy, power dissipated in the vibrating capacitor, attraction between plates, input power, and the power developed in the load resistor, R. In addition, a general relation, (69), between the instantaneous powers in the system is established. The behavior of the capacitor voltage, V_c, can be obtained directly from the results already presented for (i/i_0). The other quantities involve functions of the integral F_1 (Appendix II) which has not been evaluated in closed form. It is of interest, therefore, to calculate the time averages of these quantities where possible.

The average power output, equal to the average power input, is obtained in Appendix VII in a series form valid for arbitrary frequency. Averages of the other quantities may also be obtained as infinite series, but only the first few terms, applicable for $z \gg 1$, are calculated in the Appendix. It will be noted from (71) that the average output power, $\langle P_{out} \rangle$, equals $A^2V_0^2/2R$ where A is the ordinate of Fig. 5 and has been defined earlier. For $z \gg 1$, $\langle P_{out} \rangle$ approaches $m^2V_0^2/2R$. For $z=0.01$ and $m=1$, the results of Fig. 5 show that $\langle P_{out} \rangle \simeq V_0^2/8R$, indicating that the output power has not dropped off tremendously even at this low z value. The input power calculated in Appendix VII is the ideal minimum and involves only the power required to move the charged plates of the capacitor against the electrical forces involved. In practice, there will be unavoidable electromechanical conversion losses, but such power dissipation can often be made small.

The rich harmonic generation shown in Fig. 6(b) suggests that a vibrating capacitor device could be used for efficient high harmonic production. However, the conversion efficiency is lowered by the efficiency of whatever electromechanical, piezoelectric, electrostrictive, or magnetostrictive device is used to vibrate one of the capacitor plates, and the resulting over-all efficiency may not be comparable to that obtained with all-electric harmonic converters. When a voltage-dependent capacitance[13,14] is used in place of the mechanically driven capacitor, the nonlinearity of this device may possibly contribute even greater high harmonic generation if C_{min}/C_{max} can be made sufficiently small.

It is often desired to obtain high harmonics from a frequency-stabilized quartz crystal since the resulting harmonics will themselves be well frequency stabilized. The vibrating capacitor may possibly be useful here. Consider a quartz crystal vibrating in a longitudinal mode. One end of it is metallized and may be considered the vibrating plate of a capacitor. The quartz crystal is attached to a rigid rectangular C-shaped structure in such a way that the top of it vibrates very close to the top of the C, which can be the fixed plate of the capacitor. By forming this fixed capacitance of aluminum with a thin anodized insulating surface, values of m at

[11] J. R. Macdonald, "An ac cathode follower circuit of very high input impedance," *Rev. Sci Instr.*, vol. 25, pp. 144–147; February, 1954.

[12] J. R. Macdonald, "Some augmented cathode follower circuits," IRE Trans. on Audio, vol. AU-5, pp. 63–70; May–June, 1947.

[13] D. B. Leeson and S. Weinreb, "Frequency multiplication with nonlinear capacitors—a circuit analysis," Proc. IRE, vol. 47, pp. 2076–2084; December, 1959.

[14] L. J. Giacoletto and J. O'Connell, "A variable-capacitance germanium junction diode for UHF," *RCA Rev.*, vol. 17, pp. 68–85; March, 1956.

least as large as 0.95 should be achievable and operation with $m=1$ should also be possible on elimination of this layer. Operation with $z=0.01$ or below will then lead to the generation of harmonics of high order and accurately controlled frequency.

Appendix I

Power Series Expansion

To second order in m $(x=my)$, we may write

$$e^{\pm x\cos\phi} = 1 \pm x\cos\phi + (x^2\cos^2\phi)/2. \qquad (14)$$

The quantity i/i_0 may be expressed in general as the Fourier series given in (8) of the body of this work. Substituting (14) in (4), simplifying, and comparing with (8) yields

$$\left.\begin{aligned} a_0 &= 0 \\ a_1 &= -mz/(1+z^2) \\ a_2 &= \frac{-3m^2z^2}{(1+z^2)[1+(2z)^2]} \\ b_1 &= -mz^2/(1+z^2) \\ b_2 &= -\frac{m^2z(2z^2-1)}{(1+z^2)[1+(2z)^2]} \end{aligned}\right\}. \qquad (15)$$

The harmonic amplitudes $\alpha_n \equiv \sqrt{a_n{}^2+b_n{}^2}$ are, for $n=1$, 2, 3,

$$\left.\begin{aligned} \alpha_1{}^0 &= mz/\sqrt{1+z^2} \\ \alpha_2{}^0 &= m\alpha_1{}^0/\sqrt{1+(2z)^2} \\ \alpha_3{}^0 &= 3m\alpha_2{}^0/4\sqrt{1+(3z)^2} \end{aligned}\right\}, \qquad (16)$$

where the zero superscript indicates that the quantities in question are of lowest order in m. Note that as $z\to\infty$, $i/i_0\to -m\sin\phi$, the correct result in this limit.

Appendix II

Transformation of (4)

Let

$$i/i_0 \equiv 1 - yMF_1, \qquad (17)$$

where

$$F_1(\phi) \equiv e^{-y\phi+x\cos\phi}\int e^{y\phi-x\cos\phi}d\phi$$

$$= e^{-y\phi+x\cos\phi}\left[c+\int_0^\phi e^{y\lambda-x\cos\lambda}d\lambda\right]. \qquad (18)$$

In the last equation, c is an integration constant. Next, we wish to find the steady-state or periodic part of $F_1(\phi)$. We have

$$F_1(\phi+2\pi k) = e^{-2\pi yk}\left\{e^{-y\phi+x\cos\phi}\left[c+\int_0^\phi e^{y\lambda-x\cos\lambda}d\lambda\right]\right.$$

$$\left.+\,e^{-y\phi+x\cos\phi}\int_\phi^{\phi+2\pi k} e^{y\lambda-x\cos\lambda}d\lambda\right\}$$

$$= e^{-2\pi yk}\left[F_1(\phi) + e^{-y\phi+x\cos\phi}\sum_{s=0}^{k-1}\int_{\phi+2\pi s}^{\phi+2\pi(s+1)} e^{y\lambda-x\cos\lambda}d\lambda\right]$$

$$= e^{-2\pi yk}[F_1(\phi)+F_2(\phi)], \qquad (19)$$

where k is a positive integer.

On making the transformation $\mu\equiv\lambda-\phi-2\pi s$, the last integral becomes

$$\int_0^{2\pi} e^{y(\mu+\phi+2\pi s)-x\cos(\mu+\phi+2\pi s)}d\mu$$

$$= e^{y\phi+2\pi sy}\int_0^{2\pi} e^{y\mu-x\cos(\mu+\phi)}d\mu.$$

Thus,

$$F_2(\phi) = e^{x\cos\phi}\sum_{s=0}^{k-1} e^{2\pi sy}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu$$

$$= e^{x\cos\phi}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu\cdot\sum_{s=0}^{k-1} e^{2\pi sy}$$

$$= \left(\frac{e^{2\pi ky}-1}{e^{2\pi y}-1}\right)e^{x\cos\phi}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu$$

$$(y\neq 0). \qquad (20)$$

Substituting (20) into (19) with $a\equiv(e^{2\pi y}-1)^{-1}$ yields $F_1(\phi+2\pi k)$

$$= e^{-2\pi yk}\left[F_1(\phi) - ae^{x\cos\phi}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu\right]$$

$$+\,ae^{x\cos\phi}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu. \qquad (19')$$

If the last term is denoted $F_3(\phi)$, we have

$$F_1(\phi+2\pi k) = e^{-2\pi yk}[F_1(\phi)-F_3(\phi)] + F_3(\phi). \qquad (19'')$$

Since $F_3(\phi+2\pi k)=F_3(\phi)$, we may write

$$F_1(\phi+2\pi k) - F_3(\phi+2\pi k) = e^{-2\pi yk}[F_1(\phi)-F_3(\phi)]. \quad (21)$$

Now since $[F_1(\phi)-F_3(\phi)]$ is bounded for $0\le\phi\le 2\pi$ and $y>0$, (21) shows that F_1 approaches F_3 uniformly in ϕ for large ϕ. Thus, for large $t(\phi\to\infty)$

$$F_1(\phi)\to F_3(\phi) = ae^{x\cos\phi}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu. \qquad (22)$$

Applying (22) in (17) yields (7) of the text.

Appendix III

Fourier Analysis

Using the notation and results of Appendix II, the complex Fourier coefficients are given by

$$c_n = \frac{1}{2\pi}\int_0^{2\pi}[1-yMF_3]e^{-in\phi}d\phi$$

$$= \frac{1}{2\pi}\int_0^{2\pi} e^{-in\phi}d\phi - \frac{1}{2\pi}\int_0^{2\pi} yMF_3e^{-in\phi}d\phi. \qquad (23)$$

Thus, with

$$\frac{1}{2\pi}\int_0^{2\pi} e^{-in\phi}d\phi \equiv \delta_{0n} = 1 \ (n=0) \text{ and } 0 \ (n>0),$$

we may write

$$(\delta_{0n} - c_n)$$
$$= \frac{a}{2\pi}\int_0^{2\pi}(y + x\sin\phi)e^{-in\phi+x\cos\phi}\int_0^{2\pi} e^{y\mu - x\cos(\phi+\mu)}d\mu d\phi$$
$$= \frac{a}{2\pi}\int_0^{2\pi}(y + x\sin\phi)e^{-in\phi-y\phi+x\cos\phi}$$
$$\cdot\int_0^{2\pi} e^{y(\phi+\mu)-x\cos(\phi+\mu)}d\mu d\phi. \qquad (24)$$

Now let,

$$G(\phi) \equiv \int_0^{2\pi} e^{y(\phi+\mu)-x\cos(\phi+\mu)}d\mu, \qquad (25)$$

and

$$g(\phi) \equiv e^{-(y+in)\phi+x\cos\phi}. \qquad (26)$$

Integrating (24) by parts yields

$$(\delta_{0n} - c_n) = \frac{a}{2\pi}\left\{\left[-g(\phi)G(\phi)\right]_0^{2\pi} + \int_0^{2\pi} g(\phi)\frac{dG}{d\phi}d\phi - in\int_0^{2\pi} g(\phi)G(\phi)d\phi\right\}, \qquad (27)$$

where

$$\frac{dG}{d\phi} = \frac{d}{d\phi}\int_\phi^{2\pi+\phi} e^{y\lambda - x\cos\lambda}d\lambda = \left[e^{y(\phi+2\pi)-x\cos\phi} - e^{y\phi - x\cos\phi}\right]$$
$$= e^{y\phi - x\cos\phi}\left[e^{2\pi y} - 1\right]. \qquad (28)$$

The first term in (27) is zero and substituting (28) in (27) leads to

$$(\delta_{0n} - c_n) = \frac{1}{2\pi}\int_0^{2\pi} e^{-in\phi}d\phi - \frac{ina}{2\pi}\int_0^{2\pi} g(\phi)G(\phi)d\phi. \quad (29)$$

Thus, $c_0 = 0$ and for $n > 0$, one obtains

$$c_n = \frac{ina}{2\pi}\int_0^{2\pi} g(\phi)G(\phi)d\phi$$
$$= \frac{ina}{2\pi}\int_0^{2\pi}\int_0^{2\pi} e^{y\mu - in\phi + 2x\sin\mu/2\sin(\phi+\mu/2)}d\mu d\phi$$
$$= \frac{ina}{2\pi}\int_0^{2\pi} e^{y\mu+in\mu/2}d\mu\int_0^{2\pi} e^{-in(\phi+\mu/2)+2x\sin\mu/2\sin(\phi+\mu/2)}d\phi$$
$$= \frac{ina}{2\pi}\int_0^{2\pi} e^{y\mu+in\mu/2}d\mu\int_{\mu/2}^{\mu/2+2\pi} e^{-in\xi+2x\sin\mu/2\sin\xi}d\xi. \quad (30)$$

Since the second integral is periodic, we may write

$$c_n = \frac{ina}{2\pi}\int_0^{2\pi} e^{y\mu+in\mu/2}d\mu\int_0^{2\pi} e^{-i[n\xi+2ix\sin\mu/2\sin\xi]}d\xi,$$
$$= \frac{ina}{2\pi}\int_0^{2\pi} e^{y\mu+in\mu/2}J_n\left(-2ix\sin\frac{\mu}{2}\right)d\mu. \qquad (31)$$

On making the substitution $\mu/2 \equiv \pi/2 + \theta$,

$$c_n = 2inae^{y\pi}\int_{-\pi/2}^{\pi/2} e^{2y\theta+in\theta}e^{in\pi/2}J_n(-2ix\cos\theta)d\theta. \quad (32)$$

Since

$$I_n(x) = e^{-in\pi/2}J_n(xe^{i\pi/2}), \qquad (33)$$

and

$$J_n(xe^{-i\pi}) = e^{-i\pi n}J_n(x), \qquad (34)$$

we may write

$$e^{in\pi/2}J_n(-2ix\cos\theta) = e^{in\pi/2}e^{-in\pi}J_n(2ix\cos\theta)$$
$$= e^{-in\pi/2}J_n(2ix\cos\theta) = I_n(2x\cos\theta). \quad (35)$$

Thus,

$$c_n = 2inae^{\pi y}\int_{-\pi/2}^{\pi/2} e^{(2y+in)\theta}I_n(2x\cos\theta)d\theta. \qquad (36)$$

This result can finally be further simplified as follows:

$$c_n = 2inae^{\pi y}\int_0^{\pi/2}\left[e^{(2y+in)\theta} + e^{-(2y+in)\theta}\right]I_n(2x\cos\theta)d\theta$$
$$= \frac{2in}{\sinh\pi y}\int_0^{\pi/2}\cos\left[(n-2iy)\theta\right]I_n(2x\cos\theta)d\theta$$
$$= \frac{2in}{\sinh\pi y}\int_0^{\pi/2}\cos\left\{\left[(n-iy)-iy\right]\theta\right\} I_{(n-iy)+iy}(2x\cos\theta)d\theta. \quad (37)$$

Now, we may make use of the identity[15]

$$I_\mu(x)I_\nu(x) = \frac{2}{\pi}\int_0^{\pi/2} I_{\mu+\nu}(2x\cos\theta)\cos\left\{(\mu-\nu)\theta\right\}d\theta \quad (38)$$

to obtain the final result

$$c_n = \frac{in\pi}{\sinh\pi y}I_{iy}(x)I_{n-iy}(x). \qquad (n>0). \qquad (39)$$

Appendix IV

Recursion Relations

Using a recursion relation satisfied by $I_\nu(x)$, (10) may be written as ($n \geq 1$), hence

$$c_n = \frac{i\pi nx}{2(n-iy)\sinh\pi y}I_{iy}(x)\cdot\left[I_{n-1-iy}(x) - I_{n+1-iy}(x)\right]. \quad (40)$$

[15] G. N. Watson, "A Treatise On the Theory of Bessel Functions" Cambridge University Press, Cambridge, Eng.; 1944. See p. 150 for the equation expressed in terms of ordinary Bessel functions.

This result may be expressed in the form

$$c_n = \frac{nx}{2(n-iy)}\left[\frac{c_{n-1}}{n-1} - \frac{c_{n+1}}{n+1}\right], \qquad (40')$$

a recursion relation between the complex harmonic coefficients. Next we require similar relations between the real coefficients a_n and b_n. From (40′) we have

$$(a_{n+1} - ib_{n+1}) = \left(\frac{n+1}{n-1}\right)(a_{n-1} - ib_{n-1}) - \frac{2(n+1)}{nx}(n-iy)[a_n - ib_n], \qquad (41)$$

which leads to the final results

$$a_{n+1} = \left(\frac{n+1}{n-1}\right)a_{n-1} - \frac{2(n+1)}{nx}(na_n - yb_n), \qquad (42)$$

$$b_{n+1} = \left(\frac{n+1}{n-1}\right)b_{n-1} - \frac{2(n+1)}{nx}(nb_n + ya_n). \qquad (43)$$

Appendix V

Series Expansions

Eq. (10) may be written in the form

$$c_n = \frac{in\pi}{\sinh \pi y} H(x), \qquad (n > 0). \qquad (44)$$

The last result must be simplified. We have

$$\Gamma(n - iy + s + 1) = \prod_{r=1}^{n}(s + r - iy)\cdot\Gamma(s + 1 - iy), \qquad (46)$$

$$\Gamma(s+1-iy)\Gamma(s+1+iy) = \Gamma(iy)\Gamma(-iy)\prod_{r=0}^{s}[(r+iy)(r-iy)] = \frac{\pi}{y}\operatorname{csch}\pi y \prod_{r=0}^{s}(r^2+y^2). \qquad (47)$$

Thus,

$$c_n = iny\sum_{s=0}^{\infty} \frac{\left(\frac{x}{2}\right)^{n+2s}(n+2s)!\prod_{r=1}^{n}(s+r+iy)}{s!(n+s)!\prod_{r=0}^{s}(r^2+y^2)\cdot\prod_{r=1}^{n}[(s+r)^2+y^2]}. \qquad (48)$$

When c_n is set equal to $(a_n - ib_n)/2$, series for the real coefficients can be obtained. Because of the complex product in the numerator of (48), such separation becomes progressively more complicated as n increases. Hence, it is convenient that series results need only be obtained for $n=1$ and 2, with the recursion relations used for higher n. Separation yields the series

$$a_1 = -2y^2\sum_{s=0}^{\infty}\frac{(x/2)^{1+2s}(1+2s)!}{s!(1+s)![(1+s)^2+y^2]\prod_{r=0}^{s}(r^2+y^2)}, \qquad (49)$$

$$b_1 = -2y\sum_{s=0}^{\infty}\frac{(x/2)^{1+2s}(1+2s)!}{[s!]^2[(1+s)^2+y^2]\prod_{r=0}^{s}(r^2+y^2)}, \qquad (50)$$

$$a_2 = -4y^2\sum_{s=0}^{\infty}\frac{(x/2)^{2+2s}(2+2s)!(3+2s)}{s!(2+s)![(1+s)^2+y^2][(2+s)^2+y^2]\prod_{r=0}^{s}(r^2+y^2)}, \qquad (51)$$

$$b_2 = -4y\sum_{s=0}^{\infty}\frac{(x/2)^{2+2s}(2+2s)![(1+s)(2+s)-y^2]}{s!(2+s)![(1+s)^2+y^2][(2+s)^2+y^2]\prod_{r=0}^{s}(r^2+y^2)}. \qquad (52)$$

where

$$\begin{aligned} H(x) &\equiv I_{n-iy}(x)I_{iy}(x) \\ &= e^{-in\pi/2}J_{n-iy}(ix)J_{iy}(ix) \\ &= (i)^{-n}J_{n-iy}(ix)J_{iy}(ix) \\ &= \sum_{s=0}^{\infty}\frac{(x/2)^{n+2s}\Gamma(n+2s+1)}{s!\Gamma(n+s+1)\Gamma(n-iy+s+1)\Gamma(iy+s+1)}, \end{aligned} \qquad (45)$$

and the last result follows from a Bessel function expansion given by Watson.[16]

[16] *Ibid.*, p. 147.

These series are absolutely convergent for $0 \leq x \leq \infty$. It is worth pointing out that their initial terms $(s=0)$ agree with the results of Appendix I when $z = y^{-1}$ is used.

For numerical summation of the above series, it is worth pointing out that they all involve simple functions of s times the quantity

$$A_s \equiv \frac{(2s)!}{2^{2s}s!s!} = \frac{1}{2^{2s}}\binom{2s}{s}, \qquad (53)$$

and A_s satisfies the recursion relation $A_{s+1} = [(s+\frac{1}{2})/(s+1)]A_s (s>0)$ and $A_0 = 1$.

Appendix VI

Carbon Microphone

Consider a single-ended carbon microphone in series with a battery, V_0, and an inductance L, the primary of an input transformer. If a sinusoidal sound wave of frequency $(\omega/2\pi)$ impinges on the microphone, its resistance R will be given by $R=R_0(1+m \sin \omega t)$, where the modulation factor $m(0 \leq m \leq 1)$ depends on the amplitude of the incident wave. The pertinent differential equation for the current i is

$$\frac{di}{dt}+\left(\frac{R_0}{L}\right)(1+m \sin \omega t)=\frac{V_0}{L}. \tag{54}$$

A steady-state solution is of the form,

$$\left(\frac{i}{i_0}\right)=ye^{-y\phi+x\cos\phi}\int e^{y\phi-x\cos\phi}d\phi, \tag{55}$$

which should be compared with (4). Here $i_0 \equiv V_0/R_0$ as before, but $y=(\omega L/R_0)^{-1}$ since the time constant is now $\tau_0 \equiv L/R_0$ rather than RC_0 as in the capacitance problem.

Because of the similarity of (55) and (4), it is readily shown that the steady-state current is given by

$$\left(\frac{i}{i_0}\right)=\frac{ye^{x\cos\phi}}{(e^{2\pi y}-1)}\int_0^{2\pi} e^{y\mu-x\cos(\phi+\mu)}d\mu, \tag{56}$$

and

$$c_n=\frac{\pi y}{\sinh \pi y}I_{iy}(x)\cdot I_{n-iy}(x), \qquad (n \geq 0). \tag{57}$$

a result only slightly different from (10). Note particularly, however, that c_0 is no longer zero, and there is thus a zero-frequency component in the current. In particular, we have (using an R superscript for the present case)

$$c_0{}^R=\frac{2y}{\sinh \pi y}\int_0^{\pi/2}\cosh(2\theta y)I_0(2x\cos\theta)d\theta, \tag{58}$$

$$=y^2\sum_{s=0}^{\infty}\binom{2s}{s}\frac{(x/2)^{2s}}{\prod_{r=0}^{s}(r^2+y^2)}. \tag{59}$$

Note that $c_0{}^R \rightarrow 1$ as $m \rightarrow 0$ and also as $\omega \rightarrow \infty$. As ω approaches zero, however, the sum of the series for $c_0{}^R$ approaches a limit greater than unity for $m>0$. In this case the series may be summed and yields $c_0{}^R=(1-m^2)^{-1/2}$. The excess over unity arises from rectification of some of the incident energy by time-varying resistance of the microphone. Clearly, infinite incident energy is necessary to cause $m=1$ in the limit of zero frequency. For $n>0$ we have

$$c_n{}^R=(y/in)c_n{}^C, \qquad (n>0). \tag{60}$$

where the R and C superscripts denote the carbon microphone and capacitance values, respectively, of the complex Fourier coefficients. It immediately follows that

$$\alpha_n{}^R=(y/n)\alpha_n{}^C \tag{61}$$

$$\chi_n{}^R=\chi_n{}^C-90°. \tag{62}$$

A push-pull, or double-button, carbon microphone can be handled in the same general way as the push-pull capacitance microphone.

Appendix VII

Power Relations

Let $\xi \equiv (i/i_0)$. It is readily shown that

$$\left.\begin{aligned} q &= CV_0(1-\xi) \\ &= C_0V_0(1-\xi)/M \equiv q_0(1-\xi)/M, \\ V_c &= V_0(1-\xi), \end{aligned}\right\} \tag{63}$$

where V_c is the instantaneous voltage across the time-varying capacitance C. Denote the stored energy in the capacitance by E and the work done in moving its plate by W. The instantaneous power dissipated in the capacitor will be $P_c=dE/dt$, and the instantaneous input power will be $P_{in}=dW/dt$. The force between the plates, F_c, may be written in the form $F_c=C_0V_0{}^2(q/q_0)^2/2d_0$. Finally, define P_{out} as the power dissipated in the output resistance, and P_0 as $\equiv V_0{}^2/R$.

We may immediately write

$$\begin{aligned} E &= \frac{1}{2}CV_c{}^2=\frac{1}{2}C_0V_0{}^2(1-\xi)^2/M \\ &\equiv E_0(1-\xi)^2/M, \end{aligned} \tag{64}$$

$$\begin{aligned} P_c &= \frac{1}{2}V_c{}^2\frac{dC}{dt}+CV_c\frac{dV_c}{dt} \\ &= iV_c-\frac{1}{2}V_c{}^2\frac{dC}{dt} \\ &= P_0\left\{(1-\xi)\left[\xi-\frac{1}{2}(1-\xi)R\frac{dC}{dt}\right]\right\} \\ &= P_0\left\{(1-\xi)\left[\xi+\frac{mz\cos\phi}{2M^2}(1-\xi)\right]\right\}, \end{aligned} \tag{65}$$

where

$$i=V_c\frac{dC}{dt}+C\frac{dV_c}{dt}$$

has been used. Since

$$dW=F_c dd=F_c d_0\, m\omega\cos\phi\, dt,$$

P_{in} becomes

$$\begin{aligned} P_{in} &= F_c d_0 m\omega\cos\phi \\ &= P_0[m\omega RC_0\cos\phi(q/q_0)^2/2], \end{aligned} \tag{66}$$

and P_{out} is

$$P_{out}=P_0\xi^2. \tag{67}$$

Now, using $\xi=1-yMF_1$, where F_1 is the integral defined in Appendix II, the desired quantities may be written as

$$\left.\begin{aligned} q/q_0 &= yF_1, \\ V_c/V_0 &= MyF_1, \\ E/E_0 &= M(yF_1)^2, \\ P_c/P_0 &= (MyF_1) - (MyF_1)^2 + \frac{mz\cos\phi}{2}(yF_1)^2, \\ F_c/F_0 &= (yF_1)^2, \\ P_{in}/P_0 &= \frac{mz\cos\phi}{2}(yF_1)^2, \\ P_{out}/P_0 &= 1 - 2(MyF_1) + (MyF_1)^2, \end{aligned}\right\} \quad (68)$$

where $F_0 \equiv C_0V_0^2/2d_0$. Note that the expression for F_c shows that this force will not become infinite even when $m=1$. The above equations lead to the general relationship,

$$(P_{in}/P_0) + (i/i_0) = (P_c/P_0) + (P_{out}/P_0). \qquad (69)$$

Next, it is desirable to obtain the average values of the above quantities in the steady state. For this purpose, the quantity F_1 must be replaced by F_3. It has already been shown in Appendix III that for the steady state $c_0=0$; thus $\langle\xi\rangle=0$ and $\langle MyF_3\rangle=1$, where the pointed brackets denote time averages. This absence of a dc current means that the battery cannot supply power on the average to the load resistance. Some of the above averages are difficult to carry out because of the presence of the integral F_3 in them, but the next two results avoid such difficulties. We have,

$$\begin{aligned} \langle P_c/P_0\rangle &= \frac{1}{2\pi}\int_0^{2\pi}\frac{dE}{dt}\,d\phi \\ &= \frac{\omega}{2\pi}\left[E(2\pi) - E(0)\right] = 0, \end{aligned} \qquad (70)$$

where the last equation follows from the stationary, periodic character of the stored capacitor energy. Then, Parseval's theorem may be applied to yield the following expression for $\langle P_{out}/P_0\rangle$, valid for $0 \le x < \infty$,

$$\begin{aligned} \langle P_{out}/P_0\rangle &= \langle\xi^2\rangle = -1 + \langle(MyF_3)^2\rangle \\ &= 2\sum_{n=1}^{\infty}|c_n|^2, \end{aligned} \qquad (71)$$

noting that $c_0=0$. We may now write,

$$\left.\begin{aligned} \langle q/q_0\rangle &= \langle yF_3\rangle, \\ \langle V_c/V_0\rangle &= \langle MyF_3\rangle = 1, \\ \langle E/E_0\rangle &= \langle M(yF_3)^2\rangle, \\ \langle P_c/P_0\rangle &= 1 - \langle(MyF_3)^2\rangle + \frac{mz}{2}\langle(yF_3)^2\cos\phi\rangle \\ &= 0, \\ \langle F_c/F_0\rangle &= \langle(yF_3)^2\rangle, \\ \langle P_{in}/P_0\rangle &= \frac{mz}{2}\langle(yF_3)^2\cos\phi\rangle. \end{aligned}\right\} \quad (72)$$

Note that (69) leads to $\langle P_{in}\rangle = \langle P_{out}\rangle$, a necessary condition since the battery can supply no average power.

Although it has not been found practical to evaluate the remaining averages in (72) in closed form, it is possible to use (14) and expand F_3 in a series useful for $x \ll 1$ and $z \gg 1$. Using the resulting series, the necessary averaging may be carried out, and one finds for high relative frequencies the following terms to second order in x,

$$\left.\begin{aligned} \langle q/q_0\rangle &= 1 + (x^2/2) + \cdots, \\ \langle E/E_0\rangle &= 1 + (x^2/2) + \cdots, \\ \langle F_c/F_0\rangle &= 1 + (3x^2/2) + \cdots, \\ \langle P_{in}/P_0\rangle &= \langle P_{out}/P_0\rangle \\ &= (m^2/2) - (1 - m^2)(x^2/2) + \cdots. \end{aligned}\right\} \quad (73)$$

Acknowledgment

The authors are grateful to P. Carnahan and G. Milner for their work in programming the present problem for the IBM computer. Helpful comments from G. K. Walters and T. L. Estle are also appreciated.

Supplement to IRE Standards on Graphical Symbols for Electrical Diagrams, 1954*

54 IRE 21. S1

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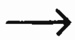
1. CONNECTORS THAT ARE INTEGRAL TO WAVEGUIDE SECTIONS AND ELEMENTS

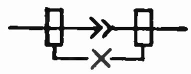
1.1 Flange, Plain.



1.2 Flange, Choke.

1.3 Intentional Direct-Current-Isolation of Waveguide Path.

1.4 Mated Plain and Choke Flanges for Rectangular Waveguide with Direct-Current Isolation (Insulation) Between Sections of Waveguide.



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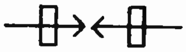
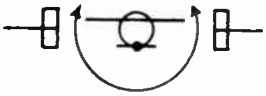
1.5 Mated Plain Flanges for Rectangular Waveguides.



2. ROTARY JOINT

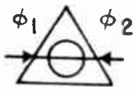
2.1 Coaxial Type in Rectangular Waveguide System.



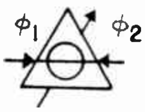
2.2 Circular-Waveguide Type in Rectangular Waveguide System.

3. DISCONTINUITY

3.1 Differential Phase Shifter. Phase shift ϕ in direction of arrowhead and its limits may be indicated.

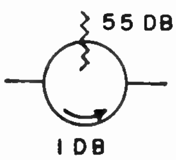


3.2 Adjustable Differential Phase Shifter.



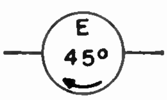
4. UNIDIRECTIONAL PAD (ISOLATOR)

Power flowing in direction of arrow is not intentionally attenuated but power flowing in opposite direction is attenuated. If required, the values of attenuation shall be shown in the locations indicated.

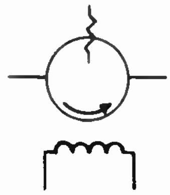


5. FIELD POLARIZATION ROTATOR

Arrow indicates direction of rotation of electric field when viewed in direction of signal flow.

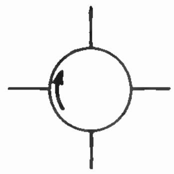


6. FIELD-POLARIZATION AMPLITUDE MODULATOR



7. CIRCULATOR

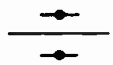
Arrowhead indicates direction of power flow from any input to next adjacent arm but not to any other arm. Circulator may have three or more ports.



8. TRANSMISSION PATH

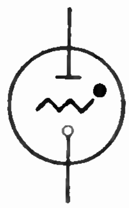
8.1 Strip Line.

8.2 Sandwich Line.



9. TUBE, ELECTRON

9.1 Radiation Counter, Gas Filled



Deletion from 54 IRE 21. S1

43.11.3 Ridged (waveguide). As this is only one of numerous special structures, it should be deleted so that only the two fundamental forms, circular and rectangular, are given in the Standard. Otherwise, all known special structures should be given distinctive graphical symbols.

Pseudo-Rectification and Detection by Simple Bilateral Nonlinear Resistors*

J. E. BRIDGES†, SENIOR MEMBER, IRE

Summary—The diode, or other devices containing a unilateral element, is generally employed for rectification or detection. Ordinary carbon resistors, however, and to a more marked degree, certain symmetrical or bilateral nonlinear resistors containing no unilateral elements, can, when a pulse-type ac waveform is impressed upon them, perform a rectification process; this process may be considered solely as a circuit effect. Four basic ac-to-dc converting circuits, some of which are largely independent of temperature, and which employ symmetrical nonlinear resistors, are analyzed.

If an asymmetrical waveform is available as a carrier for simple types of modulation, the symmetrical nonlinear resistor may be used as a detector with several signal-enhancing properties not usually present in the ordinary diode or square-law detection process. The nonlinear resistor detection circuit is capable of radically suppressing symmetrical interfering signals, such as Gaussian noise at one point in the signal spectrum following the detector, in this case, at zero frequency. For certain types of signal systems, this detection process can maintain for very weak signals a linear relationship between the impressed input and resulting output signal-to-noise ratios associated with the detector circuit.

Two illustrative examples are given: a pseudo-rectification circuit supplying 3000-volt dc potential in a television receiver, and a circuit capable of enhancing the detection of very minute amounts of radiation impressed upon a photomultiplier.

Introduction

The diode, or other devices containing a unilateral element, is widely used for rectification or detection. The property of ac-to-dc conversion is not unique to the diode-type devices but is present in a small degree in many resistors used in the electronics industry today. This ac-to-dc conversion property is quite marked in certain symmetrical nonlinear resistors.

To illustrate the basic concept, Fig. 1 shows a volt-ampere curve of a typical semiconductor diode rectifier which includes a p-n junction; and, due to its asymmetric characteristics, any signal of reasonable amplitude applied to it will undergo an ac-to-dc conversion. Fig. 2 shows a typical volt-ampere characteristic for a linear resistor and for a typical symmetrical nonlinear resistor, neither of which include a p-n junction or other unilateral elements. This nonlinearity is present in most carbon composition resistors available on the market today. Resistors which exhibit this nonlinearity to a very marked degree are sold under the names of "Thyrite" and "Varistor." These effects are further described by Schwertz and Mazenko,[1] Williams and Thomas,[2] and Ashworth, Weedham, and Sillars.[3]

Within certain amplitude limits, almost any type of signal applied to the diode which has the characteristics shown in Fig. 1 will result in an ac-to-dc conversion. If an *symmetrical* signal is applied to the resistor with the nonlinear volt-ampere characteristics of that shown in Fig. 2, no ac-to-dc conversion would take place; how-

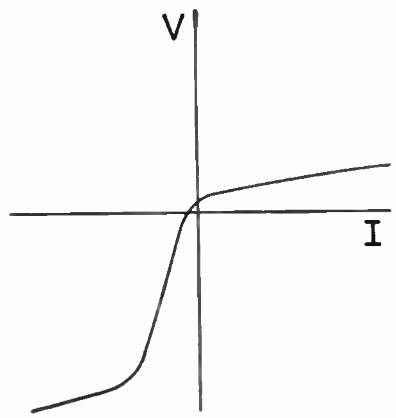


Fig. 1—Voltage-current characteristic of a typical semiconductor diode.

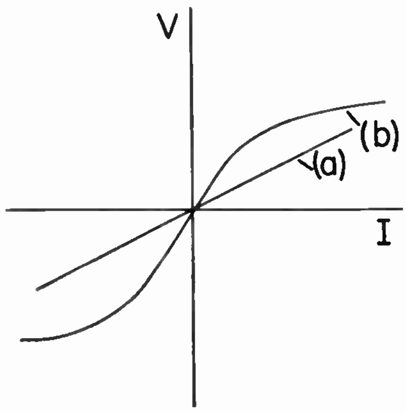


Fig. 2—Voltage-current characteristic of (a) a linear resistor and (b) a nonlinear resistor.

* Received by the IRE, July 5, 1960; revised manuscript received, November 17, 1960.

† Warwick Mfg. Corp., Chicago, Ill.

[1] F. A. Schwertz and J. J. Mazenko, "Nonlinear semiconductor resistors," *J. Appl. Phys.*, vol. 24, pp. 1015–1024; August, 1953.

[2] T. R. Williams and J. B. Thomas, "Current noise and nonlinearity in carbon resistors," IRE Trans. on Component Parts, vol. CP-5, pp. 151–153; December, 1958.

[3] F. Ashworth, W. Needham, and R. W. Sillars," "Silicon carbide non-ohmic resistors," *J. IEE*, vol. 93, pp. 385–405; September, 1946.

ever, if an *asymmetrical* signal such as a pulse with no dc component is applied, an ac-to-dc conversion will take place.

This may be better understood by examining Fig. 3. If a pulse-type repetitive ac waveform $v(t)$ having no dc component, so that

$$\bar{v}(t) = \frac{1}{T}\int_0^T v(t)dt = 0 \tag{1}$$

is applied to the linear resistor curve (a) indicated in Fig. 3(b), then a current I_1 will flow during the time of the pulse and I_2 during the absence of the pulse. In the case of the nonlinear resistor, curve (b), a much greater current flows during the pulse (t_0—t_1) than during the absence of the pulse (t_1—t_2). Assuming that the currents in both the linear and nonlinear resistor are the same during the pulse, then it is easily seen that a much smaller current I_3 flows in a nonlinear resistor during the absence of the pulse than does in the linear resistor. Thus the average value of this smaller current I_3 will not equal the average value of I_1. As a consequence, a net dc current is generated.

This property is useful in a number of ways, such as for developing a dc bias from an ac pulse-type source. In addition, of even greater interest is the use of this in noise analysis and as a signal detector. Subsequent sections of this paper will first examine in detail the ac-to-dc conversion properties of four basic circuit configurations. This will be followed by an analysis of the role of a symmetrical nonlinear resistor in the role of a detector. These analyses will then be followed by a discussion of several practical applications of this effect. The mathematical details of the analysis will be presented in the Appendix.

This property of ac-to-dc conversion by simple symmetrical nonlinear resistors has not been analyzed in detail in the past, as far as can be ascertained. The general idea of using nonlinear elements to effect signal enhancement of some form of detection is well known. Most of these approaches, which will be cited later, are, for the most part, generalized approaches which determine the optimum nonlinear transfer function requirements and other signal-processing specifications. Here the objective is somewhat different and is aimed toward making a specific analysis of the signal-enhancing properties of a simple, easily obtained component. This type of analysis does not necessarily result in optimum methods for signal enhancement, but it does provide an illustration of a method of signal enhancement which may not greatly increase the complexity of a communication or instrumentation system.

AC-to-DC Conversion

The current and voltage relationships of a nonlinear resistor can be given in many ways. The most common is defined as follows,

$$I = |KV^n|, \tag{2}$$

where I is resistor current of the same sign as the voltage, K is a constant, V is applied potential, and n is an exponent.

Now, assume that there is available a rectangular current or voltage waveform, as indicated in Fig. 3(a), which has no dc component, where r is defined as the ratio of the duration of the pulse, t_1—t_0, to the remainder of the period, t_2—t_1, when the pulse is absent. This ratio r of ON to OFF time and the exponent n of the nonlinear resistors are the important parameters of the following analysis, which is summarized in Fig. 4.

If a nonlinear resistor is placed in series with a zero-impedance voltage source [Fig. 4(a)], a dc current will result. As developed in the Appendix, the value of the dc current flowing will be as follows:

$$I_{dc} = K(V_{o-p})^n\left(\frac{r - r^n}{1 + r}\right). \tag{3}$$

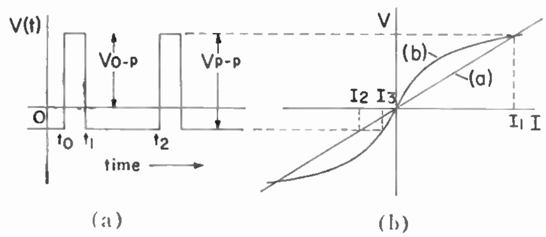


Fig. 3—A time-amplitude function of an asymmetrical waveform having no dc component applied to a linear resistor [curve (a)] and a nonlinear resistor [curve (b)].

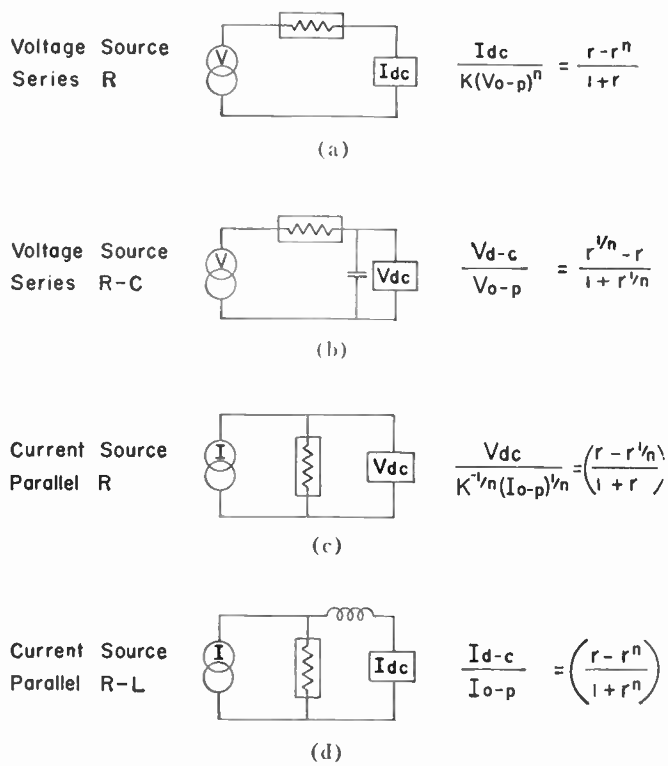


Fig. 4—Summary of four basic circuit configurations for pseudo-rectification and the resulting relationships from an idealized rectangular pulse source.

Fig. 5(a) shows the normalized (K and V_{o-p} equated to 1) dc output, as a function of the ON and OFF time of the pulse with the exponent as the additional parameter. Note that for a given exponent, a maximum of the ac-to-dc conversion occurs. Obviously when the pulse is extremely small, that is, as r approaches zero, very little dc current will flow. As r is increased and approaches the value of 1, a maximum of dc conversion appears first, and then the amount of dc converted drops off again to zero, where the waveform applied to the nonlinear resistor is perfectly symmetric.

If a nonlinear resistor is placed in series with a condenser and a zero-impedance voltage source of pulses, an ac-to-dc conversion again takes place. In this case, assuming no load is placed across the condenser, a dc voltage is developed equal to

$$V_{dc} = V_{o-p}\left(\frac{r^{1/n} - r}{1 + r^{1/n}}\right). \tag{4}$$

Here again, at very narrow pulse widths, hardly any dc voltage is developed. As the ON time of the pulse is increased, the voltage maximum is quickly reached and generally falls off linearly as r is increased to 1. An important advantage of the RC combination of Fig. 4(b) is that the output voltage is independent of the specific conductivity K of the nonlinear resistor. Therefore, any variation in output due to changes in conductivity with temperature is minimized. The maximum theoretical dc voltage is limited to 50 per cent of the pulse height. Fig. 6 shows the normalized, ($V_{o-p}=1$), dc-voltage output as a function of r and n.

If a nonlinear resistance is placed across an infinite-impedance current source of pulses which has no dc component [Fig. 4(c)], a dc voltage will result. Fig. 4(c) and normalized data given in Fig. 7 summarize the behavior of this circuit configuration.

If a nonlinear resistor is placed in parallel with a very large inductor and an infinite-impedance current source of pulses [Fig. 4(d)] which has no dc component, again a dc component will be developed. As in the

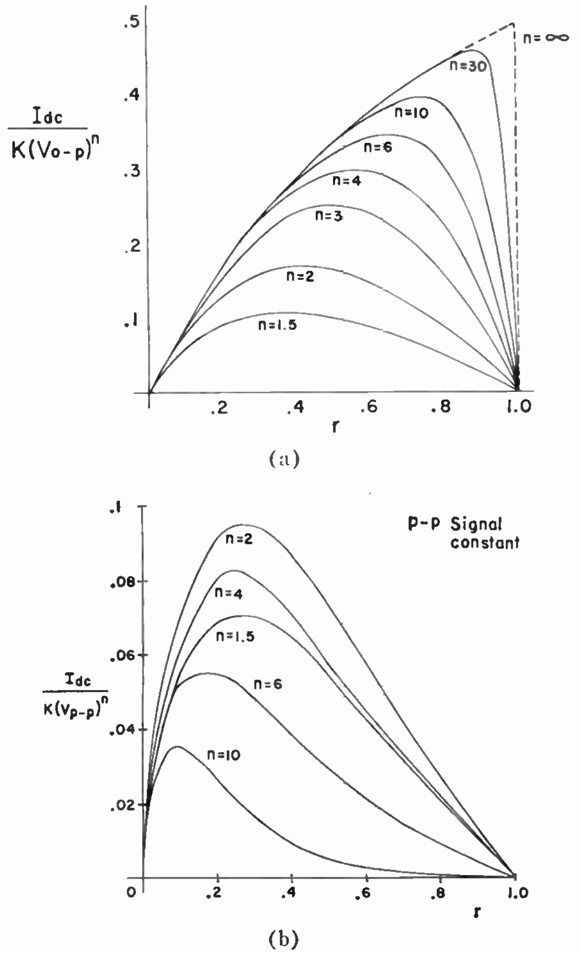


Fig. 5—(a) Normalized dc current output from an idealized pulse voltage source and series nonlinear resistor, based on V_{o-p} held constant. (b) Normalized dc current output from an idealized pulse voltage source and series nonlinear resistor, based on V_{p-p} held constant.

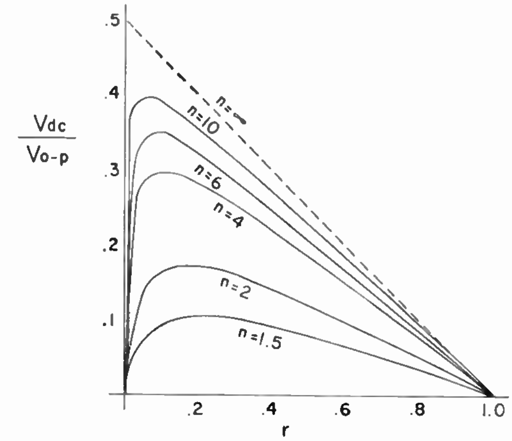


Fig. 6—Normalized dc voltage output from an idealized pulse voltage source in series with a nonlinear resistor and a condenser of large value, based on V_{o-p} held constant.

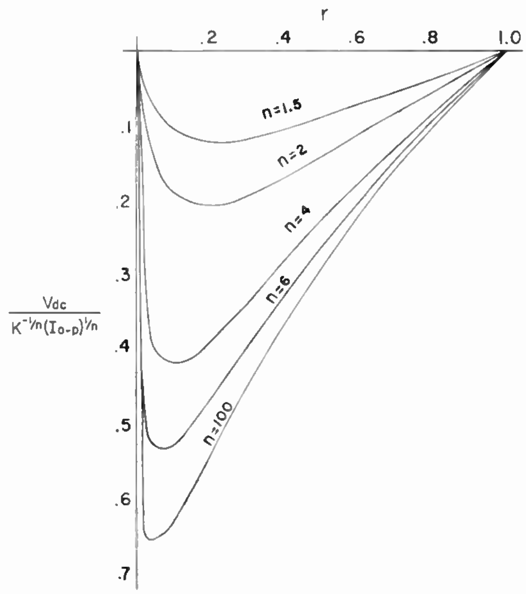


Fig. 7—Normalized dc voltage output from an idealized pulse current source in parallel with a nonlinear resistor, based on I_{o-p} held constant.

case of the voltage source and nonlinear resistor in series with a capacitor, the dc current developed is independent of the conductivity K of the nonlinear resistor. Fig. 8 presents the time and exponent relationships.

For certain types of efficient pulse generators such as transistorized blocking oscillators, and for certain types of automatic-gain controlled pulse amplifiers, the zero-to-peak voltage is held constant. The curves presented in Figs. 5(a) and 6–8, therefore, are based on fixed zero-to-peak voltages or currents. If peak-to-peak signals are held constant instead of the zero-to-peak signals (which is the case for some commercial laboratory pulse generators), the current maximums are generally shifted toward a smaller value of r, as illustrated in Fig. 5(b). Other basic circuit configurations are also possible, such as voltage source with a nonlinear resistor in series with an inductor; however, a little study will show that some of these other configurations are not quite as meaningful as the ones which are shown. It is also worthwhile to note that the circuit of Fig. 4(a) is not the dual of circuit Fig. 4(c).

Detection

Detection may be thought of as the rectification of modulated RF energy with the recovery of the modulating signal. Within this definition, symmetrical nonlinear resistors may be regarded as detectors, provided the "carrier" has a nonsymmetrical waveform.

A symmetrical nonlinear resistor, as a detector, has one unique property which makes it interesting from a noise and interference point of view. It has the ability to suppress symmetrical signals. This, of course, both enhances and limits its usefulness. In addition, as the subsequent analysis will show, this type of detector has other advantages as well.

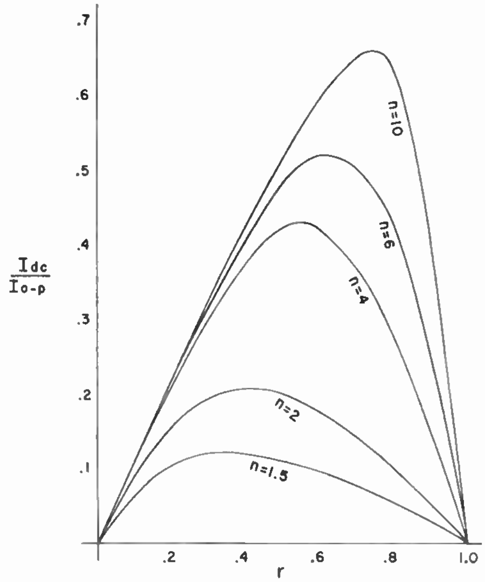


Fig. 8—Normalized dc current output from an idealized current source in parallel with a nonlinear resistor and a very large inductor, based on I_{o-p} held constant.

The over-all effects of interference and noise on the performance of a symmetrical nonlinear detector may be better understood by assuming the circuit of Fig. 4(a) to be modified to include a signal $s(t)$, plus a noise or interference signal $n(t)$, and a suitable filter circuit as indicated in Fig. 9. The current output of the detector is then passed through a low-pass filter into either a dc current indicator or an rms current indicator.

Some choice is permitted in the selection of the frequency response prior to the detector and is exemplified by response $Y_0(\omega)$, a broad-band video-type response, or $Y_1(\omega)$, a narrow-band comb-filter response. The narrow-band filter responses are centered about the harmonics of the pulse-type carrier. This pulse-type carrier is assumed to be turned off-and-on at an equivalent angular modulation rate of no greater than Δ radians per second. This limits the minimum bandwidth of each filter response to approximately 2Δ. The low-pass filter following the detector would have a minimum bandwidth of Δ. These idealized filter responses are shown in Fig. 10.

First, consider the effects of a sinusoidal interfering signal,

$$n(t) = b \cos yt, \tag{5}$$

with the desired signal carrier $s(t)$ consisting of the first two harmonics of a pulse,

$$s(t) = a_1 \cos \omega_c t + a_2 \cos 2\omega_c t. \tag{6}$$

Referring to Fig. 10, assume a predetection response of $Y_1(\omega)$, and assume that the interfering signal can appear anywhere within the two band-pass responses. The resultant current through R_1 of Fig. 9, will be, assuming n is 3:

$$i(t) = K[s(t) + n(t)]^n \tag{7}$$

$$= s^3(t) + 3[s^2(t)n(t)] + 3[s(t)n^2(t)] + n^3(t). \tag{8}$$

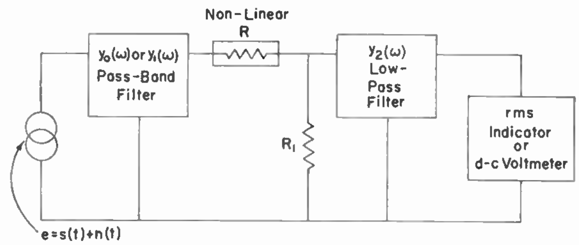


Fig. 9—Nonlinear resistor pseudo-detector circuit.

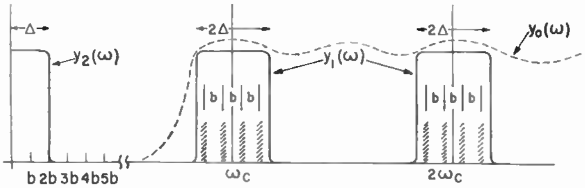


Fig. 10—Filter responses prior to detector and subsequent to detector.

Substituting the sinusoidal functions for $n(t)$ and $s(t)$, and carrying out the indicated operations and keeping only the low-frequency terms, $i_{l-f}(t)$, the results are as follows:

for $\omega_c-\Delta/2<y<\omega_c+\Delta/2$,

$$i_{l-f}(t) = \frac{3a_1^2a_2}{4} + \frac{3a_1a_2}{2} b \cos(\omega_c - y)t + \frac{3}{4} a_2b^2 \cos 2(\omega_c - y)t; \tag{9}$$

for $\omega_c-\Delta<y<\omega_c-\Delta/2$ and $\omega_c+\Delta/2<y<\omega_c+\Delta$,

$$i_{l-f}(t) = \frac{3a_1^2a_2}{4} + \frac{3a_1a_2}{2} b \cos(\omega_c - y)t; \tag{10}$$

for $2\omega_c-\Delta<y<2\omega_c+\Delta$,

$$i_{l-f}(t) = \frac{3a_1^2a_2}{4} + \frac{3}{4} a_1^2b \cos(2\omega_c - y)t. \tag{11}$$

This shows that if no signal is present ($a_1=a_2=0$), the effect of the interfering sinusoidal signal is completely rejected. In other words, no low-frequency terms are present if the signal is absent.

Only if the interfering sinusoidal signal is within Δ radians per second of a harmonic of the pulse, does an interfering signal appear at the output of the filter.

If an rms indication, rather than a dc indication, is used following the detector, the presence of the interfering sinusoid can only enhance the output indication. In other words, the output contains a signal term and an interference term but no interference terms by themselves. If a simple binary output indication is required, the rms value of both the signal and signal-interference term aggregate could be interpreted as the total signal. Under these conditions, the signal-to-interference ratio would be very large.

If the video-type predectection filter response $Y_0(\omega)$ is assumed, (9), (10) and (11) still hold. If the interfering signal is not within Δ radians per second of any of the pulse frequencies, only the desired dc output, the first term of either (9), (10) or (11), is present, due to the filtering action of the $Y_2(\omega)$ response.

This situation does not prevail as the number of interfering signals are increased, because the interfering signals are so numerous that they eventually approximate harmonic relationships between themselves. It is possible, therefore, for a group of interfering signals to be related so that a pulse train is momentarily duplicated, which results in a detected output independent of the signal. If the interfering signals were selected at random, the sense of the pulses and the resulting sign of a detected output would obviously vary from plus to minus, and the detected output would not be expected to maintain a net average dc value. Such would be the case for thermal noise.

A meaningful approximate approach to analysis of the effects of thermal noise would be to assume that the first two harmonics of the pulse to be centered in a rectangular pass band 2Δ radians wide, and the detector low-pass filter response to be Δ radians wide as in Fig. 10. A number of mathematical approaches could be employed to analyze the case for thermal noise; however, the method of dividing the signal response prior to the detector into discrete channels,[4] each containing a sinusoidal noise signal phased at random with any set of points in time, will be used because the spectrum characteristics are easily preserved and because this method is more easily grasped from an intuitive point of view.

The signal remains the same as in the previous analysis, but instead of one sinusoidal interference signal, a large number of noise sinusoids are assumed. The sum of the signal and ensemble of noise sinusoids are substituted in (2) and the resultant low-frequency terms ascertained.

As might be expected, the low-frequency terms of the aforementioned expansion are similar to (8). As developed in the Appendix, the rms value of these low-frequency terms may be expressed in terms of the input rms signal-to-noise ratio σ, the amplitude of the fundamental of the carrier c, as follows:

$$i_{l-f}(t)_{\text{rms}} = \left\{(c^3f_1)^2 + \left(\frac{c^3f_2}{\sigma}\right)^2 + \left(\frac{c^3f_3}{\sigma^2}\right)^2 + \left(\frac{c^3f_4}{\sigma^3}\right)^2\right\}^{1/2} \tag{12}$$

for $n=3$, where f_1, f_2, f_3, and f_4 are constants and functions, respectively, of $s^3(t)$, $s(t)n^2(t)$, $s(t)n^2(t)$. and $n^3(t)$. The signal input $s(t)$ and the noise input $n(t)$ to the detector are each expressed as an ensemble of sinusoidally varying signals as follows:

$$s(t) = \sum_k c_k \cos[k\omega_c t + \theta_k]$$

$$n(t) = \sum_n a_n \cos(\omega_n t + \phi_n).$$

Specific terms of (12) are c^3f_1, the desired detected signal; c^3f_2/σ and c^3f_3/σ^2, low-frequency signals arising from cross-products between signal and noise terms; and c^3f_4/σ^3, low-frequency terms due to noise terms alone. From these terms the detected output signal-to-noise ratio can be defined.

As developed in the Appendix, the pass band prior to the detector $Y_1(\omega)$ is divided into equally-spaced noise generator channels, and a computation is carried out to develop the approximate value of the terms of (12). The ratio of the second harmonic to the fundamental of the pulse is assumed to be 0.7.

[4] J. E. Bridges, Detection of television signals in thermal noise," Proc. IRE, vol. 42, pp. 1396–1405; September, 1954.

The results of this computation mentioned in the Appendix is summarized in Fig. 11 for $N=3$. Curve A shows the ratio of the dc output c^3f_1 to the remaining terms. For large value of σ, a linear relationship between input and output is maintained. As σ approaches unity, a threshold is reached, and for very small values of σ, the output signal-to-noise ratio is deteriorated at three times the rate that the input signal-to-noise ratio is deteriorated. It should be pointed out that the rate of deterioration, after the threshold, is dependent on the value of the exponent n, and obviously, increasing or decreasing the value of n would affect the rate of deterioration subsequent to detection.

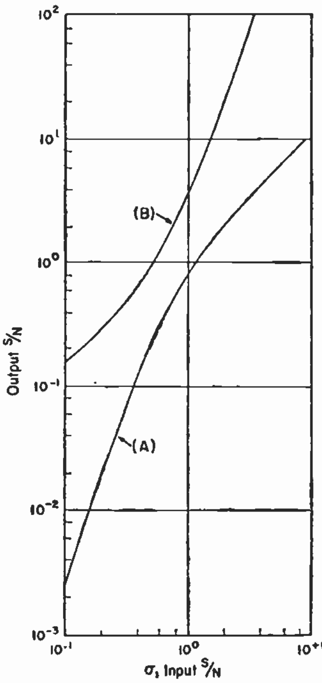


Fig. 11—Approximate input and output signal-to-noise amplitude ratios for detector of Fig. 10. Curve A DC output-to-noise ratio. Curve B DC output plus signal-and-noise aggregate to noise alone ratio.

Although noise alone does develop low-frequency ac signals, Gaussian noise and other types of symmetrical noise are still incapable of developing an average dc component. Where simple presence or absence of signal information is sought, time integration by readily constructed RC filter networks could enhance the output signal-to-noise ratio proportional to the square root of the ratio of the respective integration times. This is an important property, since no other simple passive detector, either diode or square-law, is capable of developing a dc output independent of other noise signals.

A second important property is that the bulk of the resultant output items of (12) contain a signal component of either signal alone or signal-and-noise cross products. The detection process is such that if the signal is absent, these signal and signal-and-noise products are absent also. Thus the signal-and-noise cross-product terms, c^3f_2/σ and c^3f_3/σ^2, as well as the signal term c^3f_1, indicate the presence of the signal prior to the detector. Since these signal-and-noise terms have no dc component, an rms indication or related display must be substituted for the dc indicator to utilize the signal-and-noise terms. If this can be done, then the aggregate of all terms containing a signal function could be considered as the detected output signal. Under these conditions the output signal-to-noise ratio would be improved over the dc indication case. Whether this is meaningful or not would depend on the type of signal-processing and use following the detector.

The results of this approach, the rms ratio of signal, and signal-and-noise terms to noise terms alone, are shown in Fig. 11, curve B. For large input signal-to-noise ratio σ, the aggregate output signal is enhanced and improves three times faster than the input signal-to-noise ratio is improved. As σ is decreased and approaches 1, this enhancement is lost, and as the input signal-to-noise ratio σ is reduced below unity, a linear relationship is maintained between the input and output signal-to-noise ratio.

Singleton[5] has demonstrated the benefits of the use of nonlinear filters with transfer characteristics tailored to the amplitude and phase statistics of the signal and of the noise. In some cases, Singleton's method results in a filter which requires constant amplitude inputs for optimum operation. In practical circuits this may be difficult to achieve, whereas the output signal-to-noise ratios of the nonlinear resistor detector circuit is less dependent on the combined signal and noise amplitudes applied to it. Generalized discussions on optimum approaches to nonlinear filters are given by Zadeh[6,7] and Balakrishnan and Drenick[8], and specific treatments concerning other types of nonlinear elements are given by Slattery,[9] White,[10] and Shooley and George.[11]

Discussion

The application of the foregoing can be divided roughly into two areas—circuit effects and signal properties. So far as circuit effects are concerned, this property has been utilized to develop high-voltage fo-

[5] H. E. Singleton, "Optimum nonlinear filters," Res. Lab. of Electronics, Mass. Inst. Tech., Cambridge, Rept. no. 160; August, 1950.

[6] L. A. Zadeh, "Optimum nonlinear filters," *J. Appl. Phys.*, vol. 24, pp. 396–1405; May, 1953.

[7] L. A. Zadeh, "A contribution to the theory of nonlinear systems," *J. Franklin Inst.*, vol. 255, pp. 387–408; May, 1953.

[8] A. V. Balakrishnan and R. Drenick, "On optimum nonlinear extraction and coding filters," IRE Trans. on Information Theory, vol. IT-2, pp. 166–172; September, 1956.

[9] T. G. Slattery, "The detection of a sine-wave in the presence of noise by a nonlinear filter," Proc. IRE, vol. 40, pp. 1232–1236; October, 1952.

[10] W. D. White, "Role of nonlinear filters in electronic systems," *Proc. NEC*, vol. 9, pp. 505–512; September, 1953.

[11] A. H. Shooley and S. F. George, "Input vs output signal-to-noise ratio characteristics of linear, parabolic, and semicubical detectors," *Proc. NEC Chicago*, vol. 17, pp. 151–161; October, 1951.

cusing potential bias for cathode-ray picture tubes,[12] such as the 17FP4 and 17GP4. The focusing potential required for this tube is approximately 20 per cent of the anode potential at a very low load current. A pulse voltage similar to those indicated in Fig. 3 is normally available from the flyback transformer in a commercial television receiver. Circuit developed is shown in Fig. 12

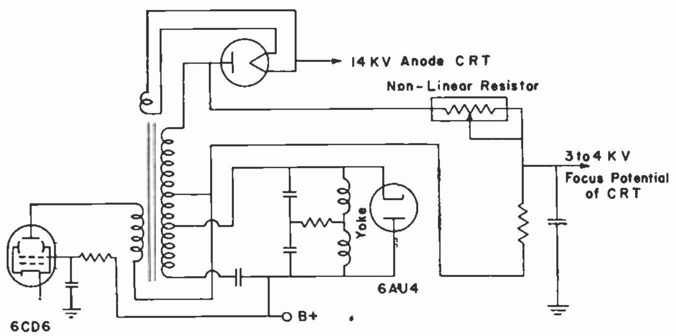


Fig. 12—Focus voltage supply using nonlinear resistor pseudo-rectifier for a cathode-ray picture tube.

with the indicated waveforms and a sketch of the nonlinear resistor pseudo-rectifier. Referring to Fig. 6, it is seen that an output voltage of approximately 25 per cent of the peak is readily obtained from nonlinear resistors having an exponent varying between 3 and 6, and pulses with r in the order of 0.2. In addition, it should be noted that this particular arrangement makes the voltage output relatively independent of the specific conductance K. Consequently, the variation in output voltage due to changes of K with temperature is minimized if the load is small.

Some variation in the focusing potential is required, and to accomplish this variation, the length of a long, cylindrical, silicon-carbide nonlinear resistor was varied. This was accomplished by substituting a sliding conductive rubber sleeve for the usual sprayed metal end connection. This changes not only K, but n as well, since n is partially dependent on the applied voltage-per-unit length of the rod-type resistor. The resistance loading provided by the load resistor R_L should be substantially greater than the load imposed by the focusing anode of the picture tube.

The efficiency of the nonlinear converter was such that the total power drain on the high-voltage system was the same when the nonlinear resistor focusing potential circuit was substituted for the 1X2 high voltage rectifier tube and potentiometer circuit that was normally used. This circuit has the advantage of long life, since it contains no filament. Slight variations in output were observed over approximately a preliminary 300-hour life test but were within the tolerance required for the picture tube. In terms of cost, the circuit appeared to be comparable to the tube circuit and less expensive than other solid-state rectifiers, such as selenium stacks and silicon stacks.

The efficiency of the ac and dc conversion circuits of this type tends to be low and depends on the pulse circuit type, pulse duration and amplitude, and impedance of the load. As a consequence, these ac-to-dc conversions would generally be limited to circuits requiring very little power. Also, as indicated in Fig. 6, the output voltage is only a small fraction of the applied pulse.

At times, the small nonlinearities found in the ordinary $\frac{1}{2}$-watt carbon resistors widely used in the electronics industry can be of consequence, particularly in the design of very stable pulse oscillators and pulse-utilizing phase detectors. The ac-to-dc conversion properties of ordinary carbon resistors may partially explain why resistors of different manufacturers are not always interchangeable in these critical pulse circuits.

So far as the signal properties of this effect are concerned, a detailed and quantitative comparison between the symmetrical nonlinear type of pseudo-detector and other well-known types of detectors, such as the diode detector and synchronous detector, is beyond the scope of this paper; however, certain worthwhile observations on a qualitative basis can be made. Assuming a bandwidth prior to the detector of 2Δ radians per second or more, which is substantially larger than the bandwidth of the low-pass circuit following the detector of Δ radians per second width, which is a common occurrence, the following can be stated in terms of the dc output of the detector and the rms noise output following the detector. For a properly functioning synchronous or phase-lock detector, the dc voltage output is independent of the noise and is proportional to the carrier input. In the case of a symmetrical nonlinear pseudo-detector operating on a pulse train, the dc output is independent of the noise applied to it and is a function of the carrier input. In the case of a diode detector, the dc output is approximately proportional to the square root of the sum of the squares of the rms noise amplitudes and signal amplitudes applied to it, and is only proportional to the carrier level for high-input signal-to-noise ratios. For small input signal-to-noise ratios, the dc output due to the carrier is completely masked by the dc developed by the noise. The rms noise in the pass band following the synchronous detector is contributed only by the noise signals immediately adjacent to the carrier and within a total pass band of 2Δ radians per second. So far as the symmetrical nonlinear pseudo-detector is concerned, some low-frequency noise is contributed by all spectral areas prior to the detector. For the diode detector, low-frequency noise is contributed by all spectral areas prior to the detector.

Thus, the nonlinear symmetrical pseudo-detector is similar to the synchronous detector, inasmuch as its dc output is independent of the noise amplitude applied to it. On the other hand, it is similar to the diode detec-

[12] J. E. Bridges, U. S. Patent No. 2,628,326.

tor inasmuch as most of the low-frequency noise is contributed by all segments of the spectrum prior to the detector. Fortunately, this drawback can be somewhat overcome if the signal rate permits further reduction of the bandwidth of the filter circuit following the detector.

Another apparent disadvantage of the nonlinear symmetrical pseudo-detector can be overcome if the character of the signal is such that the aggregate of the detected signal and signal-and-noise products can be utilized. If this is permissible, the rate of signal fall-off of signals below threshold will parallel that of the synchronous detector.

So far as practical applications of this effect are concerned, in view of the limited spectrum space and recent successful developments in the field of synchronous or phase-lock communications systems, it does not appear that this type of signal transmission and detection would be of great interest. However, in certain specific fields where the below-threshold signals are important, where wide bandwidths prior to detection are inherent, where synchronous or phase-lock detection is either costly or not possible due to radically varying Doppler shifts, and where additional bandwidth is available for two or more harmonically related signals, this type of detection may be of value.

In the field of instrumentation, the following will illustrate this. Fig. 13 illustrates an application of this

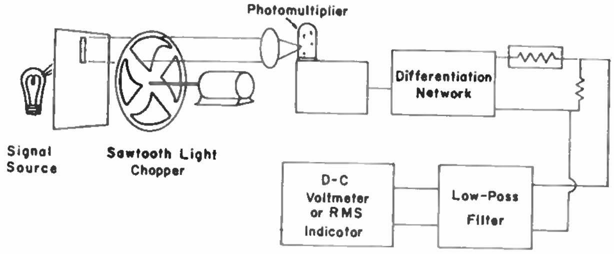


Fig. 13—A nonlinear resistor pseudo-detection circuit suitable for detecting very weak below threshold signals as applied to a photoelectric pickup device.

type of detector circuit to a photoelectric sensing device. Because of a large amount of nonsymmetrical noise present in a photomultiplier at the very-low frequencies, an additional precaution must be implemented. First, both the noise and signal must be suitably processed such that the noise applied to the nonlinear detector is symmetrical. Second, the processed signal must be in the form of a pulse train. The first requirement can be met by simply differentiating the output signal from the photomultiplier. The second requirement can be met by chopping the input radiation to the photomultiplier into the form of a saw-tooth signal as indicated. Care must be taken such that all circuits are linear except for the nonlinear detection circuit. This type of signal treatment is similar to that suggested by Balakrishnan and Drenick.[8]

A preliminary experimental evaluation of this circuit, using a Type 931A photomultiplier tube, indicated substantial improvements over either the common diode peak-riding or average detection process for comparable detector bandwidths.

Appendix

The current and voltage relationships of a nonlinear resistor can be given in a number of ways. The most common is

$$I = \left| K(V)^n \right|, \tag{13}$$

where I is the resistor current having same sign as V, K is a constant, V is the applied potential, and n is an exponent. Now assume that a rectangular pulse of current or voltage waveform is available which has no dc components and is defined as indicated in Fig. 3. From this the following mathematical relationships can be developed:

$r = \dfrac{t_1 - t_0}{t_2 - t_1}$, duty, cycle ratio;

$T = t_2 - t_0$, pulse repetition period;

I_{o-p}, V_{o-p}, zero-to-peak pulse height during pulse;

rI_{o-p}, rV_{o-p}, zero-to-peak signal during trace or pulse absence;

$(1 + r)I_{o-p}$, $(1 + r)V_{o-p}$, peak-to-peak pulse height;

$\dfrac{t_1 - t_0}{t_2 - t_0} = \dfrac{r}{1 + r}$, duration of pulse-to-period ratio;

$\dfrac{t_2 - t_1}{t_2 - t_0} = \dfrac{1}{1 + r}$, duration of trace-to-period ratio.

For the circuit of Fig. 4(a), a voltage source and a series nonlinear resistor, the following relationships can be developed:

$$I_p = K(V_{o-p})^n, \text{ current during pulse;} \tag{14}$$

$$I_t = K(rV_{o-p})^n, \text{ current during trace;} \tag{15}$$

$$I_{dc} = \frac{I_p(t_1 - t_0) - I_t(t_2 - t_1)}{t_2 - t_0}, \text{ average dc current;} \tag{16}$$

$$I_{d-c} = \frac{Kr(V_{o-p})^n - Kr^n(V_{o-p})^n}{1 + r}$$

$$= K(V_{o-p})^n \frac{(r - r^n)}{(1 + r)}; \text{ also, since } V_{o-p} = \frac{V_{p-p}}{1 + r}; \tag{17}$$

$$I_{dc} = \frac{K(V_{p-p})^n(r - r^n)}{(1 + r)^{n+1}}. \tag{18}$$

For the voltage source with a series nonlinear resistor and condenser load, the following technique was used: For a stable operating condition the charge acquired by the condenser during the pulse equals the charge lost during the trace period. Thus, the following relationships hold:

$$\int_{t_0}^{t_1} I_p(t)dt = \int_{t_1}^{t_2} I_t(t)dt \tag{19}$$

$$K(V_{o-p} - V_{dc})^n(t_1 - t_0) = K(rV_{o-p} + V_{dc})^n(t_2 - t_1) \tag{20}$$

$$(V_{o-p} - V_{dc})r^{1/n} = (rV_{o-p} + V_{dc})$$

$$V_{dc} = (V_{o-p}) \frac{r^{1/n} - r}{1 + r^{1/n}}. \tag{21}$$

For the current source with a parallel nonlinear resistor, the following relationships were developed:

$$V_p = K^{-1/n}(I_{o-p})^{1/n}, \text{ voltage output during pulse;} \tag{22}$$

$$V_t = K^{-1/n}(rI_{o-p})^{1/n}, \text{ voltage output during trace;} \tag{23}$$

$$V_{dc} = \frac{V_p(t_1 - t_0) - V_t(t_2 - t_1)}{t_2 - t_0}, \text{ average dc voltage output;}$$

$$V_{dc} = K^{-1/n}(I_{o-p})^{1/n} \frac{(r - r^{1/n})}{(1 + r)}. \tag{24}$$

For the current source in parallel with a nonlinear resistor and a conventional inductor, a condition of equilibrium is reached when the average voltage across the coil V_L is equal to zero. Thus,

$$\int_{t_0}^{t_1} V_L dt = \int_{t_1}^{t_2} V_L dt \tag{25}$$

$$K^{-1/n}(I_{o-p} + I_{dc})^{1/n}(t_1 - t_0) = K^{-1/n}(I_{o-p}r - I_{dc})^{1/n}(t_2 - t_1) \tag{26}$$

$$r^n I_{o-p} + I_{dc} r^n = I_{o-p} r - I_{dc} \tag{27}$$

$$I_{dc} = I_{o-p} \frac{(r - r^n)}{r^n + 1}. \tag{28}$$

To determine the approximate behavior of the nonlinear resistor detection circuit, the following procedure was used. Referring to Fig. 11(b), one sinusoidal interference generator is assumed to appear in one of the two peak responses of $Y_1(\omega)$ prior to the detector, and two harmonics of the signal are present so that signal amplitude $s(t) = a_1 \cos \omega_c t + a_2 \cos 2\omega_c t$, and a single sinusoidal interference signal is assumed to be $n(t) = b \cos yt$. If this voltage signal was applied to a nonlinear resistor with $n = 3$, the following current would result:

$$i(t) = K(a_1 \cos \omega_c t + a_2 \cos 2\omega_c t + b \cos yt)^3. \tag{29}$$

Carrying out the indicated operations and eliminating the higher-frequency terms, the following low-frequency terms result:

$$i_{l-f}(t) = \frac{3a_1^2 a_2}{4} + \frac{3a_1 a_2 b}{2} \cos(\omega_c - y)t + \frac{3}{4} a_2 b^2 \cos 2(\omega_c - y)t, \tag{30}$$

$$\text{for } \omega_c - \Delta/2 < y < \omega_c + \Delta/2;$$

$$i_{l-f}(t) = \frac{3a_1^2}{4} a_2 + \frac{3}{2} a_1 a_2 b \cos(\omega_c - y)t, \tag{31}$$

$$\text{for } \omega_c - \Delta < y < \omega_c - \Delta/2 \text{ and } \omega_c + \Delta/2 < y < \omega_c + \Delta;$$

$$i_{l-f}(t) = \frac{3}{4} a_1^2 a_2 + \frac{3}{4} a_1^2 b \cos(2\omega_c - y)t \tag{32}$$

$$\text{for } 2\omega_c - \Delta < y < 2\omega_c + \Delta.$$

In the case of thermal noise, a multiplicity of interfering signals is assumed, each within its discrete channel and phased at random with any set of points in time. For an approximate solution, the bandwidth prior to the detector, $Y_1(\omega)$, is divided into equally-spaced channels, such that the peak responses are divided into eight channels, each with a noise generator within it. This results in an input of the following form:

$$s(t) + n(t) = c_1 \cos \omega_c t + c_2 \cos 2\omega_c t + \sum_{n=0}^{n=3} a_n \cos\left[\left(\omega_c - \Delta\left(\frac{2n-3}{2}\right)\right)t + \phi_n\right] + \sum_{n=4}^{n=7} a_n \cos\left[\left(2\omega_c - \Delta\left(\frac{2n-11}{2}\right)\right)t + \phi_n\right]. \tag{33}$$

If this voltage is applied to a nonlinear detector, the following current would result. Assuming $N=3$ for simplicity and equal noise amplitudes, such that $a_n = a$,

$$i(t) = K[s(t) + n(t)]^3. \tag{34}$$

Carrying out the indicated operations results in a fairly large quantity of individual low-frequency terms, and if they are carried out in suitable tabular and chart form, the following coefficients of the various low-frequency amplitude terms (Table I) can be developed by collecting various categories of signals and combining each group by taking the square root of the sum-of-the-squares of the appropriate components. The data in Table I can be converted into terms of the input signal-to-noise ratio σ by noting

$$\sigma = \left[\frac{c_1^2 + c_2^2}{\sum_n a_n^2}\right]^{1/2}. \tag{35}$$

Assuming $c_2 = 0.7c_1$, then

$$\sigma = \sqrt{\frac{1.5}{8a^2}}, \quad a = \frac{c_1}{2.3\sigma}, \quad a^2 = \frac{c_1^2}{5.3\sigma^2}, \quad a^3 = \frac{c_1^3}{12.3\sigma^3}.$$

Table I can be further simplified by noting the above and by combining the various waveforms in any common category by taking the square root of the sum-of-the-squares of the different amplitudes (Table II).

In the case of Fig. 10, where

$$\Delta = 2b \text{ and } Y_2(\omega) = 1 \quad \text{for } \omega < 2b$$

$$Y_2(\omega) = 0 \quad \text{for } \omega > 2b,$$

TABLE I

Low-Frequency Channel	Signal	First-Order Noise	Second-Order Noise	Third-Order Noise
0 to b	$\frac{3}{4}C_1^2C_2$	$\frac{\sqrt{2.5}}{2}c_1^2a+\frac{\sqrt{6}}{2}c_1c_2a$	$\frac{\sqrt{12}}{2}c_1a^2+\frac{\sqrt{6}}{2}c_2a^2$	$\frac{\sqrt{35}}{2}a^3$
b to $2b$		$\frac{\sqrt{2.5}}{2}c_1^2a+\frac{\sqrt{6}}{2}c_1c_2a$	$\frac{\sqrt{19}}{2}c_1a^2+\frac{\sqrt{9.5}}{2}c_2a^2$	$\frac{\sqrt{29}}{2}a^3$
$2b$ to $3b$			$\frac{\sqrt{11}}{2}c_1a^2+\frac{\sqrt{5}}{2}c_2a^2$	$\frac{\sqrt{17}}{2}a^3$
$3b$ to $4b$			$\frac{\sqrt{6}}{2}c_1a^2+\frac{\sqrt{2.5}}{2}c_2a^2$	$\frac{\sqrt{6.5}}{2}a^3$
$4b$ to $5b$				$\frac{\sqrt{3.5}}{2}a^3$

TABLE II

Low-Frequency Channel	c^3f_1 Signal	$\frac{c^3f_2}{\sigma}$ First-Order Noise	$\frac{c^3f_3}{\sigma}$ Second-Order Noise	$\frac{c^3f_4}{\sigma}$ Third-Order Noise
0 to b	$0.53c^3$	$\frac{0.51c^3}{\sigma}$	$\frac{0.40c^3}{\sigma^2}$	$\frac{0.24c^3}{\sigma^3}$
b to $2b$		$\frac{0.51c^3}{\sigma}$	$\frac{0.46c^3}{\sigma^2}$	$\frac{0.22c^3}{\sigma^3}$
$2b$ to $3b$			$\frac{0.34c^3}{\sigma^2}$	$\frac{0.17c^3}{\sigma^3}$
$3b$ to $4b$			$\frac{0.25c^3}{\sigma^2}$	$\frac{0.10c^3}{\sigma^3}$

the dc output-to-noise ratio is

$$\frac{0.53\sqrt{2}}{\left[\left(\frac{.51}{\sigma}\right)^2+\left(\frac{.51}{\sigma}\right)^2+\left(\frac{.40}{\sigma^2}\right)^2+\left(\frac{.46}{\sigma^2}\right)^2+\left(\frac{.24}{\sigma^3}\right)^2+\left(\frac{.22}{\sigma^3}\right)^2\right]^{1/2}}. \tag{36}$$

The factor $\sqrt{2}$ in the numerator is necessary to correlate the heating or power effect of the dc signal to the remaining ac sinusoidal amplitudes. If a sinusoidal output or other type of modulation were used, this $\sqrt{2}$ factor would be correspondingly reduced. The signal and signal-and-noise to noise ratio is

$$\frac{\left[(0.53\sqrt{2})^2+\left(\frac{.51}{\sigma}\right)^2+\left(\frac{.51}{\sigma}\right)^2+\left(\frac{.40}{\sigma^2}\right)^2+\left(\frac{.46}{\sigma^2}\right)^2\right]^{1/2}}{\left[\left(\frac{.24}{\sigma^3}\right)^2+\left(\frac{.22}{\sigma^3}\right)^2\right]^{1/2}}. \tag{37}$$

The Amplitude Distribution and False Alarm Rate of Noise After Post-Detection Filtering*

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Summary—**A digital computer has been used to simulate the passage of white, Gaussian noise through a narrow-band filter, followed by a detector and a post-detection filter. The amplitude distribution of the output of the post-detection filter has been obtained for several different detectors. In addition, the variation of false alarm rates with detector law, post-detection filtering, time constants, and threshold setting has been investigated. Not only the numerical results but also the approach, and the new detectors described are believed of interest.**

INTRODUCTION

A DIGITAL simulation has been used to obtain the amplitude distribution and false alarm rate of noise after post-detection filtering for several different detectors. Simulation was chosen as the method of obtaining this distribution because it placed less of a burden upon the analyst than experimentation or the analytic methods of Kac and Siegert[1] or Emerson[2] and because it was more generally applicable to a variety of circuits than the Pearson type III distribution assumed by Bussgang, *et al.*[3] In particular, the square law and envelope detectors are looked at, along with some constant false alarm rate circuits—called "mean-level detectors" and "amplitude-ratio detectors." In addition, several post-detection filters have been studied.

The amplitude distribution and false alarm rate were obtained by using a digital computer (an IBM 709) to simulate a narrow-band filter, the detectors, and the post-detection filters through which noise (generated by a Monte Carlo method) was passed. The results include relations found between false alarm rate, amplitude-probability density, filter time constants, and threshold setting. After a summary of the results, the detectors and the methods used to analyze them are described. The product of the duration of the test in seconds times the noise bandwidth of the predetection filter is 318,200.

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[1] M. Kac and A. J. F. Siegert, "On the theory of noise in radio receivers with square-law detectors," *J. Appl. Phys.*, vol. 18, pp. 383–397; April, 1947.

[2] R. C. Emerson, "First probability densities for receivers with square-law detectors," *J. Appl. Phys.*, vol. 24, pp. 1168–1176; September, 1953.

[3] J. J. Bussgang, P. Nesbeda, and H. Safran, "A unified analysis of range performance of CW, pulse, and pulse Doppler radar," PROC. IRE, vol. 47, pp. 1753–1762; October, 1959.

RESULTS

The noise-amplitude distribution after detection, but before smoothing, is plotted in Figs. 1 and 2 (next page). Note that on semilog paper the noise distribution after the square-law detector follows a straight line, as is expected theoretically. Figs. 3–8 present the amplitude distribution after post-detection filtering. It is hoped that these data will be especially useful to those who wish to fit an empirical distribution function to the amplitude distribution of the noise after post-detection filtering.

It was observed that when a threshold was set sufficiently far from the mean level of the noise, the false alarm rate was proportional to the probability density of the noise amplitude. The constant of proportionality was the expected number of positive crossings of the mean divided by the probability density at the mean. Thus,

$$F = M\,\frac{Q(T)}{Q(0)}, \tag{1}$$

where F is the false alarm rate, M is the expected number of positive crossings of the mean, Q is the probability density of the noise amplitude, and T is the threshold setting in units of standard deviations measured from the mean. The observed number of crossings of the mean agrees with that predicted by Rice,[4] even though the distribution of the noise was non-Gaussian. The expected number of positive crossings of the mean, and thus the false alarm rate, is proportional to the geometric mean of the post-detection bandwidth and predetection bandwidth for the case of the single pole post-detection filter. Thus,

$$M_1 = k_1\sqrt{B/\tau}, \tag{2}$$

where B is the predetection noise bandwidth and τ is the post-detection time constant. When two post-detection filters with time constant τ were cascaded, M was found to be almost independent of predetection bandwidth and varied linearly with post-detection bandwidth For this case,

$$M_2 = k_2(1/\tau). \tag{3}$$

The k's are constants of proportionality.

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

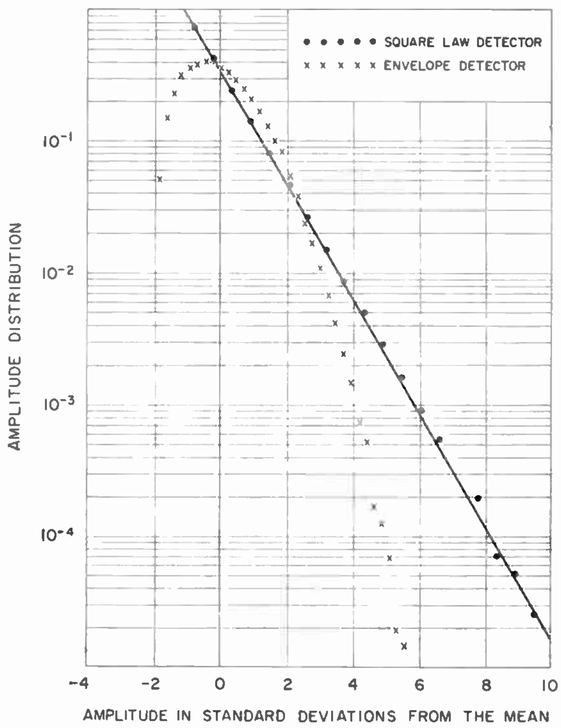

Fig. 1—Amplitude distribution after detection.

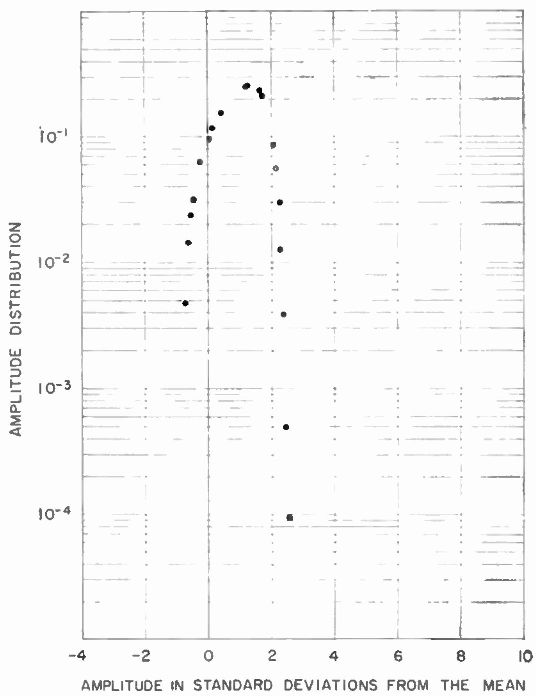

Fig. 2—Amplitude distribution after detection (amplitude-ratio detector).

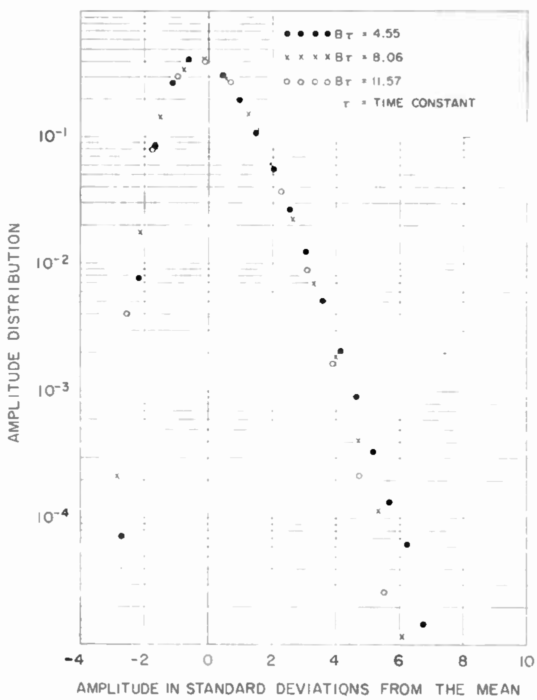

Fig. 3—Amplitude distribution after post-detection filtering (square-law detector—single pole filter)

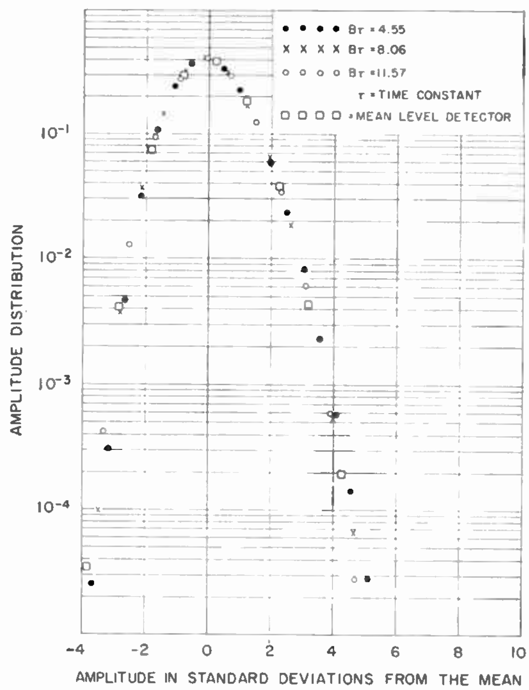

Fig. 4—Amplitude distribution after post-detection filtering (envelope detector—single pole filter).

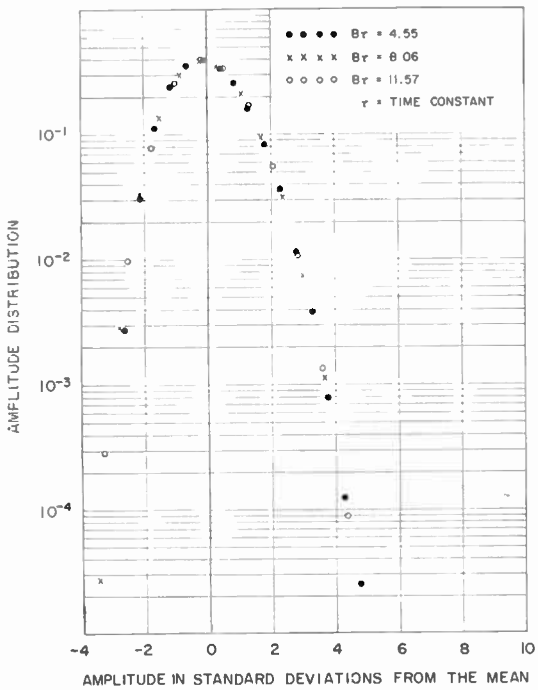

Fig. 5—Amplitude distribution after post-detection filtering (amplitude-ratio detector—single pole filter).

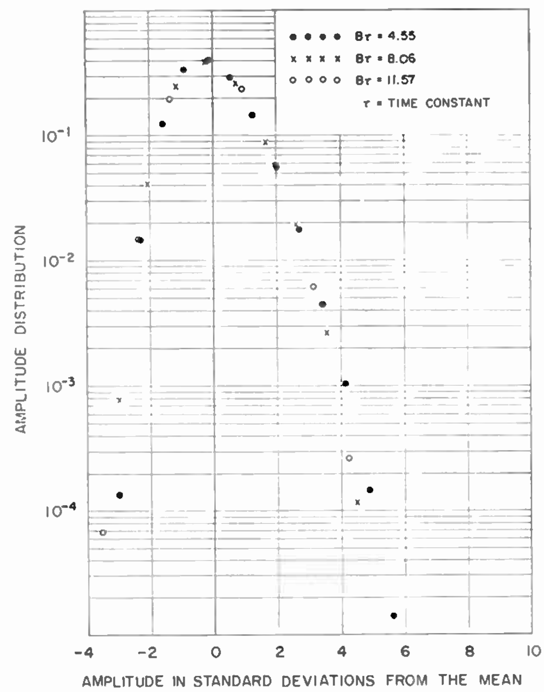

Fig. 6—Amplitude distribution after post-detection filtering (square-law detector—double pole filter).

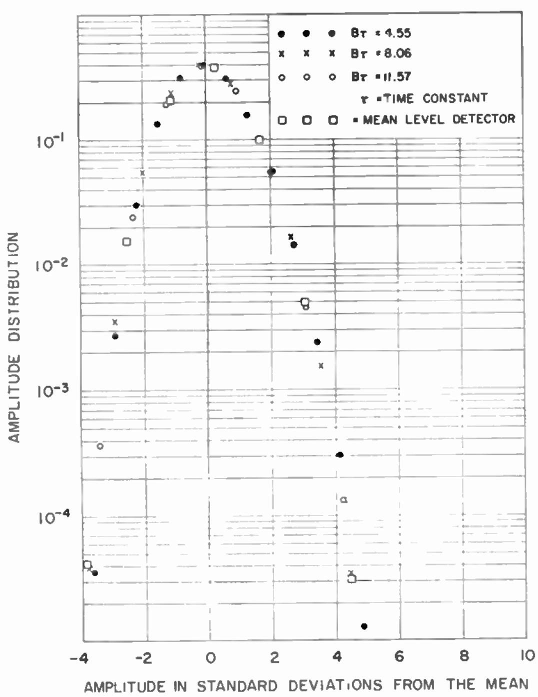

Fig. 7—Amplitude distribution after post-detection filtering (envelope detector—double pole filter).

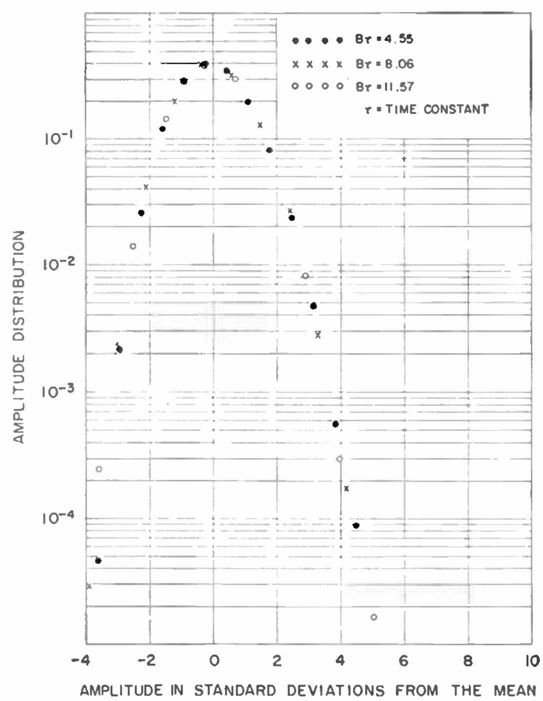

Fig. 8—Amplitude distribution after post-detection filtering (amplitude-ratio detector—double pole filter).

The Detection System

The detection system simulated consisted of a narrow-band noise source, a narrow-band predetection filter, one of several different detectors—square-law, amplitude-ratio, or envelope detector, one of several post-detection filter systems, and a threshold. A block diagram of the detection system is shown in Fig. 9.

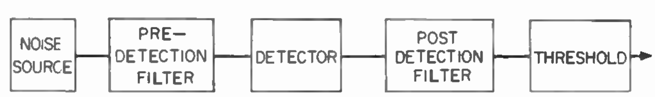

Fig. 9—The detection system.

A. Noise Source

A narrow-band voltage (*e.g.*, noise) is conveniently represented by[5]

$$N(t) = \alpha(t) \cos \omega t + \beta(t) \sin \omega t, \tag{4}$$

where α and β are two slowly varying random functions of time. The approximate center frequency of $N(t)$ is ω radians per second. If $N(t)$ represents a thermal noise voltage, α and β will be independent random functions of time having zero mean, Gaussian amplitude distribution, and bandwidth equal to half the bandwidth of $N(t)$. According to a sampling theorem developed by Shannon,[6] the waveforms α and β can be completely described by two sequences of numbers $\alpha(n/2W)$ and $\beta(n/2W)$ spaced $1/2W$ seconds apart, provided $\alpha(t)$ and $\beta(t)$ contain no frequencies greater than W. The waveform $\alpha(t)$ may be reconstructed from the sequence $\alpha(n/2W)$ by using the relation,

$$\alpha(t) = \sum_n \alpha(n/2W) \frac{\sin \pi(2Wt - n)}{\pi(2Wt - n)} \cdot \tag{5}$$

However, we shall use the representation

$$\alpha_1(t) = (1/2W) \sum_n \alpha(n/2W)\delta(t - n/2W), \tag{6}$$

where $\delta(x)$ is the unit impulse at $x=0$. The latter representation is preferred because it leads naturally to Z-transforms.

At this point it should be noted that the response of a low-pass circuit (whose gain at frequencies greater than W may be neglected) is the same to the inputs $\alpha(t)$ or $\alpha_1(t)$ since, as is easily shown, both have the same low-frequency spectrum. Without real loss in generality, we can restrict α so that it is zero for negative time. Then its Laplace transform is

$$A(s) = \int_0^\infty \alpha(t)e^{-st}dt. \tag{7}$$

Similarly, the Laplace transform of α_1 can then be written

$$A_1(s) = \int_0^\infty \alpha_1(t)e^{-st}dt. \tag{8}$$

By combining (6) and (8), one obtains

$$A_1(s) = (1/2W) \sum_{n=0}^{\infty} \alpha(n/2W)z^{-n}, \tag{9}$$

where $z=\exp(s/2W)$. The summation may be considered as defining the Z-transform, $A_1(z) = 2WA_1(s)$. It can be shown[6] that

$$A_1(s) = \sum_{k=-\infty}^{\infty} A(s + j4\pi kW). \tag{10}$$

It is immediately evident that $A_1(s)$ is a periodic function of $f=s/2\pi j$ with period $2W$ and has the same values as $A(s)$ for real frequencies less than W, provided $A(s)$ is negligible for real frequencies greater than W.

The generation of the sequences $\alpha(n/2W)$ and $\beta(n/2W)$ is the first step of the simulation program. α and β are independent and have a Gaussian distribution. Thus, the only restrictions on α and β are those needed to achieve a flat spectrum and insure a Gaussian amplitude distribution. The sampling interval, $1/2W$, was chosen so that the bandwidth of $N(t)$ was 2.83 times the noise bandwidth of the predetection filter—the part of the detection system which had the widest bandwidth. At this bandwidth (2.83 times the noise bandwidth) the filter attenuation was about 20 decibels and was increasing at about 12 decibels per octave.

B. Predetection Filter

The predetection filter was assumed to be a second-order Butterworth filter having a noise bandwidth B and an impulsive response

$$G(t) = 2h(t) \cos \omega t, \tag{11}$$

where $h(t)$ is the impulsive response of the low-pass prototype with transfer function

$$H(s) = \frac{8B^2}{s^2 + 4Bs + 8B^2} \cdot \tag{12}$$

In Appendix II, it is shown that the quadrature components of the output of the predetection filter are approximately

$$x(t) = \int_0^t \alpha(t - \tau)h(\tau)d\tau,$$

and

$$y(t) = \int_0^t \beta(t - \tau)h(\tau)d\tau. \tag{13}$$

The Laplace transform of x and y are

$$X(s) = H(s)A(s) = H(s)A_1(s),$$

and

$$Y(s) = H(s)B(s) = H(s)B_1(s), \tag{14}$$

where $A_1(s)$ may be used as an approximation of $A(s)$, provided $H(s)$ is negligible at real frequencies greater than W. Under these circumstances, for real frequencies less than W, $H(s)$ may also be approximated[7] by $H_1(z)$, a rational fraction in $z=\exp(s/2W)$. This procedure defines Z-transforms for the sample sequences $x(n/2W)$ and $y(n/2W)$. For example, if the approximation leads to

[5] See, for example, W. B. Davenport and W. L. Root, "Random Signals and Noise," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 158–159; 1958.

[6] J. R. Ragazzini and G. F. Franklin, "Sampled-Data Control Systems," McGraw-Hill Book Co., Inc., New York, N. Y.; ch. 2.

[7] R. Boxer and S. Thaler, "A simplified method of solving linear and nonlinear systems," Proc. IRE, vol. 44, pp. 89–101; January, 1956.

$$H_1(z) = \frac{a_0 + a_1z^{-1} + a_2z^{-2}}{b_0 + b_1z^{-1} + b_2z^{-2}}, \tag{15}$$

where $2WH_1(z)$ is approximately the Z-transform of the sequence $h(n/2W)$, then

$$x_1(z) = 2WH_1(z)A_1(z) = H_1(s)A_1(z). \tag{16}$$

Eq. (16) may then be interpreted by using the relation

$$b_0x(n/2W) + b_1x\left(\frac{n-1}{2W}\right) + b_2x\left(\frac{n-2}{2W}\right)$$
$$= a_0\alpha(n/2W) + a_1\alpha\left(\frac{n-1}{2W}\right) + a_2\alpha\left(\frac{n-2}{2W}\right). \tag{17}$$

A similar equation specifies the value of $y(n/2W)$ as a function of the values of $\beta(n/2W)$.

C. Detectors

Three different detector laws—square-law envelope, linear envelope, and amplitude-ratio—were simulated. (The mean-level detector consists of a bridge filter following a linear envelope detector.) Since the square of the envelope at time $n/2W$ is given by adding the squares of $x(n/2W)$ and $y(n/2W)$, the simulation of the first two detector laws is readily implemented. The amplitude-ratio detector, shown in Fig. 10, presents greater difficulty.

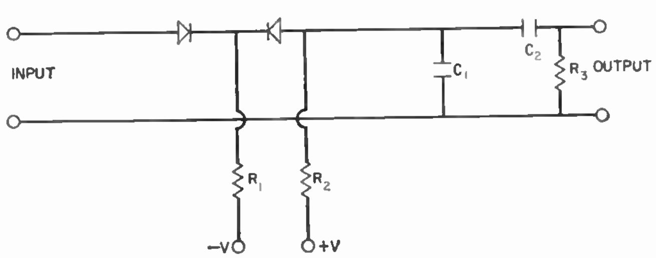

Fig. 10—Amplitude-ratio detector.

An approximate mathematical model can be constructed by assuming that the stray capacity of the diode junction may be neglected, that the currents $-V/R_1$ and V/R_2 are independent of the input voltage, and that the forward impedance of the diodes may be neglected. Under these assumptions, the current flowing toward the junction of C_1 and C_2 is $V/R_2 - V/R_1$, when the input voltage is more negative than that across C_1. Otherwise, the current is V/R_2. If this current is averaged over one cycle of the input (which is assumed to consist of a narrow-band waveform) one obtains an average current, w_3, given by

$$\begin{aligned} w_3 &= V/R_2 - (V/R_1)(1/\pi) \arccos (v_2/w_2), & w_2 > v_2 \\ &= V/R_2 & w_2 < v_2 \end{aligned} \tag{18}$$

where v_2 is the magnitude of the voltage across C_1 and w_2 is the envelope of the input signal. The Laplace transform of v_2 is given by

$$V_2(s) = (1/2C)\frac{s + 2/R_3C - 1}{s^2 + (2s/R_3C)} W_3(s), \tag{19}$$

where $C = C_1$ and $W_3(s)$ is the Laplace transform of $w_3(t)$.

It may be noted that the current w_3 is a function of the ratio w_2/v_2, which is the ratio of w_2, the present value of the input envelope, to v_2, an average of the envelope over the recent past. The name "amplitude-ratio detector" derives from this property. The amplitude-ratio detector and its associated filter perform, in many respects, similarly to a logarithmic receiver followed by differentiation of the detected output.

Using the techniques of Boxer and Thaler[7] we obtain the following set of equations:

$$\begin{aligned} w_3(n/2W) &= + V/R_2, - (V/\pi R_1) \arccos \frac{v_2(n/2W)}{w_2(n/2W)} & w_2 > v_2 \\ &= + V/R_2, & w_2 < v_2 \end{aligned}$$

$$\frac{v_2(n/2W) + f_1v_2\left(\frac{n-1}{2W}\right) + f_2v_2\left(\frac{n-2}{2W}\right)}{f_3w_2(n/2W) + f_4w_2\left(\frac{n-1}{2W}\right) + f_5w_2\left(\frac{n-2}{2W}\right)} = 1. \tag{20}$$

The numbers f_1, f_2, f_3, f_4, and f_5 result from the approximation,

$$(1/2C)\frac{s + 2/RC - 1}{s^2 + (2/RC)s} = \frac{f_3z^2 + f_4z + f_5}{z^2 + f_1z + f_2}. \tag{21}$$

Three sets of values of the f's were computed for the cases $BR_3C/2 = 4.55$, 8.06, 11.57, where B is the noise bandwidth of the predetection filter. In the actual computations, $v_2(n/2W)$ was replaced by $v_2[(n-1)/2W]$. This approximation produces very little error because the sampling period $1/2W$ is very small compared to the time constants associated with the amplitude-ratio detectors.

D. Post-Detection Filtering

The first stage of post-detection filtering used with the envelope and square-law detectors was assumed to consist of a simple series resistor-capacitor combination having a transfer function of the form,

$$V(s) = [1/(sRC + 1)]W(s), \tag{22}$$

where $V(s)$ and $W(s)$ are the transforms of the output and input voltages, respectively. The transfer function of the filter used with the amplitude-ratio detector is the same as that above, provided $W(s)$ is taken to be the transform of the input current $w_3(t)$. BRC was set equal to 4.55, 8.06, and 11.57. The mean-level detector shown in Fig. 11 has a bridge-type filter having transfer function,

$$k[1/(1 + RC_3s)] - 1/(1 + R_2C_2s), \tag{23}$$

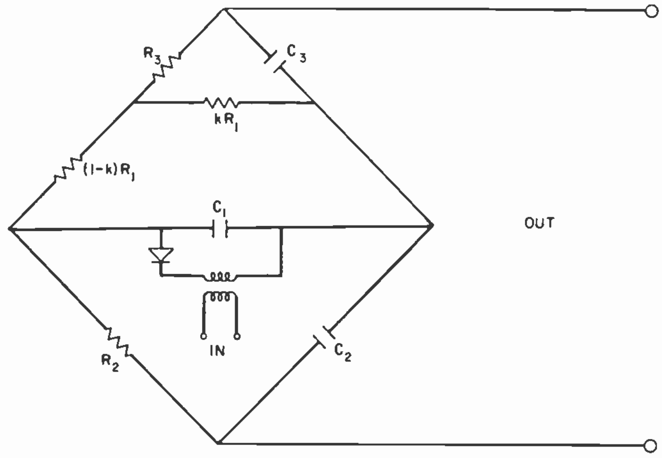

Fig. 11—The mean-level detector.

where $R=R_3+k(1-k)R_1$. This circuit was simulated for $k=\frac{2}{3}$, and $BRC_3=R_2C_2B/10=8.06$.

The time constant R_2C_2 is chosen rather long so that the voltage across capacitor C_2—which is largely proportional to the mean value of the noise—cannot follow fluctuations in the detector output, which are due to signal return from targets. The time constant RC_3 is chosen so that fluctuations because of target echoes appear across capacitor C_3 along with noise fluctuations due to noise power in the same frequency band. Since the noise fluctuations and the mean level of the noise are nearly proportional to each other, a suitable choice of k allows the direct current voltage across capacitors C_2 and C_3 in series to be used as a threshold bias. Since the two voltages vary together (proportionally), the false alarm rate is largely independent of noise amplitude and receiver gain.

A second stage of post-detection filtering was also employed. In all cases the transfer function of the second stage of filtering was that given by (22). The second stage had the same time constant as the first stage and was assumed not to load the first stage.

E. The Threshold

The outputs of the first and second stages of post-detection filtering were compared with a number of equally spaced thresholds and a count was made of the number of times each threshold was crossed. The fraction of the time each post-detection filter output (and the detector outputs as well) remained between two levels was also recorded, thus giving an experimental probability-density curve. All data were read out after the amplitudes were normalized so that the output amplitudes had an rms value of unity.

Appendix I

Approximating Electrical Filtering by Difference Equations

The approximation of electrical filtering by difference equations is one of the first steps to using a digital computer to simulate the operation of electric filters. Since one customarily describes filters by their transfer functions (that is, the Laplace transform of their impulse response) it is often quite convenient to use Z-Forms[7–9] to directly convert the transfer function to a difference equation. Customarily,

$$F(s) = G(s)H(s), \tag{24}$$

where $F(s)$, $H(s)$, and $G(s)$ are, respectively, the Laplace transforms of $f(t)$, the output; $h(t)$, the input; and $g(t)$, the impulsive response of the filter.

If we arrange $G(s)$ as a rational fraction in powers of $1/s$ and then substitute for s^{-n} the equivalent Z-form, the first three being

$$\frac{1}{s} = \frac{T}{2}\frac{z+1}{z-1},$$
$$\frac{1}{s^2} = \frac{T^2}{12}\frac{z^2+10z+1}{(z-1)^2},$$
$$\frac{1}{s^3} = \frac{T^3}{2}\frac{z^2+z}{(z-1)^3}, \tag{25}$$

the result is a rational fraction in z of the form

$$G_1(z) = \frac{A_m z^m + A_{m-1}z^{m-1} + \cdots + A_0}{z^m + b_{m-1}z^{m-1} + \cdots + b_0}. \tag{26}$$

The coefficients A_n, b_n may then be used to construct the difference relation sought:

$$\begin{aligned} f(nT) = &- b_{m-1}f(nT-T) - b_{m-2}f(nT-2T) - \cdots \\ &- b_0 f(nT-mT) + A_m h(nT) + A_{m-1}h(nT-T) \\ &+ \cdots + A_0 h(nT-mT). \end{aligned} \tag{27}$$

If the system is quiescent for negative values of time, $f(nT)=h(nT)=0$ for $nT<0$.

The formulation in (27) is especially applicable when the input $h(t)$ is available only as a sequence of sampled values $h(nT)$. In the foregoing, it is assumed that little information is lost if $g(t)$ and $h(t)$ are sampled every T seconds, *i.e.*, $g(t)$ and $h(t)$ contain little energy at frequencies greater than $W=1/2T$ cps.

Appendix II

Calculation of the Response of the Predetection Filter to Noise

The use of a digital computer to simulate the action of a predetection filter on noise is greatly facilitated if only the enevelope of the output of the filter is of interest. Fortunately, this is so, since the detector laws under consideration respond only to the envelope of narrow-band signals.

If a low-pass filter has a transfer function $H(s)$ and an impulsive response $h(t)$, then the band-pass equiva-

[8] J. R. Ragazzini and G. F. Franklin, "Sampled Data Control Systems," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 295–304.

[9] E. I. Jury, "Sampled Data Control Systems," John Wiley and Sons, Inc., New York, N. Y.

lent with transfer function,

$$G(s) = H\left(\frac{s^2+\omega^2}{2s}\right), \tag{28}$$

has an impulsive response which is given approximately by

$$g(t) = 2h(t)\cos\omega t. \tag{29}$$

[If we assumed the band-pass equivalent to be $G(s)=H(s+j\omega)+H(s-j\omega)$, then (29) would have been immediately evident from the inverse Laplace transform of $G(s)$.] To demonstrate the validity of (29), let us recall that the impulsive response is given by

$$g(t) = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} G(s)e^{st}ds. \tag{30}$$

Utilizing (28), we have

$$g(t) = \frac{1}{2\pi j}\int_{-j\infty}^{j\infty} H\left(\frac{s^2+\omega^2}{2s}\right)e^{st}ds. \tag{31}$$

Now $H(z)$ is supposed to be quite small for z large so that the bulk of the contributions to the integral occur for s near $\pm j\omega$. As a result, the impulsive response is given approximately by

$$g(t) \cong \frac{1}{2\pi j}\int_{-j(\omega+A)}^{-j(\omega-A)} H\left(\frac{s^2+\omega^2}{2s}\right)e^{st}ds + \frac{1}{2\pi j}\int_{j(\omega-A)}^{j(\omega+A)} H\left(\frac{s^2+\omega^2}{2s}\right)e^{st}ds, \tag{32}$$

where A is much less than ω but much larger than the bandwidth of the filter. Now, let $s=-j\omega+x$ in the first integral and $s=j\omega+x$ in the second integral, obtaining

$$g(t) \cong \frac{e^{-j\omega t}}{2\pi j}\int_{-jA}^{+jA} H\left[\frac{x(x-2j\omega)}{2(x-j\omega)}\right]e^{xt}dx + \frac{e^{-j\omega t}}{2\pi j}\int_{-jA}^{jA} H\left[\frac{x(x+2j\omega)}{2(x+j\omega)}\right]e^{xt}dx. \tag{33}$$

For A much less than ω,

$$(x-2j\omega)/2(x-j\omega) \cong 1 \cong (x+2j\omega)/2(x+j\omega),$$

so that $g(t)$ becomes

$$g(t) \cong \frac{\cos\omega t}{\pi j}\int_{-jA}^{jA} H(x)e^{xt}dx. \tag{34}$$

For A much larger than the bandwidth of the filter, the integral in (34) is close in value to $2\pi jh(t)$, so that

$$g(t) \cong 2h(t)\cos\omega t, \tag{35}$$

which is the desired result. The above heuristic demonstration can be made rigorous by evaluating the errors associated with the several approximations made.

A narrow-band signal or noise-plus signal can be represented by

$$n(t) = \alpha(t)\sin\omega t + \beta(t)\cos\omega t, \tag{36}$$

where ω is the approximate frequency of the signal and $\alpha(t)$ and $\beta(t)$ change very little over a time as short as one period of the signal ($T=2\pi/\omega$). If $n(t)$ is the input to a filter having the impulsive response $g(t)$, the output is given by

$$m(t) = \int_0^t n(t-\tau)g(\tau)d\tau. \tag{37}$$

Substituting from (29) and (36), we obtain

$$m(t) = 2\int_0^t [\alpha(t-\tau)\sin\omega(t-\tau)h(\tau)\cos\omega\tau + \beta(t-\tau)\cos\omega(t-\tau)h(\tau)\cos\omega\tau]d\tau. \tag{38}$$

If α and β change only slowly during a time equal to $2\pi/\omega$, $m(t)$ is given approximately by

$$m(t) \approx \left[\sin\omega t\int_0^t \alpha(t-\tau)h(\tau)d\tau + \cos\omega t\int_0^t \beta(t-\tau)h(\tau)d\tau\right], \tag{39}$$

and the square of the envelope of the output of the filter is given by

$$R^2(t) \cong \left[\int_0^t \alpha(t-\tau)h(\tau)d\tau\right]^2 + \left[\int_0^t \beta(t-\tau)h(\tau)d\tau\right]^2. \tag{40}$$

The quadrature components of $m(t)$ are

$$x(t) = \int_0^t \alpha(t-\tau)h(\tau)d\tau,$$

$$y(t) = \int_0^t \beta(t-\tau)h(\tau)d\tau. \tag{41}$$

The two quadrature components of noise are known to have amplitudes that are distributed according to the normal law. Therefore, if $n(t)$ represents noise, we need only generate two independent random sequences of noise representing the amplitdues $x(t)$ and $y(t)$ of the two quadrature components of noise, and then separately smooth each component by passing it through a low-pass filter having the same shape (considering negative as well as positive frequencies) and bandwidth as the band-pass filter. The smoothing was performed using a difference equation derived from the low-pass transfer function by replacing powers of $1/s$ by Z-forms as described in Appendix I.

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IRE Standards on Radio Transmitters: Definitions of Terms, 1961*

61 IRE 15. S1

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Alternator Transmitter. A radio transmitter which utilizes power generated by a radio-frequency alternator.

Amplitude-Modulated Transmitter. A transmitter which transmits an amplitude-modulated wave.

Arc Converter. A form of negative resistance oscillator utilizing an electric arc as the negative resistance.

Automatic Frequency Control. An arrangement whereby the frequency of an oscillator or the tuning of a circuit is automatically maintained within specified limits with respect to a reference frequency.

Balanced (Push-Pull) Amplifier. An amplifier circuit in which there are two identical signal branches connected so as to operate in phase opposition and with input and output connections each balanced to ground.

Baud. A unit of signaling speed. The speed in bauds is the number of signal elements per second.

Buffer Amplifier. An amplifier designed to isolate a preceding circuit from the effects of a following circuit.

Carrier-Frequency Range of a Transmitter. The continuous range of frequencies within which the transmitter may be adjusted for normal operation. A transmitter may have more than one carrier-frequency range.

Class A Modulator. A Class A amplifier which is used specifically for the purpose of supplying the necessary signal power to modulate a carrier.

Class B Modulator. A Class B amplifier which is used specifically for the purpose of supplying the necessary signal power to modulate a carrier.

Electron-Coupled Oscillator. An oscillator employing a multigrid tube in which the cathode and two grids operate in any conventional manner as an oscillator and in which the load is in the plate circuit.

Feedback. In a transmission system or a section thereof, the returning of a fraction of the output to the input.

Fixed-Frequency Transmitter. A transmitter designed for operation on a single carrier frequency.

Fixed Transmitter. A transmitter that is operated in a fixed or permanent location.

Frequency Band. A continuous range of frequencies extending between two limiting frequencies.

Frequency-Modulated Transmitter. One which transmits a frequency-modulated wave.

High-Level Modulation. Modulation produced at a point in a system where the power level approximates that at the output of the system.

Impulse (Shock) Excitation. A method of producing oscillator current in a circuit in which the duration of the impressed voltage is relatively short compared with the duration of the current produced.

Linear Rectifier. A rectifier, the output current or voltage of which contains a wave having a form identical with that of the envelope of an impressed signal wave.

Low-Level Modulation. Modulation produced at a point in a system where the power level is low compared with the power level at the output of the system.

Master Oscillator. An oscillator so arranged as to establish the carrier frequency of the output of an amplifier.

Modulation Index. In angle modulation with a sinusoidal modulating wave, the ratio of the frequency deviation to the frequency of the modulating wave.

Multi-RF-Channel Transmitter. A radio transmitter having two or more complete radio-frequency portions capable of operating on different frequencies either individually or simultaneously.

Multi-Frequency Transmitter. A radio transmitter capable of operating on two or more selectable frequencies, one at a time, using preset adjustments of a single radio-frequency portion.

Phase-Modulated Transmitter. A transmitter which transmits a phase-modulated wave.

Radio Broadcasting. Radio transmission intended for general reception.

Ring Oscillator. An arrangement of two or more pairs of tubes operating as push-pull oscillators around a ring, usually with alternate successive pairs of grids and plates connected to tank circuits. Adjacent tubes around the ring operate in phase opposition. The load is supplied by coupling to the plate circuits.

Service Band. A band of frequencies allocated to a given class of radio service.

Shock Excitation. See *Impulse Excitation.*

Simplex Operation of a Radio System. A method of operation in which communication between two stations takes place in one direction at a time. *Note:* This includes ordinary transmit-receive operation, press-to-talk operation, voice-operated carrier and other forms of manual or automatic switching from transmit to receive.

Spark Transmitter. A radio transmitter which utilizes the oscillatory discharge of a capacitor through an inductor and a spark gap as the source of its radio-frequency power.

Spurious Transmitter Output. Any part of the radio-frequency output which is not implied by the type of modulation (AM, FM, etc.) and specified bandwidth.

Spurious Transmitter Output Conducted. Any spurious output of a radio transmitter conducted over a tangible transmission path. *Note:* Power lines, control leads, radio frequency transmission lines and waveguides are all considered as tangible paths in the foregoing definition. Radiation is not considered a tangible path in this definition.

Spurious Transmitter Output Extraband. Spurious output of a transmitter outside of its specified band of transmission.

Spurious Transmitter Output Inband. Spurious output of a transmitter within its specified band of transmission.

Spurious Transmitter Output Radiated. Any spurious output radiated from a radio transmitter. *Note:* The radio transmitter does not include the associated antenna and transmission lines.

The Quadratic Invariances of a Generalized Network*

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Summary—**The Manley-Rowe relation, as applied in the small signal linearized approximation, may be stated as a quadratic form that is invariant under the operation of the system. It is, however, only one of the set of such forms that is invariant through a given type of system.**

It is shown that the existence of quadratic invariances is a consequence of the eigenvalues of the system operator being either of unit magnitude or else grouped in pairs such that one is the conjugate reciprocal of the other. If this condition applies, then there exists at least *n* such linearly independent forms, where *n* is the number of degrees of freedom of the system. Each form then specifies a quantity that is conserved by the system.

Methods of determining the quadratic invariant forms from the matrix operation of the system are developed. Application is made to certain simple two-port networks to illustrate the analysis and the significance of the resulting invariances. Parametric circuits are also studied. The Manley-Rowe relation is found, as expected. Other relations, applicable to subclasses of such networks are also found.

Finally, application is made to a lossy parametric shunt element, such as an imperfect nonlinear capacity. The quadratic invariances for such a device, for the two-frequency case, are derived.

Introduction

OUR purpose here is to study the quadratic invariances of the system operator of a generalized two-port network. These invariances are significant in that they specify many of the important properties of such systems. Among the known examples of such invariances may be cited the conservation of energy in conventional lossless circuits, the Manley-Rowe relation in parametric circuits,[1] and Chu's kinetic power theorem in electron beam devices.[2–4] It is our intention to determine when relations of this form which are invariant in the given system exist, and to develop methods whereby they can be found.

We shall be concerned with systems in which the operator is linear, homogeneous, and finitely dimensioned. In some of the situations of interest, linearity is obtained only in the small signal approximations, *e.g.*, parametric circuits and electron beams. Nevertheless, we may hope that such an approach will serve to unify the theory of these physically diverse situations, and will provide additional insight into their behavior.

We shall consider as our central problem that of finding the relation between an operator and its invariants. We will ask what the invariants of a given operator are.[5]

An important part of conventional two-port network theory can be described as the manipulation of such invariants. The whole class of reactive circuits is that for which the power flow is invariant. A subclass is defined as that for which a given impedance is invariant, *i.e.*, that have a particular characteristic impedance, etc. While a discussion of conventional circuits in terms of such invariants will seem, perhaps, unusual, it is useful since, as we shall show, it is generalizable to more complex situations such as parametric networks.

We shall develop the theory first on a formal basis, so as to obtain results which will be applicable to a broad range of problems. We shall be able to show the conditions under which invariants of the types being sought exist, and to obtain a procedure for their determination.

We shall then apply this formalism to conventional two-port networks by considering examples which illustrate the principal situations that arise. We shall do this, not so much in the hope of adding much of significance to this body of knowledge, but to give a well-known reference point from which this method of approach can be better understood.

We shall finally apply this theory to circuits which are obtained from conventional reciprocal two-port networks by periodic variation of their parameters.

We shall show that the Manley-Rowe power theorem depends upon one of these invariants. The generality of this theorem derives from the fact that the invariant is applicable to the entire class of circuits obtained by the parametric variation of reactive networks.

Finally, we shall study the invariances of a shunt parametric capacity in parallel to a resistance. This describes the parametric units that are actually available[6] more adequately, and is a circuit element to which the Manley-Rowe theorem does not apply. We shall, however, be able to obtain analogous invariants.

I. Abstract Development

A. State Vector

We start by defining an *n*-dimensional vector that describes the state that exists at a port. We express such

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[1] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements, Part I. General energy relations," PROC. IRE, vol. 44, pp. 904–913; July, 1956.

[2] L. J. Chu, "A Kinetic Power Theorem," presented at the IRE-PGED Electron Tube Res. Conference, Durham, N. H.; June, 1951.

[3] H. A. Haus, "Electron Beam Waves in Microwave Tubes," Mass. Inst. Tech., Rept. No. 316; April 8, 1958.

[4] H. A. Haus, "Signal and Noise Propagation Along Electron Beams," in "Noise in Electron Devices," L. D. Smuller and H. A. Haus, Eds., Technology Press, Cambridge, Mass., 1959; and John Wiley and Sons, Inc., New York, N. Y., 1959.

[5] For the theory of quadratic forms and invariants in general, see L. E. Dickson, "Algebraic Theories," Dover Publications, Inc., New York, N. Y.; 1959.

[6] A. Uhlir, Jr., "Progress Report on Varactors," presented at the Natl. Symp. on Microwave Theory and Technique, Cambridge, Mass.; June 1–3, 1959.

a vector as an n-row column vector.

$$\mathbf{x} = \begin{bmatrix} \cdot \\ \cdot \\ \cdot \\ x_n \\ \cdot \\ \cdot \\ \cdot \end{bmatrix}.$$

The components of this vector may be any suitable quantities. For most purposes, the time dependency need not be included. It is sufficient if the x_n include only the amplitudes of the various components. These components may be the voltages and currents at the various frequencies of interest, or they may be the forward and backward waves. In the electron beam case, they may be the amplitudes of any sufficient set of independent variables, *e.g.*, voltage and current on the circuit and RF velocity and space charge on the beam.

Furthermore, we may note that the definition of "port" is largely arbitrary. We may, for example, regard the entire circuit as being fed through sundry circulators from some source, and connected to some load through the other terminals of the circulators. In this case, the entire set of inwardly directed waves constitutes the state at one port, and the set of outwardly directed waves the other state. This gives us the "scattering" description of the system.

The totality of voltages may describe one state, and the totality of currents the other. This gives us either the impedance or admittance description, depending upon which we take as the load port.

The term "load" is used here in a somewhat loose manner. We are not precluding the possibility that the "load" may supply power. It is arbitrary which port we call the load.

Broadly speaking, then, we divide the whole set of external parameters into two groups, calling one the first state $\mathbf{x}_1$, and the other the second state $\mathbf{x}_2$. Although it is not strictly necessary, we assume that this division has been made in an equal fashion, so that the (finite) dimensionality of $\mathbf{x}_1$ is the same as that of $\mathbf{x}_2$.

B. The State Operator, L

Given a definition of the two state vectors of a system, then, we assume that we know the operator, L, that connects these vectors.

$$\mathbf{x}_1 = L\mathbf{x}_2. \tag{1}$$

We assume that this operator is linear and homogeneous. Since its dimensionality is finite, it may be expressed in matrix form. It is, then, a square $n \times n$ matrix.[7]

[7] For general matrix theory and manipulation, see Richard Bellman, "Introduction to Matrix Analysis," McGraw-Hill Book Co., Inc., New York, N. Y.; 1960.

C. Quadratic Invariances

Given such a description, then, we wish to determine if there are invariant forms whose values are unchanged by the operation, L. One such invariance is, for example, the net real power flow into a reactive network. Its value at the output is the same as at the input, and it is, therefore, "invariant under the operation L."

This particular invariant can be written as a quadratic form. If the state vector is written in terms of voltages and currents, then the net real power flow at one generalized port is the sum of terms of the form $\frac{1}{2}(EI^* + E^*I)$. More generally, if the state vector is otherwise defined, then it is the sum of terms which involve a constant, times one component of the state vector, times the conjugate of a second component. We can then write it in the form

$$s(K) = \mathbf{x}^\dagger K\mathbf{x}, \tag{2}$$

where K is an $n \times n$ square matrix. The symbol ($\dagger$) is used to indicate the "Hermitian conjugate," the complex conjugate of the transposed matrix. Since $\mathbf{x}$ is a column matrix, $\mathbf{x}^\dagger$ is a row matrix, and $s(K)$ is a scalar. The conservation of real power flow states that this form is invariant under the operation L if K is such that the form yields the proper expression for the real power.

It is, then, reasonable to ask if there are other matrices for which $s(K)$ is invariant under L. In fact, our example was based on a reactive network. We may ask what happens in the presence of loss. Are there appropriate K's in this case? Again, in parametric circuits for which the signal power is negligible compared to the pump, the system may be approximated as a linear system in the signal, idle, and, perhaps, other frequencies. The approximate system, then, does not conserve power (since we do not keep track of the pump power) but does conserve the power normalized by the frequency in accordance with the Manley-Rowe relation. This, again, is expressible as a quadratic form.

Geometrically, we normally regard the scalar $\mathbf{x}^\dagger\mathbf{x}$ as giving the square of the length of $\mathbf{x}$. This, however, implies a "Euclidean space," and there is nothing in the definition of the state vector, $\mathbf{x}$, which determines the nature of the space in which it is "embedded." Its components may have different units. They may be at different frequencies. The term "length" has no obvious physical significance. We are, therefore, not bound to any particular type of space and may choose it as is convenient.

An arbitrary space is specified by its "metric." The metric tells us how the length of a vector is to be computed, given the components of the vector. The formula that connects these is precisely (2). If K is the metric of the space in which $\mathbf{x}$ is embedded, then $s(K)$defined by (2), is by definition the square of the length of $\mathbf{x}$.

Our problem can be described as asking whether we can so choose the space in which we consider x to be embedded that its length is unchanged by the operation L. Given L, do there exist one or more spaces in which it describes a rotation?

As the first step in this problem, we can substitute (1) into (2), and require that the result be independent of x:

$$x_2^\dagger K x_2 = x_1^\dagger K x_1 = (Lx_2)^\dagger K(Lx_2) = x_2^\dagger L^\dagger K L x_2, \tag{3}$$

since the conjugate transpose of a product of matrices is the product of the conjugate transposes in reverse order.

If (3) is to be true for any x_2, then we must have

$$L^\dagger K L = K. \tag{4}$$

Conversely, given L, any K which satisfies (4) yields an $s(K)$ which is invariant under L.

This does not tell us whether or not such a K matrix exists. It says only that, if it does, it generates an invariant $s(K)$ which is conserved by the system.

D. The Condition for the Existence of K-Matrices

Our first concern must be to determine the conditions necessary for the existence of suitable K matrices for an operator L. These can be obtained directly. If we restrict our attention to nonsingular K's, then we can write (4) in the form

$$L^\dagger = KL^{-1}K^{-1}. \tag{5}$$

$L^\dagger$ is seen to be related to L by a "similarity transformation." It is well-known, from elementary matrix theory, that the eigenvalues of a matrix are unchanged by a similarity transformation. Hence, the eigenvalues of $L^\dagger$ must be the same as those of L^{-1}. However, the eigenvalues of $L^\dagger$ are clearly the conjugates of the eigenvalues of L, and those of L^{-1} are the reciprocals of those of L. Hence, a necessary condition for the existence of a nonsingular K is that every eigenvalue of L be either its own conjugate reciprocal, *i.e.*, of unit magnitude, or else can be paired with another that is its conjugate reciprocal. That is, for every λ_n that is an eigenvalue of L, either

$$\lambda_n\lambda_n^* = 1, \tag{6}$$

or else that there exists another eigenvalue λ_m with the same degree of multiplicity such that:

$$\lambda_n\lambda_m^* = 1. \tag{7}$$

If K is allowed to be singular, so that K^{-1} does not exist, this condition can be weakened. This does not appear to be a situation of interest, however.

We have shown, here, that the condition (6) and (7) is necessary. That it is also sufficient will be shown by the formal determination of the K's.

E. The Determination of the K-Matrices

To determine the K-matrices, we proceed by the usual method of diagonalization. The eigenvalues of L are determined as the roots, λ_i, of the determinantal secular equation

$$|L - \lambda I| = 0. \tag{8}$$

To each such value λ_i, there corresponds at least one eigenvector x_i, defined by

$$Lx_i = \lambda_i x_i. \tag{9}$$

We shall assume that the set $\{x_i\}$ is complete, *i.e.*, that it is a set of n linearly independent vectors, in terms of which any n-dimensional vector can be expressed. This is not always true, but is so in most cases of interest. The more general case in which the Jordan canonical form[8] is convenient, is considered briefly in the Appendix.

If this is so, then we may define a matrix T as:

$$T = (x_1 x_2 \cdots x_n). \tag{10}$$

That is, T is an n-row matrix whose components are n-term column matrices. Hence, T is an $n \times n$ square matrix.

Then, as is well-known from matrix theory,

$$L_1 = T^{-1}LT \tag{11}$$

is a diagonal matrix, whose elements are the eigenvalues of L in the order in which the corresponding eigenvectors appear in T.

The matrix T, we may note, is not defined uniquely. The eigenvectors may be taken in any order. Furthermore, they may each be multiplied by a scalar factor. However, this is unimportant. It implies that we can find at least one nonsingular T that diagonalizes L.

If we substitute (7) into (4) it can be written

$$L_1^\dagger K_1 L_1 = K_1, \tag{12}$$

where K_1 is a transformed K.

$$K_1 = T^\dagger K T \tag{13}$$

The matrix $L_1^{\dagger -1}$, if L obeys the conditions deduced in the preceeding section, has the same elements as L_1 since these are both diagonal matrices whose components on the diagonal are the eigenvalues of L. These may be arranged in different order, however. If λ_n is of unit magnitude, it will occupy the same position in both. If, however, $\lambda_n\lambda_m^* = 1$ with $n \neq m$, then the positions of λ_n and λ_m will be interchanged.

We can easily find a permutation matrix which, when used in a similarity transformation, will correct this in-

[8] For a detailed discussion of the Jordan canonical form, see B. Friedman, "Principles and Techniques of Applied Mathematics," John Wiley and Sons, Inc., New York, N. Y., ch. 2, pp. 57–130; 1956.

terchange of position. If, for example, λ_n and λ_m are interchanged, the matrix which has the (nn) and the (mm) terms equal to zero, and the (n, m) and (m, n) terms equal to unity, will interchange them. This type of permutation matrix (built up of 2-cycles) is also convenient in that it is its own reciprocal.

Hence, we can easily find P such that

$$L_1^{\dagger-1} = P^{-1}L_1P$$

or

$$L_1^{\dagger} = P^{-1}L_1^{-1}P, \tag{14}$$

since the inverse of a product is the product of the inverses in reverse order.

Substituting (14) into (12) and rearranging,

$$K_2L_1 = L_1K_2, \tag{15}$$

where

$$K_2 = PK_1 = PT^{\dagger}KT. \tag{16}$$

Eq. (15) is a commutation law. Any matrix K_2 which commutes with L_1 is acceptable and, in turn, by (16) determines an appropriate K for which $s(K)$ is invariant under L.

Since L_1 is diagonal, any diagonal matrix commutes with it. There are n such forms which are linearly independent. These may be chosen so that all are nonsingular. (For example, we may take the set where we put $+1$ everywhere on the diagonal except for one position at which we put -1. All such are nonsingular. They are linearly independent, and there are n of them, depending on the position of the minus sign.)

It may be easily shown that, if the K_2's are nonsingular and independent, then so are the K's. Hence, if the conditions on the eigenvalues of L are obeyed, there are at least n matrices leading to n independent invariant quadratic forms; however, there may be more.

If L is degenerate so that more than one eigenvector has the same eigenvalue, then L_1 will contain elements that repeat along the diagonal. In this case, K_2 may also be any permutation matrix among the repeating elements. Each such additional K_2 leads to a new K which is linearly independent of the others.

This procedure shows that, if the conditions on the eigenvalues of L are satisfied, then there will indeed exist appropriate metrics which will generate quadratic forms whose values are invarient in the system. It shows, further, that there exists a whole set of such metrics, each leading to its own conservation law. If the system has n degrees of freedom, there will be at least n such conservation laws, and more if the system is degenerate.

The procedure outlined also gives a method for determining these invariances. It is often simpler to find them by direct substitution in (4). The computation of the eigenvectors of L, necessary for the construction of T, and the determination of the inverse of T, are tedious processes if the system is at all complicated. Nevertheless, it is significant that a systematic method can be found and is available if necessary.

F. Change of Basis

Before proceeding with specific examples, we will consider two subsidiary problems. The first is the behavior of these invariants under a change of basis. Secondly, we shall consider whether or not, or when, K should be restricted to Hermitian form ($K^{\dagger}=K$).

Considering first a change of basis, this is, in essence, a change in the manner of describing the system. If, for example, we are given L as relating states that are described in terms of voltages and currents, we may find it convenient to transform to the operator, M, that relates the forward and backward waves at the same two ports (the "wave" representation), the waves being defined relative to a normalizing characteristic impedance. Or, if we have a wave representation, we may wish to transform to a new representation with a different wave normalization. We may even find it convenient to use a representation that combines a voltage at one frequency with a current at another in a kind of "generalized" wave representation.

All of these changes may be written as

$$x = Rv, \tag{17}$$

where R is an $n\times n$ nonsingular matrix, x is on the given basis, and y is the same vector expressed on the new basis.

Then on the new basis, the operator becomes the matrix M, which is related to L by

$$M = R^{-1}LR. \tag{18}$$

Substituting (18) in (4), we see that K is transformed into a new associated operator, J, which is related by

$$J = R^{\dagger}KR, \tag{19}$$

and the resultant scalar invariant becomes

$$s(J) = y^{\dagger}Jy. \tag{20}$$

Thus, it is a straightforward matter to obtain the invariants on one basis, given the solution on another. The Manley-Rowe relation, for example, can easily be converted to an expression in terms of waves or generalized waves. The existance of a given conservation law that is appropriate to the system is independent of the description of the system, and may be translated directly from one type of description to another without difficulty.

G. Hermitian K's

Finally, we shall consider the form of K. If K equals its own Hermitian conjugate,

$$K = K^{\dagger},$$

then it is said to be of Hermitian form. Such a form is the matrix analog of real numbers.

Our development of the theory has nowhere required K to be Hermitian. In fact, the outlined procedure will, in general, yield non-Hermitian K's. This follows since we will generally choose the K_2's with real numbers on the diagonal. A Hermitian K_2 does not assure a Hermitian K.

From the point of view of the geometric interpretation, a non-Hermitian K is awkward since $s(K)$ is then not necessarily real. Since $s(K)$ is interpreted as the square of the length, it is somewhat disconcerting to realize that, for some vectors x, it may be complex. While this is unimportant in the formal sense, it does suggest that it might be worthwhile to consider K's such that $s(K)$ will always be at least real. The condition that this requires is that K be Hermitian.

We note that (4) is linear in the K's. If we have two K's which satisfy (4), then so does any linear combination of them. Furthermore, if K satisfies (4), then so does $K^\dagger$, as may be seen by taking the Hermitian conjugate of (4). Hence, if K satisfies (4), then so does K' and K''. Where

$$\begin{aligned} K' &= K + K^\dagger \\ K'' &= j(K - K^\dagger), \end{aligned} \tag{21}$$

K' and K'' are both Hermitian.

Hence, given a set of K's, which may not be Hermitian, we can find a set of Hermitian K's. Our set of conservation laws, then, can always be changed into a set that is non-negative definite.

II. Normal Networks

We shall now illustrate these results by three cases from normal network theory. Since our purpose is to illustrate, rather than to discover anything new, we shall pick cases that are particularly simple.

A. Lossless Transmission Line

The operator for a lossless transmission line of real length θ and real impedance Z is

$$L = \begin{pmatrix} \cos\theta & jZ\sin\theta \\ j(1/Z)\sin\theta & \cos\theta \end{pmatrix}. \tag{22}$$

The eigenvalues of this are exp $j\theta$ and exp $(-j\theta)$. Since these are distinct if $\theta \neq n\pi$, the diagonal form is

$$L_1 = \begin{pmatrix} \exp j\theta & 0 \\ 0 & \exp(-j\theta) \end{pmatrix}, \tag{23}$$

which is obtained with

$$T = \frac{1}{\sqrt{2}}\begin{pmatrix} Z & Z \\ 1 & -1 \end{pmatrix}. \tag{24}$$

(It will be recognized that, in this case, we have simply changed to a wave basis with impedance Z. This is not generally true, however.) In this case $L_1^{\dagger -1} = L_1$, so $P = I$, the identity matrix.

$$K_2 = \begin{pmatrix} A & 0 \\ 0 & B \end{pmatrix}. \tag{25}$$

Putting (24) and (25) into (16), we find that the most general K is

$$K = \begin{pmatrix} \beta/Z & \alpha \\ \alpha & \beta Z \end{pmatrix}, \tag{26}$$

where α and β are new constants, possibly complex.

That (26) is acceptable can be verified directly in (4). If we now write out the scalar invariances resulting from (26), we find that there are two independent ones that can be written. Setting first $\alpha = 1$ and $\beta = 0$, and then $\alpha = 0$, $\beta = 1$:

$$\begin{aligned} s_1 &= (VI^* + V^*I) \\ &= \frac{1}{2Z}\{(V + ZI)(V + ZI)^* \\ &\qquad - (V - ZI)(V - ZI)^*\} \end{aligned} \tag{27}$$

$$\begin{aligned} s_2 &= VV^*/Z + ZII^* \\ &= \frac{1}{2Z}\{(V + ZI)(V + ZI)^* \\ &\qquad + (V - ZI)(V - ZI)^*\}. \end{aligned} \tag{28}$$

The scalar, s_1, is simply the net real power flow in the direction from input to output. That it is invariant is an expression of the conservation of energy. As may be seen from the first form of (27), it is independent of Z and, hence, is common to all such lines.

The second scalar, s_2, in its first form, is somewhat unfamiliar. In its second form, however, it is the sum of the powers in the forward and backward waves, referred to the characteristic impedance. In combination with s_1, it indicates that the power flow is conserved separately in each wave.

Since s_2 includes Z explicitly, it is specialized to lines of this particular impedance. If, however, we consider lossy networks, then s_2 is common to a larger class of operators. Hence, a lossless transmission line of specified characteristic impedance can be characterized as the intersection of the classes associated with these invariants.

B. Transmission Line Below Cutoff

Let us now consider a line below cutoff so that we may see what happens in a block band. We shall, in this case, simplify the process somewhat by assuming normalization so that we can write L as

$$L = \begin{pmatrix} \cosh\gamma & j\sinh\gamma \\ -j\sinh\gamma & \cosh\gamma \end{pmatrix}, \tag{29}$$

where γ is real. This situation is a trifle more complicated since the eigenvalues are $\exp(\pm\gamma)$ which are not of unit magnitude. They are, however, (conjugate) reciprocals, so that solutions are possible.

The diagonal form is obtained with

$$T = \frac{1}{\sqrt{2}}\begin{pmatrix} 1 & 1 \\ -j & j \end{pmatrix}, \tag{30}$$

and is

$$L_1 = \begin{pmatrix} \exp\gamma & 0 \\ 0 & \exp -\gamma \end{pmatrix}. \tag{31}$$

Then,

$$L_1^{\dagger-1} = \begin{pmatrix} \exp -\gamma & 0 \\ 0 & \exp\gamma \end{pmatrix}$$
$$= \begin{pmatrix} 0 & 1 \\ 1 & 0 \end{pmatrix} L_1 \begin{pmatrix} 0 & 1 \\ 1 & 0 \end{pmatrix}. \tag{32}$$

The most general matrix that commutes with L_1 is, again, a diagonal matrix. Hence, from (16), the most general form of K is

$$K = \begin{pmatrix} \beta & \alpha \\ \alpha & -\beta \end{pmatrix}, \tag{33}$$

and the scalar invariances are

$$s_1 = V^*I + VI^*$$
$$s_2 = VV^* - II^*. \tag{34}$$

The invariant s_1 is the same as before, and again expresses the conservation of energy. However, s_2 is unique to this type of operator. An expression analogous to (28) can be written also.

C. Lumped Shunt Reactance

As a final example we will consider a shunt element whose admittance is pure imaginary jb:

$$L = \begin{pmatrix} 1 & 0 \\ jb & 1 \end{pmatrix}, \tag{35}$$

which has only a single eigenvector and a chain of length 2. (See Appendix.) The only eigenvalue is unity. The chain is

$$x_1 = \begin{pmatrix} 0 \\ 1 \end{pmatrix} \qquad x_{1,1} = \begin{pmatrix} -j/b \\ 0 \end{pmatrix}. \tag{36}$$

Hence,

$$T = \begin{pmatrix} 0 & -j/b \\ 1 & 0 \end{pmatrix}, \tag{37}$$

and the Jordan canonical form (*cf.* Appendix)

$$L_1 = \begin{pmatrix} 1 & 1 \\ 0 & 1 \end{pmatrix}. \tag{38}$$

Then it is easily found that

$$L_1^{\dagger-1} = \begin{pmatrix} 1 & 0 \\ -1 & 1 \end{pmatrix} = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix} L_1 \begin{pmatrix} 0 & 1 \\ -1 & 0 \end{pmatrix}. \tag{39}$$

The most general matrix that commutes with L_1 is

$$K_2 = \begin{pmatrix} A & B \\ 0 & A \end{pmatrix}. \tag{40}$$

Then, from (13),

$$K = \begin{pmatrix} \beta & \alpha \\ \alpha & 0 \end{pmatrix} \tag{41}$$

and the invariants are

$$s_1 = VI^* + V^*I$$
$$s_2 = VV^*, \tag{42}$$

where, again, s_1 expresses the conservation of power. In this case, s_2 expresses, in somewhat weakened form, the constancy of the voltage across the network.

The failure of the eigenvectors to be a complete set, requiring consideration of generalized eigenvectors, with the resultant complications of the Jordan canonical form, has not been disastrous. We have been able to handle the problem in spite of this complication.

We have, then, given simple examples illustrating what may happen in the application of our abstract formalism to various operators. We are now ready to proceed with parametric circuits.

III. Parametric Networks

We shall consider circuits built by the ladder connection of elements formed by the parametric variation of normal reciprocal elements.

We shall assume that all signals at the frequencies of interest do not themselves affect the parameters. Under the usual circumstances then, we are assuming that these signals are small compared to the pump, although we may think of other reasons for this decoupling.

We shall also assume that the parametric variation is expressible as a Fourier series with fundamental at a given frequency, ω, the pump frequency. The harmonics of ω may either be in the pump itself, or generated by a nonlinear effect in the circuit.

If we assume a signal with components at $\omega_n = \omega_0 + n\omega$, then the differential equations for various simple elements may be solved. If we define the state vector as the column matrix of, first, the coefficients V_n of the components of voltage with time dependence $\exp(j\omega_n t)$, and then as the column matrix of the coefficients of the current components

$$x = \begin{pmatrix} V \\ I \end{pmatrix} = \begin{pmatrix} (V_n) \\ (I_n) \end{pmatrix}. \tag{43}$$

TABLE I

OPERATORS FOR PARAMETRIC ELEMENTS

Parametric Expansion		L_{ij} {i designates the row; j designates the column}
Shunt Capacity $C(t) = c_0 + \sum (c_m \exp jm\omega t + c_{-m} \exp -jm\omega t)$	$c_{-m} = c_m^*$	$\boldsymbol{A} = \boldsymbol{D} = \boldsymbol{I}$; $\boldsymbol{B} = (0)$; $C_{ij} = \omega_i c_{i-j}$
Shunt Inductance $\frac{1}{L(t)} = \frac{1}{l_0} + \sum\left(\frac{1}{l_m} \exp jm\omega t + \frac{1}{l_{-m}} \exp -jm\omega t\right)$	$l_{-m} = l_m^*$	$\boldsymbol{A} = \boldsymbol{D} = \boldsymbol{I}$; $\boldsymbol{B} = (0)$; $C_{ij} = -1/\omega_j l_{i-j}$
Series Inductance $L(t) = l_0' + \sum (l_m' \exp jm\omega t + l_{-m}' \exp -jm\omega t)$	$l_{-m}' = l_m'^*$	$\boldsymbol{A} = \boldsymbol{D} = \boldsymbol{I}$; $B_{ij} = \omega_i l_{i-j}'$; $\boldsymbol{C} = (0)$
Series Capacity $\frac{1}{C(t)} = \frac{1}{c_0'} + \sum\left(\frac{1}{c_m'} \exp jm\omega t + \frac{1}{c_{-m}'} \exp -jm\omega t\right)$	$c_{-m}' = c_m^*$	$\boldsymbol{A} = \boldsymbol{D} = \boldsymbol{I}$; $B_{ij} = -1/\omega_j c_{i-j}'$; $\boldsymbol{C} = (0)$
Series Resistance $R(t) = r_0 + \sum (r_m \exp jm\omega t + r_{-m} \exp -jm\omega t)$	$r_{-m} = r_m^*$	$\boldsymbol{A} = \boldsymbol{D} = \boldsymbol{I}$; $B_{ij} = -jr_{i-j}$; $\boldsymbol{C} = (0)$
Shunt Conductance $G(t) = g_0 + \sum (g_m \exp jm\omega t + g_{-m} \exp -jm\omega t)$	$g_{-m} = g_m^*$	$\boldsymbol{A} = \boldsymbol{D} = \boldsymbol{I}$; $\boldsymbol{B} = (0)$; $C_{ij} = -jg_{i-j}$

then it is convenient to write the operator

$$\boldsymbol{L} = \begin{pmatrix} \boldsymbol{A} & j\boldsymbol{B} \\ j\boldsymbol{C} & \boldsymbol{D} \end{pmatrix}, \tag{44}$$

where $\boldsymbol{A}$, $\boldsymbol{B}$, $\boldsymbol{C}$, and $\boldsymbol{D}$ are themselves square matrices.

It should be remembered in this that, as in normal network theory, we are actually writing only half of the complete set of coefficients. With V_n, for example, which is the coefficient at ω_n, there is associated a V_n^* at $-\omega_n$. This is not, however, V_{-n} which is the coefficient at $\omega_{-n} = \omega_0 - n\omega$, not at $(-\omega_n)$.

We should also note that some of the frequencies of interest may have $\omega_n = \omega_0 + n\omega$ negative. For these frequencies then, V_n and I_n are the conjugates of the coefficients normally used in conventional theory. Hence, it is important that we keep in mind, at all times, the possibility of ω_n having a negative sign.

On the basis of (43), and using the form of (44), Table I gives the operators for various elements with an assumed form of Fourier expansion of the parametric dependence.

It is now our concern to find the $\boldsymbol{K}$ operators which apply to these operators. While in principle this is possible for any finite number of frequencies by the methods we have developed, in practice this is not feasible. Nevertheless, we can solve the problem for, *e.g.*, two frequencies. We can then test a likely looking generalization of the two-frequency case against $\boldsymbol{L}$ directly in (2).

If we do this, our attention is drawn to operators of the form

$$\boldsymbol{K} = \begin{pmatrix} 0 & \boldsymbol{\kappa} \\ \boldsymbol{\kappa}^\dagger & 0 \end{pmatrix}, \tag{45}$$

where $\boldsymbol{\kappa}$ is itself a matrix.

One form of $\boldsymbol{\kappa}$ that has considerable generality is

$$\boldsymbol{\kappa} = \begin{pmatrix} \ddots & & \cdots & \\ & 1/\omega_{n-1} & 0 & 0 \\ \vdots & 0 & 1/\omega_n & 0 \\ & 0 & 0 & 1/\omega_{n+1} \end{pmatrix} \tag{46}$$

$$= (\delta_{ij}/\omega_i) = (\delta_{ij}/\omega_j),$$

where δ_{ij} is the "Kronnecker delta" and equals zero if $i \neq j$, and one if $i = j$.

We may show that this $\boldsymbol{K}$ applies to the first four operators of Table I, *i.e.*, to those elements that are derived from normal elements which are reactive. For substituting (44) and (45) in (4), this requires that

$$A^\dagger \kappa C = C^\dagger \kappa^\dagger A, \quad \text{(Hermitian)} \tag{47}$$

$$B^\dagger \kappa D = D^\dagger \kappa^\dagger B, \quad \text{(Hermitian)} \tag{48}$$

$$A^\dagger \kappa D + C^\dagger \kappa B = \kappa. \tag{49}$$

The relations (47) and (48) are analogous to the requirement on the $ABCD$ representation of a normal network that, if it is purely reactive, A, B, C, and D are purely real; and (49) is analogous to the requirement that the determinant of the $ABCD$ matrix be unity.

Now, for the elements of Table I, $\boldsymbol{A}=\boldsymbol{D}=\boldsymbol{I}$ and either $\boldsymbol{B}$ or $\boldsymbol{C}=(0)$. Hence, (49) and either (47) or (48) automatically apply. The κ of (46) may be found to apply to the nontrivial one of (47) and (48) if the element is a series or shunt inductance or capacity. For example, if it is a shunt capacity, then (47) requires that

$$\sum_j (\delta_{ij}/\omega_j)\omega_j c_{k-j} = \sum_j \omega_j c_{i-j}^*(\delta_{jk}/\omega_j),$$
$$= c_{k-i} \qquad = c_{i-k}^* \tag{50}$$

and this is true.

The corresponding scalar invariant is

$$\begin{aligned} s(K) &= (V^\dagger I^\dagger)\begin{pmatrix} 0 & \kappa \\ \kappa^\dagger & 0 \end{pmatrix}\begin{pmatrix} V \\ I \end{pmatrix} \\ &= V^\dagger \kappa^\dagger I + I^\dagger \kappa V \\ &= \sum_n (V_n^* I_n + V_n I_n^*)/\omega_n, \end{aligned} \tag{51}$$

which will be recognized as the Manley-Rowe power relation.

We observe that, since $\boldsymbol{K}$ is an invariant that is common to all series and shunt capacities and inductances, it is common to all ladder connections of them. Explicitly, if $\boldsymbol{K}$ is such as to satisfy (4) for both $\boldsymbol{L}$ and $\boldsymbol{M}$ then for the product $\boldsymbol{LM}$

$$\begin{aligned} (LM)^\dagger K(LM) &= M^\dagger(L^\dagger KL)M \\ &= M^\dagger KM \quad = K. \end{aligned} \tag{52}$$

Hence, $s(\boldsymbol{K})$ is also applicable to all such "reactive" networks.

We may also readily find that, for shunt capacities and series inductances with the same shape of parametric variation (ratios of Fourier coefficients the same), we may use $\boldsymbol{K}$ as in (41), but with

$$\kappa_{ij} = c_{i-j}^*/(c_0\omega_j) = l_{i-j}'/(l_0'\omega_j). \tag{53}$$

For similarly shaped series capacities and shunt inductances,

$$\kappa_{ij} = l_0/(\omega_i l_{i-j}^*) = c_0'/(\omega_i c_{i-j}'^*). \tag{54}$$

Both of these lead to scalar invariances involving both the real power flow and "cross power" terms, *i.e.*, terms involving $V_0^* I_{-1}$, etc. These are, however, useful since they apply to reasonably large classes of elementary units, and hence to ladder networks formed from these classes.

If we prefer, the diagonal terms of these can be removed by subtracting (46) from (53) and (54). Then the scalar invariances will involve only the cross power terms.

For the series and shunt resistances, we can still find $\boldsymbol{K}$'s in the form of (45). The simplest and most general of these is with $\kappa = jI$. In this case, as Manley and Rowe have pointed out, the scalar invariant is

$$s(K) = -j\sum_n (V_n^* I_n - V_n I_n^*), \tag{55}$$

and involves only reactive power flow terms.

Two Frequency Shunt Elements

We shall further explore the above by considering a particular case of some practical interest—that of a shunt element at ω_{-1} and ω_0 only, which is both reactive and lossy. This is important because available microwave parametric elements are far from being the idealized parametric capacities that we have considered so far, and to which the Manley-Rowe theorem, in its original form, is limited. Available elements do, in fact, contain loss. The equivalent circuit apparently includes a constant resistor in shunt to the capacity, a second constant resistance in series with both, and possibly other terms as well.[6] Since this circuit is not derived by parametric modulation of a reactive network, the invariance of (51), the Manley-Rowe invariance, does not apply. It is pertinent to ask what is the comparable invariance of such a unit.

As the basic operator, we take

$$L = \begin{pmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ ja & jb & 1 & 0 \\ jc & jd & 0 & 1 \end{pmatrix} = \begin{pmatrix} I & 0 \\ jC & I \end{pmatrix}, \tag{56}$$

where we impose no restrictions on the quantities a, b, c, d, or on the form of $\boldsymbol{C}$.

The eigenvalues of $\boldsymbol{L}$ are all unity. It has two eigenvectors, $\boldsymbol{e}_1$ and $\boldsymbol{e}_2$, and two generalized eigenvectors of rank 2, $\boldsymbol{e}_3$ and $\boldsymbol{e}_4$. The transformation matrix can, therefore, be written as

$$\begin{aligned} T &= (e_1\, e_3\, e_2\, e_4) \\ &= \begin{pmatrix} 0 & -jd/(ad-bc) & 0 & jb/(ad-bc) \\ 0 & jc/(ad-bc) & 0 & -ja/(ad-bc) \\ 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{pmatrix}, \end{aligned} \tag{57}$$

and the Jordan canonical form is

$$L_1 = \begin{pmatrix} 1 & 1 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 1 \\ 0 & 0 & 0 & 1 \end{pmatrix}. \tag{58}$$

Then,

$$P = \begin{vmatrix} 0 & 1 & 0 & 0 \\ -1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 \\ 0 & 0 & -1 & 0 \end{vmatrix}. \qquad (59)$$

The most general form of K_2 which commutes with L_1 is

$$K_2 = \begin{vmatrix} \alpha_1 & \beta_1 & \alpha_2 & \beta_2 \\ 0 & \alpha_1 & 0 & \alpha_2 \\ \alpha_3 & \beta_3 & \alpha_4 & \beta_4 \\ 0 & \alpha_3 & 0 & \alpha_4 \end{vmatrix}. \qquad (60)$$

Putting (57), (59), and (60) into (16), we obtain the most general form of $\boldsymbol{K}$. The resulting expression is rather involved. We find that it is of the form

$$K = \begin{pmatrix} A & B \\ C & 0 \end{pmatrix}, \qquad (61)$$

where $\boldsymbol{A}$, $\boldsymbol{B}$, $\boldsymbol{C}$ are 2×2 matrices.

The β terms of (60) determine $\boldsymbol{A}$, and allow it to be any 2×2 matrix. These terms, therefore, simply express the constancy of the voltages across the shunt element.

The detailed expression resulting from the α's is

$$K = -j\begin{vmatrix} 0 & 0 & (\alpha_1 a^* + \alpha_3 c^*) & (\alpha_2 a^* + \alpha_4 c^*) \\ 0 & 0 & (\alpha_1 b^* + \alpha_3 d^*) & (\alpha_2 b^* + \alpha_4 d^*) \\ (\alpha_1 a + \alpha_2 c) & (\alpha_1 b + \alpha_2 d) & 0 & 0 \\ (\alpha_3 a + \alpha_4 c) & (\alpha_3 b + \alpha_4 d) & 0 & 0 \end{vmatrix}. \qquad (62)$$

If we confine our attention to Hermitian $\boldsymbol{K}$'s, then α_1 and α_4 must be pure imaginaries, and $\alpha_3^* = -\alpha_2$.

We can now enumerate a set of invariants by setting $(\alpha_1, \alpha_2, \alpha_3, \alpha_4)$ equal to $(j, 0, 0, 0)$, $(0, 0, 0, j)$, $(0, j, j, 0)$ and $(0, 1, -1, 0)$. The resultant $\boldsymbol{K}$'s are singular, but could, if we like, be recombined into a set of nonsingular $\boldsymbol{K}$'s. This is legitimate since the derivation of the invariant form from the $\boldsymbol{K}$'s does not depend on any assumption of nonsingularity. We have, then

$$s_1 = (aV_{-1} + bV_0)I_{-1}^* + \text{c.c.}$$
$$s_2 = (cV_{-1} + dV_0)I_0^* + \text{c.c.}$$
$$s_3 = (cV_{-1} + dV_0)I_{-1}^* + (aV_{-1} + bV_0)I_0^* + \text{c.c.}$$
$$s_4 = j\{(cV_{-1} + dV_0)I_{-1}^* - (aV_{-1} + bV_0)I_0^*\} + \text{c.c.}, \qquad (63)$$

where c.c. means the complex conjugate.

In the case of a pure capacity, it is possible to combine these invariants so as to eliminate the capacity and the amplitude and phase of its variation. In so doing, the cross products, $V_{-1}I_0^*$, etc., are also eliminated. The result is the Manley-Rowe relation Since the constants do not appear in it, it is common to all devices of that type.

In the general case this is not possible. Since these invariants are four equations in the four cross products, it would be possible to derive a new invariant that eliminates the cross products, leaving only the terms in the real and reactive power flows at the two frequencies. It does not appear, however, that this form would be of sufficiently simple form in the general case to be useful. The invariants of (63) appear to be in as useful a form as is possible in the general case.

IV. Conclusions

The analysis of linear homogeneous systems in terms of the invariant quadratic forms of the operator appears to be a useful and productive procedure of considerable generality. The physical interpretation of such forms is that they describe quantities which are unchanged by the system. One such, for example, describes the net real power flowing into a port. In a reactive two-port network, this quantity is conserved by the system. This is indicated by the invariance of the appropriate quadratic form.

It appears likely that such invariances always exist in a physical situation providing the problem is properly set up so that the description of the state at the output is compatible with that at the input. (This, perhaps, is the proper definition of compatibility.) The mathematical condition on the matrix operator can be explicitly stated. Specifically, its eigenvalues must either be of unit magnitude, or else can be taken in pairs that are reciprocal conjugates of each other. If this condition is met, then appropriate conservation laws can be written.

Furthermore, if the condition holds, then there exists at least n independent conservation laws, where n is the number of degrees of freedom of the system.

As a simple example, a lossless transmission line with fixed characteristics has two such invariant forms. As one of these, we may take a form which expresses the net real power flow. As the other, we may choose a form which, when combined with the first, indicates the conservation of real power flow in an appropriately defined forward or backward wave.

These concepts can be extended to include parametric devices in which the signal, idle, and other signals of interest are small compared to the pump signal (the linearized approximation).

In the case of reactive parametric devices (devices obtained by periodic variation of the parameters of a lossless network), one of these conservation laws is the Manley-Rowe relation. This quadratic form is unique the ultimate packing factor[2] of tape has not been approached, the authors of this paper undertook a study and an experimental investigation of the feasibility of reading magnetic tape recordings with a cathode-ray beam.

Electron Beam Readout Method

The resolving power of a beam of electrons is several orders of magnitude higher than that of visible light; furthermore, an electron beam can be shifted much faster than any mechanical system. Consequently, in cases where the reading medium—rather than the storage medium or the writing medium—is the limiting factor, the replacement of the magnetic readout head with an electron beam should permit a substantial improvement in the information packing density. If a beam of electrons is passed through the magnetic field extending over a track recorded on magnetic tape, and if the electrons are permitted to impinge on a fluorescent screen, it will be possible under certain conditions to observe on the screen a pattern showing details of the magnetic field. If a narrow slit is placed in the plane of the screen, it is possible to collect the electrons passing through the slit as the pattern is moved across it. The current produced by the collected electrons will vary and produce a time-varying output signal in accordance with the variations of the magnetic field.

Because of the high current-density requirement, it would be difficult to use an electron beam for the recording of a magnetic signal on tape. However, the readout process presently presents a greater limitation than the recording process. Although it is possible to record a magnetic pattern having a wavelength smaller than the effective width of the magnetic gap in the recording head, it is not practical to read wavelengths that are about the same size or smaller than the gap length. The reason for this is, that it is possible for a point on the tape to experience one or more changes in polarity while it moves across the gap and still be magnetized by the magnetic field which prevails at the time it leaves the field of the gap. This is not possible during readout. A full wavelength of the recorded magnetic field will not produce any signal. For this reason, the development of a system, that increases the packing factor of the readout process only, would be a step forward.

Historical Background

Since J. J. Thomson's first experiments, deflection of electron beams by magnetic fields has been used in many applications, most notably in cathode-ray tubes for oscilloscopes, television, and radar. In practically all such applications, the cross section of the beam is small compared to the dimensions of the magnetic field. In very few cases have experiments been conducted with an electron beam having a cross section equal to or larger than the wavelength of the magnetic field. Comparatively large beams have been used in the following experiments.

Marton and Lachenbruch[3] of the National Bureau of Standards have performed experiments with small magnetic fields (for example, fields surrounding magnetic recording wires), a comparatively large electron beam, and a dark-field method (Schlieren method). Miller[4] observed the magnetic field surrounding a magnetic tape by shooting an electron beam across the tape. In Miller's work, the tape was outside the cathode ray tube so that only that portion of the magnetic field that penetrated the wall of the tube was accessible to the electron beam. Spivak and his co-workers[5] imaged the surface of a magnetized target with secondary electrons. Blackman and co-workers[6] used an electron beam to investigate the magnetic domains of unmagnetized single crystals of hexagonal cobalt. Mayer[7] has used an electron-mirror method to study magnetic fields such as those associated with magnetic domains, magnetic recording heads and patterns recorded on magnetic tape. The electron-mirror technique has also been used in the IBM Research Laboratories.[8]

II. Description of an Electron Beam Readout System

When a steady sinusoidal signal is recorded in the conventional manner on magnetic tape, the resulting magnetic field above the surface of the tape follows:

$$B_x = B_0\, e^{-2\pi y/\lambda} \cos \frac{2\pi x}{\lambda} \tag{1}$$

$$B_y = B_0\, e^{-2\pi y/\lambda} \sin \frac{2\pi x}{\lambda} \cdot \tag{2}$$

The field is composed of two superimposed sinusoidal magnetic fields, B_x and B_y. B_y is a field in the vertical y direction, normal to the tape surface [see Fig. 1(a)], and B_x is a field parallel to the tape surface in the horizontal x direction, the direction of the magnetic track [see

[2] The number of bits of information per unit area.

[3] L. Marton and S. H. Lachenbruch, "Electron optical mapping of electromagnetic fields," *J. Appl. Phys.*, vol. 20, pp. 1171–1182; December, 1949.

[4] R. L. Miller, "Visual monitor for magnetic tape," *J. SMPTE*, vol. 61, pp. 309–312; September, 1953.

[5] G. V. Spivak, N. G. Kanavina, I. N. Chernyshev, and I. S. Sbitnikova, "Electron-optical method for imaging objects with magnetic nonhomogeneities," *Dokl. Akad. Nauk SSSR*, vol. 92, pp. 541–543; 1953. Transl.: NSF-tr-202, AEC, Oak Ridge, Tenn., W40995; February, 1954.

[6] M. Blackman and E. Gruenbaum, "Observation of magnetic domains in electron shadow photographs," *Nature*, vol. 178, pp. 584–585; September 15, 1956. (Letter to the Editor.)

M. Blackman, G. Haigh, and N. D. Lisgarten, "A new method of observing magnetic transformations," *Nature*, vol. 179, pp. 1288–1290; June 22, 1957.

[7] L. J. Mayer, "Research to Investigate the Feasibility of Electron Mirror Microscopy in the Study of Magnetic Domains," Wright Air Dev. Center, Wright-Patterson AFB, Dayton, Ohio, Tech. Rept. 57-585, ASTIA Doc. No. AD 131084; September, 1957.

L. L. Mayer, "Electron mirror microscopy of magnetic stray fields on grain boundaries," *J. Appl. Phys.*, vol. 30, pp. 1101–1104; July, 1959.

[8] J. D. Kuehler, "A new electron mirror design," *IBM J. Res. and Dev.*, vol. 4, pp. 202–204; April, 1960.

Fig. 1(b)]. The two fields have equal maximum amplitudes, but are spatially 90° out of phase. The resulting field has a constant vector amplitude. The field vector rotates with constant angular velocity while moving in the direction of the track [see Fig. 1(c)]. The amplitude of the resulting field decreases exponentially as the distance from the tape surface increases. The smaller the wavelength, the more rapidly the field decreases with increasing distance. At a distance of $\lambda/2$ from the tape, the field has been attenuated by 27 db; at a distance of λ, the field has been attenuated by 54 db, etc.

We propose a readout system based on these field characteristics. The system is described in the following paragraphs.

The signal is recorded on tracks that run across the width of the magnetic tape—not as usual in the direction of the length of tape. This cross magnetization, which is now used in several high-density recording systems[9] is accomplished by a rotating recording head that records at high speeds across the width of the tape while the tape itself moves much more slowly (see Fig. 2). The rotating head has three or four recording gaps that are magnetized by the same signal. As soon as one gap finishes the recording of one track at the far edge of the tape, the next gap begins a new track at the near edge.

In our readout process, the tape is driven over a rotating drum or a smooth cylindrical support. The cylinder axis is perpendicular to the length of the tape. An electron beam is directed toward the tape. Its axis is perpendicular to the cylinder axis (Fig. 3). Its height is adjusted so that the lower half of the beam is intercepted by the tape. The upper half of the beam passes over the surface of the tape and is intercepted by a fluorescent screen. The electrons at the lower edge of the upper half of the beam pass through the magnetic field associated with the uppermost track on the tape. Each electron is deflected in a direction perpendicular to the beam axis and to the magnetic field vector existing at the position where the electron crosses the track [Fig. 1(d)].

The amplitude of the deflection observed on the screen depends on the amplitude of the magnetic field vector, the track width, the electron velocity, and the distance between the track and the screen. Since the field vector of a recorded steady ac signal has a constant amplitude but changes its direction with constant angular velocity along the track, a cycloid pattern is produced on the screen. For deflection amplitudes that are small compared to the wavelength, the cycloids are prolate and approach a sinusoidal form [Fig. 1(e)]. For deflection amplitudes $d_c = \lambda/2\pi$, the pattern will appear as cusped cycloids [Fig. 1(f)]. For greater deflection amplitudes the pattern shows curtate cycloids [Fig. 1(g). For very large deflection amplitudes, the curtate cycloids overlap so tightly that the region of deflection appears in a uniform gray. In our readout system, the screen is located close enough to the tracks that the cycloids are either prolate or cusped. If a slit with a height equal to the height of the pattern and a width small compared to the wavelength is placed in the plane of the screen at the lower edge of the beam, and if a collector (Faraday cage) is mounted behind the slit, the current collected will vary as the slit is moved along the cycloid pattern. The collected current is an electrical reproduction of the signal recorded on the tape as long as the pattern is sinusoidal. This is approached only for small amplitudes. If the pattern consists of prolate or cusped cycloids, the collected current will be a reproduction of the signal with some nonlinear distortion. As the cycloids be-

[9] C. P. Ginsburg, "Comprehensive description of the Ampex video tape recorder" J. SMPTE, vol. 66, pp. 177–182; April, 1957.

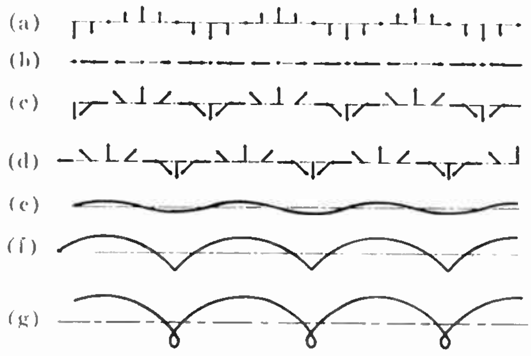


Fig. 1—Magnetic field at tape surface resulting from uniform sinusoidal recording and cycloid patterns produced by deflection of electrons by magnetic field. (a) Vertical field vector B_y. (b) Horizontal field vector B_x. (c) Resulting field vector. (d) Electron deflection due to field vector. (e) Prolate cycloid. (f) Cusped cycloid. (g) Curtate cycloid.

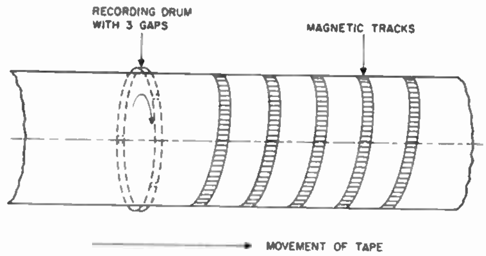


Fig. 2—Cross-tape recording with rotating recording head.

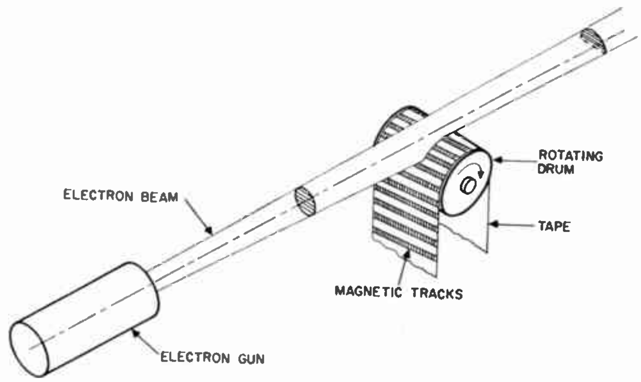


Fig. 3—Position of electron beam for readout.

come more and more curtate, not only will the distortion increase but eventually the signal amplitude will be reduced. Curtate cycloids are, therefore, to be avoided. For small wavelengths, the plane of the screen would have to be moved very close to the magnetic track to avoid curtate cycloids. In addition, for wavelengths of one mil or smaller, the width of the slit would become impractical. This problem is resolved by placing an electrostatic projection lens between the tape and the screen (Fig. 4). This permits a plane close to the track to be imaged upon a slit far away from the track.

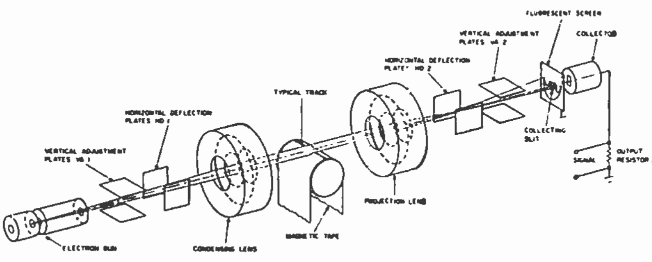


Fig. 4—Electron-optical setup.

By adjusting the voltage applied to the lens, it is possible to move the plane to a position where the pattern on the screen consists of prolate or cusped cycloids. The use of a magnification of 10 or greater makes possible the use of a slit width that is practical even for small wavelengths. By deflecting the beam with a pair of deflection plates (HD 2, Fig. 4), it is possible to move the pattern across a stationary slit.

The addition of another electrostatic lens, the condensing lens, between the electron gun and the tape increases the current density of the beam at the magnetic track. The diameter of the beam is much smaller than the length of the track—that is, the width of the tape. It is therefore necessary to have the beam scan the track; this is accomplished with deflection plates HD 1 (Fig. 4). The angle at which the beam intercepts the track must be kept very close to 90°; otherwise, electrons will pass through magnetic fields of opposite polarities, and will have deflection reversals, with a consequent reduction in signal output. The angle of incidence of the electron beam on the track must be no greater than the ratio of half the smallest recorded wavelength to the track width. To insure that this limitation will not be exceeded, the center of the deflection plates HD 1 is located in the principal focal plane of the condensing lens. The center of deflection is the plane through the center of the deflection plates; that is, the focal plane of the condensing lens parallel to the lens axis but displaced from it by a distance proportional to the deflection voltage (see Fig. 7).

The deflection plates HD 2 are located in the focal plane of the projection lens. If the deflection voltages applied to plates HD 2 and HD 1 are synchronized and adjusted in amplitude, the electron beam which enters the projection lens parallel to the tube axis will leave the deflection plates HD 2 in the tube axis; the beam therefore remains stationary on the screen while the entire length of the magnetic track is scanned.

Two pairs of vertical adjustment plates (VA 1 and VA 2, Fig. 4) are used to adjust the height of the beam. Plates VA 1 are used to optimize the beam height relative to the tape, and plates VA 2 are used to optimize the beam height relative to the collector slit.

III. Factors Determining Maximum Obtainable Information Density

The packing factor of the system described in Section II is determined by 1) the shortest readable wavelength, λ_2, and 2) the distance between the centers of adjoining tracks, s. The packing factor is

$$p = \frac{1}{\lambda_2 s} \text{ (cycles per square mil).} \tag{3}$$

The distance s between the centers of adjoining tracks is the sum of the track width w and the separation between tracks a. A theoretical estimate of the maximum packing factor, based on the assumption that the shot noise is the dominant source of noise in the collected current, shows that

$$p_{\max} = \frac{7.0}{1 + 0.92\sqrt{f_2/f_1}} \frac{B_{0_{\max}}}{S_{\max}} \sqrt{\frac{j_c}{V_0(f_2 - f_1)}} \text{ (cycles per square mil).} \tag{4}$$

An abbreviated derivation of (4) is described in the Appendix.

Note that $p_{\max}$ decreases as f_2/f_1 increases—that is, as the geometrical bandwidth of the recorded signal increases. In the case of signals with bandwidths of many octaves, it will be economical to shift the signal spectrum before recording, using carrier-modulation techniques. The signal must then be demodulated during playback to restore it to its original form. A frequency shift that "compresses" the recorded signal bandwidth into one octave will yield the maximum packing factor.

The value of the packing factor for optimum compression is given by

$$p' = 1.5 \frac{f_2}{f_2 - f_1} \frac{B_{0_{\max}}}{S_{\max}} \sqrt{\frac{j_c}{V_0(f_2 - f_1)}} \tag{5}$$

(cycles per square mil).

Table I shows the results of applying (4) and (5) in two typical cases: 1) an audio signal with a spectrum extending from 100 cps to 10 kc, and 2) a video signal with a spectrum extending from 300 cps to 3 Mc. For the audio signal, the packing factors for both direct

TABLE I
OPTIMUM DESIGN FOR TYPICAL SIGNALS

Parameter	Units	Audio Signal		Video Signal
		Direct Recording	Using Modulation Technique with $\Delta f=9.8$ kc	Using Modulation Technique with $\Delta f=3$ Mc
f_1	cps	100	100	300
f_2	cps	10,000	10,000	3,000,000
j_c	ma/cm²	350	350	350
$\sqrt{j_0/j_c}$*		1.7	1.0	2.7
V	kv	10	10	10
V_0†	volt	0.22	0.22	0.22
S_{max}	db	30	30	30
$B_{0_{max}}$	gauss	775	775	775
R	mil	31	31	31
λ_2	mil	0.055	0.02	0.14
λ_1	mil	5.5	0.04	0.28
u	inch per second	0.55	0.4	840
w	mil	0.5	0.86	2.2
s	mil	13.5	1.65	5.2
p	cycles per mil²	6.8	—	—
p'	cycles per mil²	—	14.9	0.85‡

* By demagnification between electron crossover and tape.
† For tungsten cathode at 2500°K.
‡ This value is somewhat better than the reported packing factor for conventional high-density systems.

recording and carrier modulation are given. The packing factor for a direct-recorded video signal is not given, because electronic readout of a system with a spectral bandwidth of four decades is impractical. The table shows that for audio signals with narrow bandwidths, substantial improvements can be obtained over conventional high-density systems. The value of 0.85 for the theoretical optimum packing factor for a 3-Mc video signal using frequency modulation is somewhat better than the packing factor for a conventional video recording system.

IV. EXPERIMENTS

Experimental Setup

Fig. 5 shows the experimental setup. A 6-inch-diameter 40-inch-long copper tube contains an optical bench made of a 1.5-inch-square aluminum bar (see Fig. 6). The electron-optical elements—the electron gun, the electrostatic lenses, the deflection systems, and the holder supporting the magnetic material under investigation—are mounted on riders that can be moved on the optical bench. The elements are aligned on the bench with a beam of light before the bench is installed in the vacuum chamber. To avoid cathode poisoning (which is difficult to avoid when using oxide-coated cathodes in the vacuum usually obtainable in demountable cathode-ray tubes), the electron gun uses a tungsten ribbon as a cathode. Fig. 7 shows the main electron-optical dimensions.

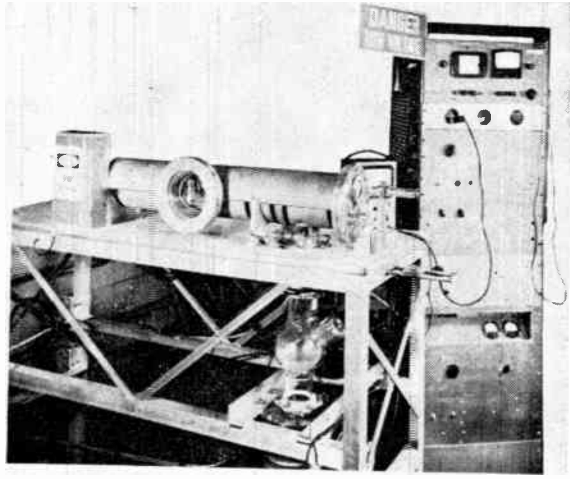


Fig. 5—Over-all view of experimental setup.

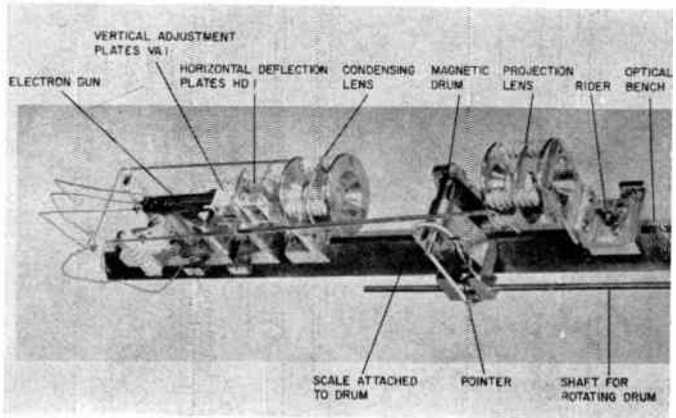


Fig. 6—Electron-optical bench.

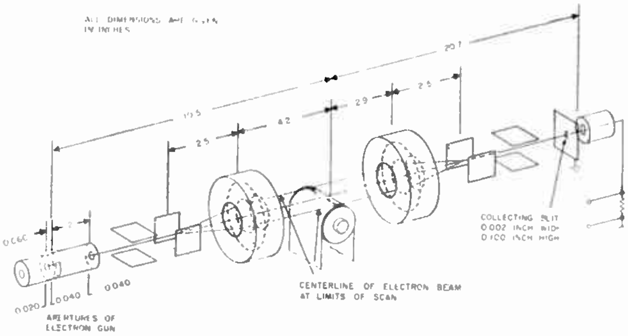


Fig. 7—Main dimensions of electron-optical system.

Most of the magnetic fields investigated were tracks that had been recorded on magnetic drums parallel to the drum axes. In some cases, magnetic tapes were studied; in those cases, short sections of a standard ¼-inch-wide magnetic tape were mounted on a drum in such a way that the track recorded on the tape was parallel to the drum axis.

The drum (Fig. 6) is rotated from the outside, by means of a shaft through a conical ground joint (Fig. 8).

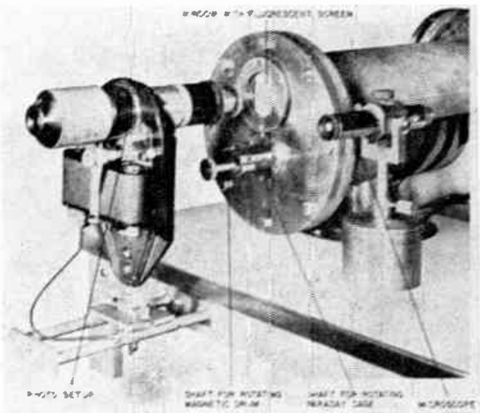


Fig. 8—Observation and recording setup.

A glass window is mounted in the end flange of the tube. and a large portion of the window is covered by a fluorescent screen. The pattern produced by the interaction of the magnetic field with the electron beam can be observed and measured on this screen with a low magnification (30×) microscope, or it can be photographed with a camera (Fig. 8). When electrical measurements are to be taken, the Faraday cage is moved in front of the window, by means of another ground joint (Fig. 9). With the aid of a mirror mounted inside the tube it is possible to observe the surface of the Faraday cage from the outside. This surface is covered with fluorescent material, making it possible to adjust the position of the beam with regard to the collecting slit.

The beam current collected by the Faraday cage is amplified and fed to the vertical deflection plates of an oscilloscope. The horizontal plates of this oscilloscope are driven from the same sawtooth voltage source that supplies the deflection voltages for scanning plates HD 1 and HD 2. All three deflection voltages can be adjusted in amplitude separately by means of potential dividers. The dc deflection biases of all three deflection systems can be adjusted independently. Three separate high-voltage dc supplies, with voltages variable from 0 to 15 kv, are used for the cathode and the two electrostatic focusing lenses. The dimensions of the unipotential lenses are shown in Fig. 10.

Preparation of Test Recordings

All test recordings were made with constant signal amplitude and constant frequency. Some of the experiments were conducted with standard $\frac{1}{8}$-inch-wide tracks recorded on $\frac{1}{4}$-inch-wide plastic-base magnetic tape. Some of the magnetic tape coatings were found to have a resistivity sufficiently low that they conducted the intercepted beam current to a ground; content touching the coating; others had too high a resistivity and were charged up by the beam. In every case, the tape base was a good insulator. To avoid charge-up, a thin film of gold was evaporated on most of the tapes.

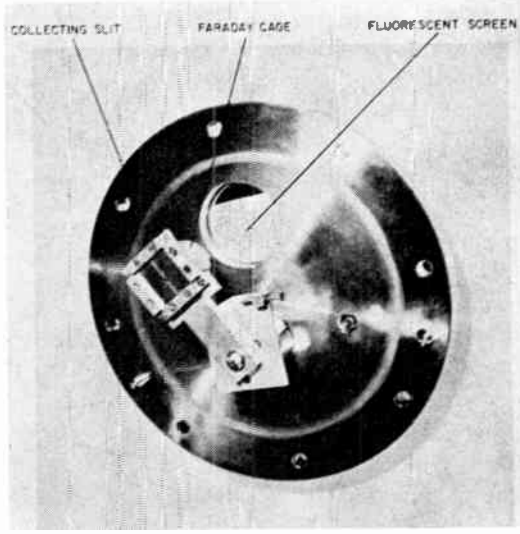


Fig. 9—Movable Faraday cage.

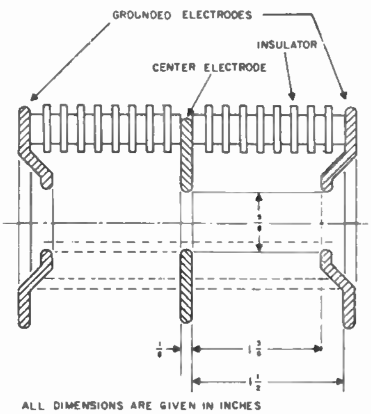


Fig. 10—Design of electrostatic unipotential lens.

At the beam voltages and intensities used during the experiments, no difficulties due to heating of the tape by the beam were encountered.

The magnetic drums used for most of the experiments were made either of brass or of nonmagnetic stainless steel. The surfaces were finished to an optical finish and then plated with a nickel-cobalt alloy. The plating process required very critical adjustments, special additives, and accurate temperature control; a combination of dc and ac currents was applied. The plating thickness varied between $\frac{1}{2}$ and $1\frac{1}{2}$ mils. The magnetic coating had very high coercivity and was difficult to demagnetize. Most of the drums were $\frac{1}{2}$ inch in diameter, but some of the experiments were conducted with drums of $\frac{1}{4}$- and $\frac{1}{8}$-inch diameter.

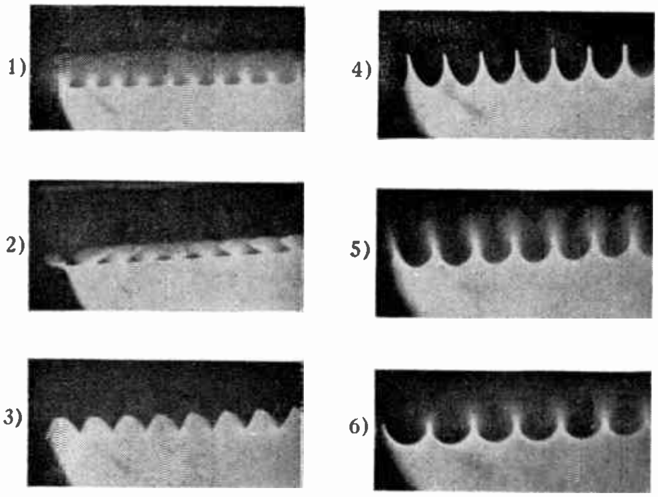
To record very narrow tracks, it was necessary to restrict the total width of the head. A single 4-mil-wide lamination of a standard recording head was used. For especially narrow tracks, the lamination's poles were honed down to a 1-mil thickness. The gap length was kept very small by inserting a piece of thin aluminum foil between the poles.

Recordings were made parallel to the axis of the drum. The drum was held in a special jig that made it possible to vary the relative speed of movement between the head and the drum. Speed and frequency were varied to record wavelengths of from 0.03 mil to 30 mils. Some recordings were made with a Brush high-resolution head (gap width: 0.050 mil).

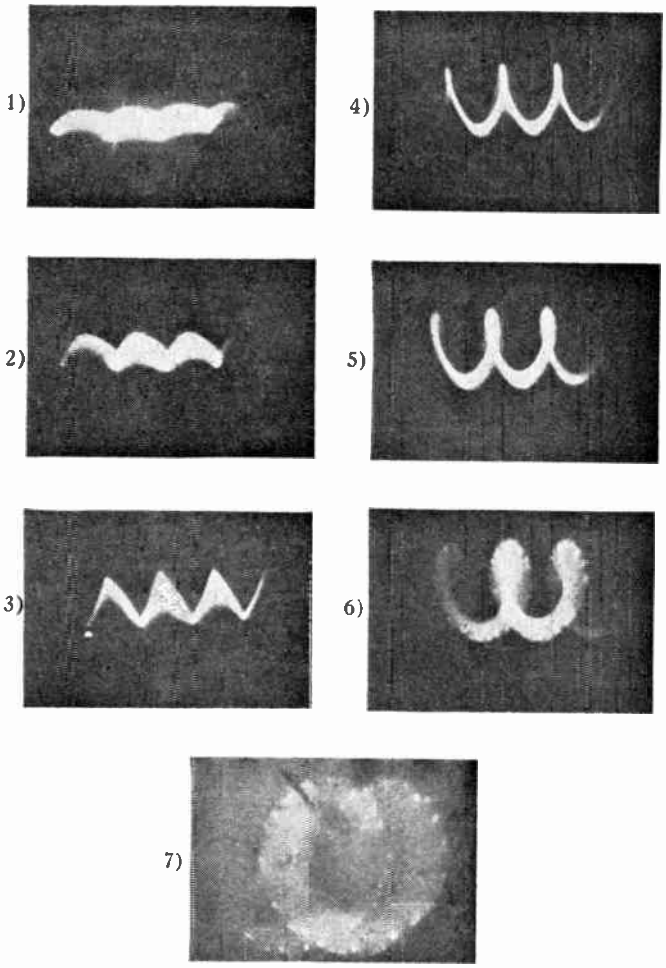
Experiments and Their Evaluation

Observation of Stationary Patterns: To observe a stationary pattern on the fluorescent screen, only the anode voltage and dc biasing voltages are applied to the deflection systems (HD 1 and VA 1) that adjust the position of the beam. As long as there is no recorded track on the top of the drum, a semicircle with a sharp edge where the drum cuts off the lower half of the beam is visible on the screen. When the drum is rotated until there is a recorded track at the top, the edge of the semicircle appears fuzzy, consisting of large curtate cycloids. When the focusing power of the projection lens is slowly increased by increasing the voltage applied to the center electrode of the lens,[10] the diameter of the semicircle on the screen will shrink until the beam crossover is imaged on the screen. With a further increase in the focusing power, the semicircle will reappear, but it will be inverted and will begin to increase in size. Again, its edge will first appear as a fuzzy region and then show a more and more distinct pattern. The pattern will show prolate cycloids when, electron-optically, the plane conjugate to the screen is located inside the magnetic track. With a further increase in the focusing power, the conjugate plane will move away from the track toward the screen, thereby increasing the observed deflection amplitude. The pattern will change from prolate to cusped cycloids, later changing into curtate cycloids and again into a fuzzy region as the focusing power is further increased. Fig. 11(a) shows the evolution of the pattern. Fig. 11(b) shows a similar evolution using a thin beam. This thin beam was obtained by pointing the beam below the top surface of the drum and showing on the screen only those electrons that were deflected over the top of the drum. Fig. 11(b)-1) shows a prolate cycloid of almost sinusoidal shape. Fig. 11(b)-7) shows the overlapping loops of the curtate cycloid. Fig. 12(a) shows cusped cycloids obtained from tracks having wavelengths of 30, 10, 3,

[10] In general, it would seem to be preferable to apply to the center electrode of the focusing lens a potential that is negative with regard to the anode, because such a voltage can be derived from the cathode voltage supply through a potential divider, and because a short focal length can be obtained with a low voltage. To obtain an equally short focal length with a positive voltage, much higher potentials (in many cases, two to three times the cathode potential) must be applied. Although positive potentials required much better insulation in the electrostatic lens, they were used almost exclusively throughout the experiments. It was found that with negative potentials, lens aberrations became more pronounced and the electrons were slowed down inside the lens to such an extent that the shielding against outside interference proved inadequate and, consequently, the resolution of the picture was deteriorated.



(a)



(b)

Fig. 11—Effect of changing focusing power of projection lens. (a) Normal-beam pattern. 1) $V_{F2}=8.4$ kv 2) $V_{F2}=9$ kv. 3) $V_{F2}=10$ kv. 4) $V_{F2}=10.7$ kv 5) $V_{F2}=11.5$ kv. 6) $V_{F2}=12.5$ kv. (Magnetic recording head: thickness, 4 mils; gap length, 0.5 mil. $\lambda=10$ mils; $V=5$ kv; $V_{F1}=5$ kv.) (b) Thin-beam pattern. 1) $V_{F2}=10$ kv. 2) $V_{F2}=10.4$ kv. 3) $V_{F2}=10.5$ kv. 4) $V_{F2}=10.8$ kv. 5) $V_{F2}=11$ kv 6) $V_{F2}=11.5$ kv. 7) $V_{F2}=12.6$ kv. (Magnetic recording head: thickness, 4 mils; gap length, 0.5 mil. $\lambda=10$ mils; $V=5$ kv; $V_{F1}=6.4$ kv.)

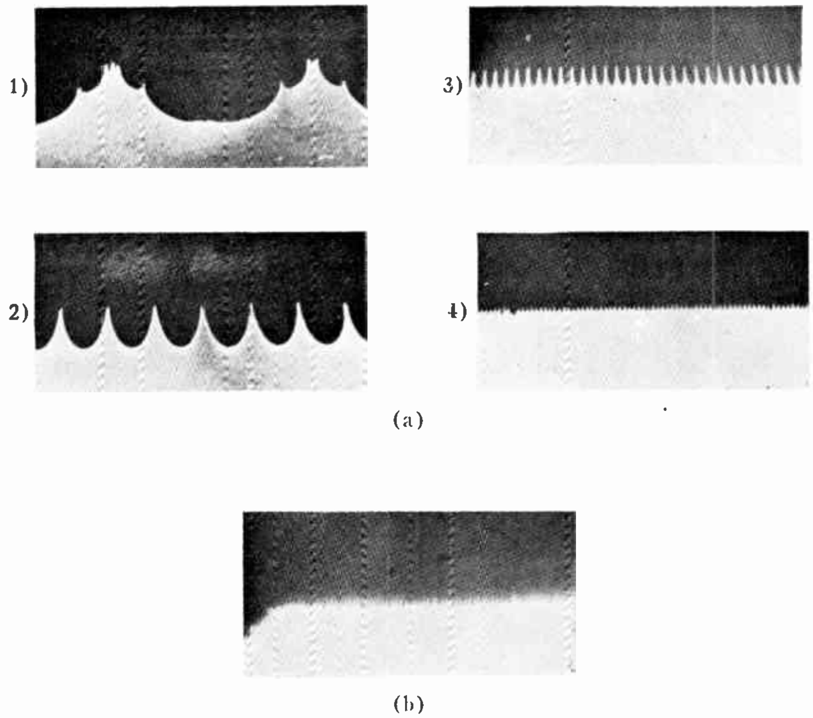


Fig. 12—Cusped cycloid patterns. (a) Obtained from magnetic drum. 1) $\lambda=30$ mils; $V=5$ kv; $V_{F1}=0$; $V_{F2}=15$ kv. 2) $\lambda=10$ mils; $V=5$ kv; $V_{F1}=0$; $V_{F2}=15$ kv. 3) $\lambda=3$ mils; $V=5$ kv; $V_{F1}=0$; $V_{F2}=11$ kv. 4) $\lambda=1$ mil; $V=5$ kv; $V_{F1}=-2.8$ kv; $V_{F2}=13$ kv. (b) Obtained from magnetic tape. $\lambda=1$ mil; $V=5$ kv; $V_{F1}=5$ kv; $V_{F2}=11.6$ kv.

and 1 mil; the tracks were recorded on a drum of $\frac{1}{8}$-inch diameter. Fig. 12(b) shows a similar pattern obtained from a plastic-based magnetic tape that was bent over a nonmagnetic drum.

Adding the condensing lens to the electron-optical setup (Fig. 4) changes the observed pattern only very little. Its main effect is to increase the beam current density j_0 at the surface of the tape.

Measurement of B_0: The height of the fuzzy region that can be observed on the fluorescent screen when no focusing voltages are applied to either the condensing or projection lenses and no ac voltages are applied to the horizontal deflection systems can be used to determine the maximum remanent magnetism ($B_{0_{max}}$) on the tape. The height d_e is measured with the calibrated microscope

$$B_{0_{max}} = \frac{4d_e\sqrt{2V/\eta}}{wD} \quad [\text{gauss}]. \tag{5}$$

The values of $B_{0_{max}}$ measured on the experimental drums varied from 19 to 350 gauss.

Output Signals: To obtain a signal output, the Faraday cage is moved in front of the fluorescent screen and deflecting voltages are applied to the horizontal deflection plates HD 1 and HD 2 of the system and to the horizontal deflection plates of the oscilloscope. The signal recorded on the magnetic tracks is then visible on the oscilloscope. Fig. 13 shows signals obtained from cusped cycloids. If the current density of the beam (j_0) in the center plane of the track is known, the peak values of the various signals can be computed as well as measured on the oscilloscope. By shining the undisturbed beam onto the collecting slit in front of the Faraday cage and measuring the beam current passing through the slit, it is possible to determine the current density in the plane of the slit. This value is multiplied by the square of the magnification between the track and the slit to determine the current density at the track. This current density j_0 will vary from experiment to experiment; it depends on the adjustment of the condensing lens. In our experiments, we measured values ranging from 1 to 15 ma/cm². The equation for computing the peak-to-peak value of a signal is:

$$E_{\text{peak-to-peak}} = \alpha r_L j_0(\lambda/\pi)^2 \sin(\pi d_s/\lambda_s) \text{ (volt)}. \tag{6}$$

We determined the signal amplitude for cusped cycloids, where $\alpha=1$. Table II shows a comparison between computed peak-to-peak signal voltages and values of the same signals measured on the face of the oscilloscope. The two methods produced results that agree fairly well.

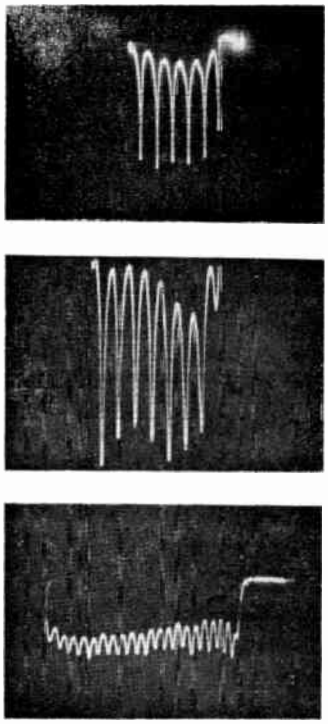
1)
2)
3)


Fig. 13—Signal patterns (magnetic drum, $\frac{1}{8}$-inch diameter). 1) $\lambda = 10$ mils. 2) $\lambda = 3$ mils. 3) $\lambda = 1$ mil

TABLE II

MEASURED AND THEORETICAL MAXIMUM SIGNAL VOLTAGES AND CURRENT DENSITIES IN PLANE OF TRACK

Radius of Drum (inch)	Recorded Wave-length (mil)	Current Density in Plane of Track (ma/cm²)	Theoretical Maximum Signal Voltage (mv)	Measured Maximum Signal Voltage (mv)
0.250	10	0.97	2.07	2.4
	3	11.3	7.1	4.6
	1	5.0	1.0	0.8
0.125	10	13.2	28.3	22.4
	3	7.0	4.42	2.3
	1	8.7	1.7	2.3
0.063	10	2.5	5.35	5.36
	3	3.76	2.4	2.5
	1	14.1	2.8	1.4

Practical Limitations on Packing Density

Limitations on Wavelength: Although in several cases tracks with wavelengths of less than 1 mil were recorded, such short wavelengths were not reproduced clearly. Random noise was observed on all tapes and drums, even where no recording had taken place. That this noise could be observed so clearly is proof that limitations in the resolving power of the system did not prevent the observation of wavelengths under 1 mil; the failure to clearly reproduce the recordings having wavelengths of less than 1 mil was caused by inhomogeneities in the magnetic materials involved. The noise was observed in stationary patterns and on low-frequency scanning patterns. It is believed that the effects of this noise will be substantially reduced in signals of limited bandwidth.

Minimum Track Center-to-Center Distance: The packing factor is also determined by the distance between the centers of two adjoining tracks, s. This distance is the sum of the track width w and the separation between tracks, a. Most of the experiments were made with tracks that were recorded with a 4-mil-wide recording head; some recordings were made with a 1-mil-wide head. These extremely narrow tracks produced satisfactory signals.

The minimum separation between tracks is determined by the maximum tolerable level of crosstalk. Consider an electron that grazes the surface of the tape while a track is at the top of the drum. When this electron crosses the two tracks on either side of the track at the top, it is deflected by the magnetic fields associated with these tracks. This is the source of crosstalk. Since the surface of the tape is curved, the electron will be further away from these tracks, the greater the separation between tracks. It has been calculated that if an electron crosses the nearer edge of the interfering track at a height $\lambda/2$, the average value of the interfering fields will be at least 40 db below the value at the surface of the tape; that is, the crosstalk ratio will be about 40 db. For this level of crosstalk, the separation between tracks must be

$$a = \sqrt{\lambda R}. \tag{7}$$

To test the validity of (7), the following experiment was performed. A scale on which the angle of rotation could be read against a fixed pointer was attached to the drum (see Fig. 6). The drum was rotated from zero signal, through maximum, and back to zero signal while the signal amplitude on the oscilloscope was observed. Fig. 14 shows a bell-shaped curve obtained from a series of such measurements. From these curves, curves showing the attenuation vs drum rotation were drawn (Fig. 15). A comparison of 40-db-point values obtained in this manner with the theoretically calculated values of minimum separations between tracks shows substantial agreement (see Table III).

The effects of artificially produced crosstalk between tracks is shown in Fig. 16. Two tracks whose wavelengths differed by 10 to 20 per cent were recorded very close together. In these pictures the superimposed frequencies can be seen clearly.

V. CONCLUSIONS

The investigations described here have shown that an electron beam can be used to advantage for measuring the magnitude of the remanent field on a recording medium. The beam can also be used to study the uniformity of magnetization and the frequency of dropouts.

Theoretical studies have indicated that a readout system such as that described in this paper can reproduce a 3-Mc-wide carrier-modulated television signal 30 db

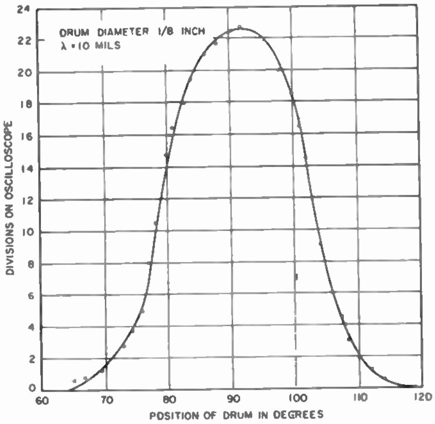
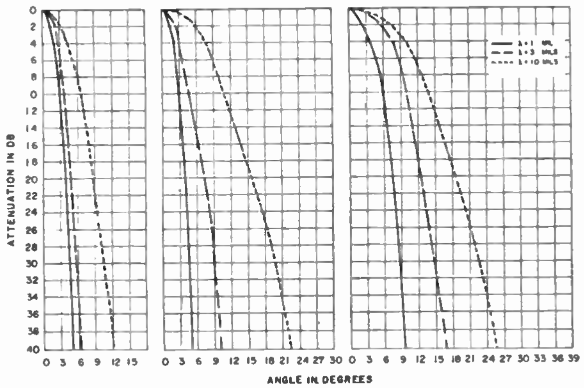


Fig. 14—Typical changes in signal output produced by moving track from zero to zero through maximum signal.



(a) (b) (c)

Fig. 15—Attenuation vs drum rotation. (a) Drum diameter: $\frac{1}{2}$ inch. (b) Drum diameter: $\frac{1}{4}$ inch. (c) Drum diameter: $\frac{1}{8}$ inch.

TABLE III

Track Center-to-Center Distance for 40-db Attenuation of Crosstalk

Radius of Drum (inch)	Recorded Wave-length (mil)	Measured Distance, s_{meas} (mil)	Theoretical Distance, s_{theor} (mil)
	10	55	55
0.250	3	28	32.5
	1	21.8	20.8
	10	48	40.4
0.125	3	19.6	24.3
	1	10.9	16.2
	10	29.2	30.0
0.063	3	17.5	18.7
	1	9.8	12.9

above shot noise from a recorded area density of 0.85 cycle per square mil. For such a density to be achieved, the recording medium must be capable of being magnetized to a maximum remanent magnetism of 775 gauss, and it must be possible to make the beam current density above the magnetic track as great as 2.5 ampere/cm^2. Both these values appear to be high, but they

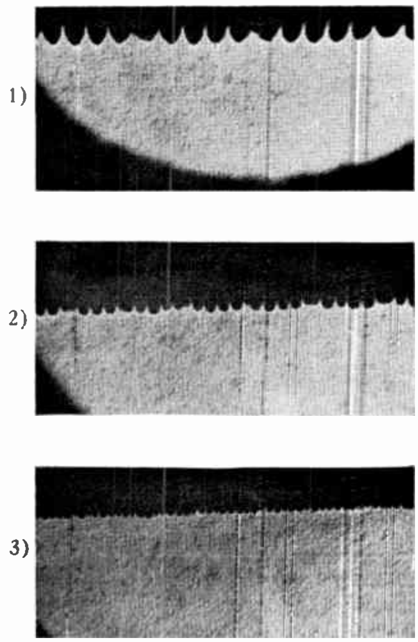


Fig. 16—Crosstalk patterns. 1) $\lambda_a=4$ mils; $\lambda_b=4.4$ mils; $s=4$ mils. 2) $\lambda_a=2$ mils; $\lambda_b=2.4$ mils; $s=4$ mils. 3) $\lambda_a=1$ mil; $\lambda_b=1.2$ mils; $s=2$ mils.

do not seem beyond reach. An area density of 0.85 cycle per square mil for a signal of 3 Mc bandwidth seems to be somewhat better than presently available area densities and comparable to area densities presently obtained in certain laboratories. For signals with bandwidths of less than 3 Mc, either much higher signal densities can be used, or the requirements upon remanent magnetism characteristics or beam-current density can be relaxed.

The experimental studies were hampered by the noise levels inherent in the recording media, both the oxide-coated tape and the nickel-cobalt-plated metallic drums. In practical applications, for example, when carrier-modulated video signals are used, a large portion of the noise will be filtered out.

The system we have described uses an electrostatic projection lens. This lens must have a usable aperture greater than the length of the track to be scanned, in this case the width of the tape. For a tape $\frac{1}{4}$-inch wide, it will not be easy to develop such a lens sufficiently free of aberrations. It will be especially difficult to develop one for a tape much wider than $\frac{1}{4}$ inch—for example, for a tape 2 inches wide.

The beam method has the great advantage over conventional mechanical systems of having a readout mechanism (the beam) that has no inertia. In mechanical readout systems, it is extremely difficult, for example, to adjust the rotational speed of the head structure to compensate for nonuniform dimensional changes occurring in the tape between recording and readout. Because of the inertia of the reading head, fast changes of speed are impossible. The electron beam is flexible. It can, for instance, be made to stop after scanning one track and to start its scanning movement when a new track arrives at the reading position. It is possible to deflect the beam so that it will be in an optimum reading position for each track, regardless of whether the track is at the exact top of the drum. This can be done with special deflection plates that bend the beam around the cylinder (see Fig. 17).

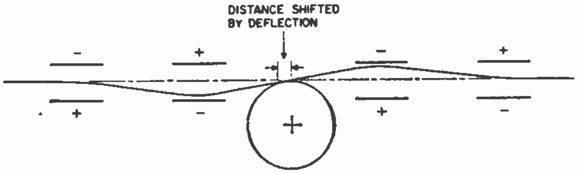


Fig. 17—System for deflecting beam around cylinder.

Putting the tape into and removing it from the vacuum chamber is not difficult to accomplish in view of recent advances in vacuum techniques and in high-speed high-vacuum pumps. It is hoped that the drying out of the tape in vacuum will not affect the life of the magnetic coating of the tape.

We would like to re-emphasize that we did not consider the recording problem in these investigations; present-day high-density recording methods are less limited than the reading methods.

We hope that the publication of this paper will promote additional interest in electron beam techniques for reproducing magnetically recorded information. We believe that the findings so far have been sufficiently promising to justify further studies of this technique.

Appendix

Theoretical Maximum Packing Factor

This Appendix outlines the derivation of the expression for the maximum packing factor given in (4). The derivation is based on two considerations:

1) The shortest wavelength (λ_2) that can be read out is determined by the permissible ratio of signal to shot noise, S_{max}.
2) The longest wavelength (λ_1) that can be read out is determined by permissible crosstalk between adjacent signal tracks.

The ratio of signal-to-shot-noise is maximized by collecting only those electrons that contribute to the signal. If this is done, the signal-to-noise ratio will be

$$S_{max} \propto \sqrt{\frac{I_0}{f_2 - f_1}} \tag{8}$$

where I_0 is the electron current collected through the collecting slit. To evaluate I_0, we note that:

1) The magnetic field extends only to a height $\lambda_2/2\pi$ above the surface of the tape and changes sign in a distance $\lambda_2/2$ along the track. Therefore, only those electrons that pass through a rectangle of area $(\lambda_2/2\pi)$ $(\lambda_2/2)$ can contribute to the signal at the high-frequency end of the spectrum of the recorded signal. Hence

$$I_0 \propto j_0\lambda_2^2 \tag{9}$$

where j_0 is the electron current density at the surface of the tape.

2) The angle between electron paths and the normal to the signal track must not exceed $\lambda_2/2D$, where D is the distance from the signal track to the collecting slit. If this angle does exceed $\lambda_2/2D$, electrons that have crossed the signal track at points further apart than $\lambda_2/2$ will enter the collecting slit simultaneously, with a resultant cancellation of signal. In other words, the divergence angle θ of the electron beam must not be greater than $\lambda_2/2D$. Langmuir's relation tells us that

$$j_0 \leq j_c\left(\frac{V}{V_0}\right)\theta^2 \propto j_c \frac{V}{V_0}\left(\frac{\lambda_2}{D}\right)^2, \tag{10}$$

where V is the voltage of the electron beam, j_c is the current density at the cathode, and V_0 is the kinetic energy of emission of electrons from the cathode expressed in electron volts.

3) The maximum angle of deflection of an electron by the magnetic field of the signal track must not exceed $\lambda_2/2\pi D$. Otherwise, the pattern of the electron beam at the screen will contain curtate cycloids with a resultant loss of signal amplitude. Mathematically, the deflection angle is

$$\frac{B_{max}w}{V} \propto \frac{\lambda_2}{D} \tag{11}$$

where w is the width of the signal track.

Upon eliminating θ and V from (8)–(11), we find

$$\frac{1}{\lambda_2 w} \propto \frac{B_{max}}{S_{max}}\sqrt{\frac{j_c}{V_0(f_2 - f_1)}}. \tag{12}$$

At the long-wavelength end of the signal spectrum, the limitation is crosstalk, and this consideration determines the minimum center-to-center separation between tracks, s. To prevent excessive crosstalk, the electrons that are being collected must not pass closer than $\lambda_1/2$ to tracks adjacent to the one being scanned. This condition is satisfied by bending the tape to a sufficiently small radius of curvature. The radius of curvature, however, must not be so small that electrons that graze the surface of the tape at the center of the scanned track cross the edge of the same track at a height greater than $\lambda_2/2\pi$; that is, electrons must pass through the full width of the magnetic field of the scanned track. To satisfy both requirements, we must have

$$\frac{s-\frac{w}{2}}{\frac{w}{2}} \propto \sqrt{\frac{\lambda_1}{\lambda_2}} = \sqrt{\frac{f_2}{f_1}}$$

or

$$\frac{s}{w} \propto 1 + \sqrt{\frac{f_2}{f_1}}. \tag{13}$$

Upon dividing (12) by (13) we find

$$p = \frac{1}{\lambda_2 s} \propto \frac{1}{1+\sqrt{\frac{f_2}{f_1}}} \frac{B_{max}}{S_{max}} \sqrt{\frac{j_c}{V_0(f_2-f_1)}}. \tag{14}$$

A more detailed analysis[11] leads to (4).

[11] M. M. Freundlich and D. I. Breitzer, "Final Report on Investigations of Techniques for Increasing Packing Factor of Magnetic Tape," Airborne Instruments Lab., Melville, N. Y., Rept. No. 3819-1; September, 1957.

ACKNOWLEDGMENT

We acknowledge with thanks the cooperation of P. A. MacLean of Airborne Instruments Laboratory for his assistance in designing and building the experimental apparatus and for his assistance during the experiments. We owe thanks to R. F. Simons of Airborne Instruments Laboratory for valuable counsel and suggestions, and to R. L. Phillips and D. R. Cross of Clevite Research Center for their help in the design of the recording equipment and in the recording.

A preliminary study for the investigation that has been discussed in this paper was initiated by Clevite Research Center and was done under contract at Airborne Instruments Laboratory in close cooperation with personnel from the Clevite Research Center. The final phase of this work was done as a cooperative effort of Clevite Research Center and Airborne Instruments Laboratory, with the Department of Defense contracting to Clevite Research Center and Clevite Research Center subcontracting to Airborne Instruments Laboratory.

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A Varactor Diode Parametric Standing-Wave Amplifier*

It has been recently pointed out by R. Landauer[1] that analysis has shown that standing wave interaction in a parametric amplifier is feasible. It is the intent of this note to present the concepts on the various types of varactor diode parametric standing-wave amplifiers originally conceived by Brett[2,3] and under investigation at this laboratory for the past 18 months.

Conceptually, the basic parametric standing-wave amplifier consists of a stable microwave generator connected to a long transmission line with a conveniently positioned detector (Fig. 1). A large amount of microwave power flows down the line from the generator and is reflected from the short, resulting in a normal voltage standing-wave pattern. The detector, a standing-wave indicator, is set close to the position of a minimum on a steep portion of the VSWR pattern. Movement of the shorting stub causes a change in the detected output, and a small amount of stub movement can account for a large detected power variation (at the same frequency). Under these conditions, the generator acts as a pump source supplying power to provide for the power gain obtained at the signal frequency.

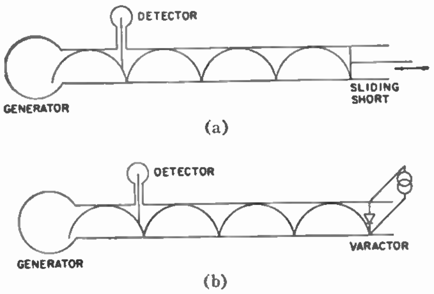

Fig. 1—(a) Basic parametric standing-wave amplifier. (b) Basic standing-wave amplifier using varactor diode.

Fig. 1(b) illustrates the basic concept of the varactor-diode parametric standing-wave amplifier. Replacement of the movable stub by a varactor diode whose reactance can be varied by an input signal gives the same results by changing the phase of the standing-wave pattern. The phase-induced power at the detector probe therefore changes at the signal frequency.

Recognizing that these essentially long line configurations were not optimum, a balanced-bridge circuit was developed (Fig. 2). Energy from an oscillator was fed into the shunt arm of a magic tee or hybrid. At the junction of the hybrid, power was divided equally between the two side arms, which are terminated by a variable position short and a varactor diode followed by an adjustable short, respectively. The series arm was terminated with a suitably matched detector.

The variable short following the varactor was adjusted such that the varactor was located at a low impedance point in the standing-wave pattern so that most of the power incident in that arm was reflected back toward the hybrid. The variable short on the second arm was adjusted so that the bridge was unbalanced slightly, allowing approximately one milliwatt to be incident on the detector. This small amount of unbalance was desirable in order to bias the detector for maximum sensitivity. Under these conditions, most of the available pump

* Received by the IRE, October 3, 1960; revised manuscript received, October 24, 1960.

[1] R. Landauer, "Parametric standing wave amplifiers," PROC. IRE, vol. 48, pp. 1328–1329; July, 1960.

[2] H. Brett, "New Modes of Parametric Amplification for Wide-Band Applications," Internal Lab. Memorandum, USASRDL, Fort Monmouth, N. J.; May 23, 1960.

[3] H. D. Webb, "Studies and Investigations Leading to the Design of a Radio Detector Finder System for the MF-HF-VHF-Range," Memorandum on Parametric Amplifier, Fourth Quarterly Rept., USASRDL Contract No. DA36-039-SC-84525; June 30, 1960.

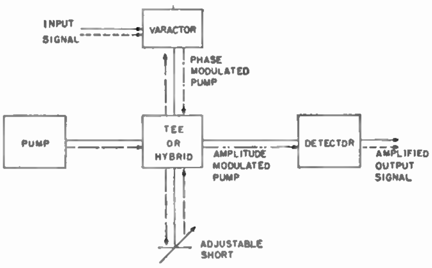

Fig. 2—Balanced-bridge standing-wave amplifier.

power reflected from the two side arms was cancelled at the junction.

When the signal was fed across the varactor diode through a suitable input circuit, the junction capacitance was modulated at the signal frequency. This resulted in a modulation of the line impedance which caused the phase of the pump power to be varied at the signal frequency. Thus, the reflected pump power from one side arm of the tee was phase modulated, while that from the other side arm was unmodulated. The hybrid functioned as an adding circuit, and the phase-modulated pump in one side arm which caused an additional unbalance in the bridge at the signal frequency was converted to an amplitude-modulated wave at the pump frequency in arm four. Clearly, if the variation in impedance caused by the input signal required less power at the signal frequency than the resultant power unbalance in the bridge, power gain would result.

The particularly interesting aspect of this kind of amplifier was that sharply tuned resonant circuits were conspicuous by their absence. In other words, this type of amplifier is, in principle, extremely broadband and as such is important for many applications. The bandwidth was limited by the properties of the input (signal) and output (detector) circuits. For proper tuning, appreciable gain (>10 db) was readily attainable, and the gain depended on pump power and the variation in diode capacitance. The conditions for optimum operation of this type of circuit were not trivial, especially in the microwave region.

With configurations similar to that illustrated in Fig. 2, low-noise (<2 db) gain has been obtained at various frequencies from dc to 3 kMc. With pump powers between 50–100 mw at 10 kMc, typical power gains from 6 to 30 db were obtained. Video gains in excess of 60 db have been observed. In all cases, low-noise performance has been achieved. Bandwidths have been disappointing however, varying from 1.0–1.5 per cent at the higher frequency ranges. The bandwidth limitations experienced thus far have been the result of tuning elements (double stub or E-H) included to compensate for imperfect matching conditions at input and output ports. More refined structures are presently under construction and results will be reported as soon as available.

The authors are indebted to their many friends in the parametric amplifier field for lively discussions, helpful criticisms and experimental verifications. Several colleagues have pointed out that these results might also be interpreted in terms of a four-frequency amplifier mode, *i.e.*, double sideband, up conversion-double sideband, down conversion. Eckhardt and Sterzer[4,5] have reported on an amplifier of this type.

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[4] W. Eckhardt and F. Sterzer, "A modulation-demodulation scheme for ultra high-speed computing and wideband amplification," *Digest of Technical Papers*, 1960 Internatl. Solid-State Circuits Conf., Philadelphia, Pa., pp. 34–35; February, 1960.

[5] W. Eckhardt and F. Sterzer, "A Four-Terminal Low-Noise Parametric Microwave Amplifier," presented at the PGMIT National Symposium, San Diego, Calif.; May 9–11, 1960.

Three-Terminal Variable Capacitance Semiconductor Device*

Consider a device of the type shown in Fig. 1, which is essentially a conventional n-p-n transistor structure with both junctions biased in the reverse direction. If the depletion regions of the two junctions just touch, a voltage variation as shown in the solid line of Fig. 2 is obtained. An increase in the reverse bias of the interaction electrode produces a voltage variation, as shown in the dashed curve of Fig. 2. This voltage variation has, at the collector, a voltage gradient which is less than that corresponding to the solid curve. Such a controlled voltage gradient has been suggested[1] for a new type of semiconductor amplifier. Associated with the change in collector voltage gradient, there is also a change in junction transition capacitance, and it is here suggested that this effect can form the basis for a three-terminal variable capacitance device. This effect will now be analyzed quantitatively.

Assume a parallel plane geometry and neglect edge effects. For a collector-to-base bias voltage V_{CB} and an interaction-to-base bias voltage V_{IB} (see Fig. 1) the voltage variation through the semiconductor body has been shown[1] to be

$$V = \frac{qn_i}{2K\epsilon_0} Dx^2 - \left(V_{CB} - V_{IB} + \frac{qn_i DW^2}{2K\epsilon_0}\right)\frac{x}{W} + V_{CB}, \quad (1)$$

provided that the two depletion regions touch or overlap. In this equation

V = the voltage measured with respect to the source electrode, with a small contact voltage usually less than one volt neglected,

* Received by the IRE, June 13, 1960; revised manuscript received, June 30, 1960. Presented at Solid-State Device Res. Conf., Carnegie Institute of Technology, Pittsburgh, Pa., June 13–15, 1960.

[1] L. J. Giacoletto, "Avalanche controlled semiconductor amplifier," Proc. IRE, vol. 47, pp. 1379–1381; August, 1959.

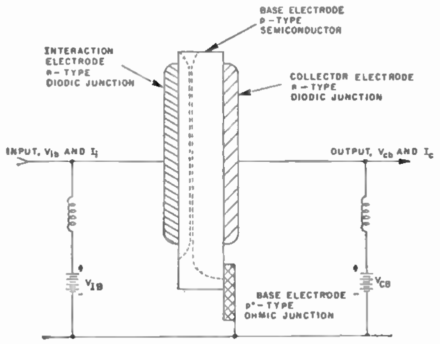

Fig. 1—Device structure and a possible circuit arrangement.

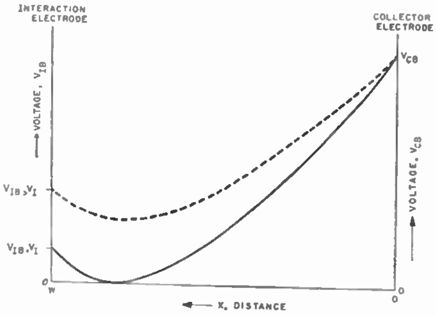

Fig. 2—Variation of voltage with distance within the semiconductor.

x = distance measured from the collector as shown in Fig. 2,

W = the base width as shown in Fig. 2,

q = the charge of a hole (1.60×10^{-19} coulombs),

ϵ_0 = the permittivity of free space ($1/36\pi \times 10^{-9}$ farads/meter),

K = the relative permittivity of the semiconductor (16 for germanium), and

D = the impurity doping of the semiconductor normalized by n_i, the electron (or hole) carrier density in the intrinsic semiconductor ($n_i = 2.4 \times 10^{11}$ carriers/meters in germanium at near room temperature, 300°K).

$$D = \frac{N - N_d}{n_i}. \quad (2)$$

From (1), the electric field at the interaction electrode, ($x = W$), E_I, and at the collector electrode, ($x = 0$), E_C, can be evaluated as

$$E_I = \frac{-qn_i D}{K\epsilon_0}\frac{W}{2} + \frac{V_{CB} - V_{IB}}{W}, \quad (3)$$

$$E_C = \frac{qn_i D}{K\epsilon_0}\frac{W}{2} + \frac{V_{CB} - V_{IB}}{W}. \quad (4)$$

Now, consider the device of Fig. 1 as a black box with an ac input per unit area interaction current, I_i, and an ac output per unit area collector current, I_o, taken as positive when flowing into the box. Since the junctions are biased in the reverse direction, the conduction currents are small and will be neglected, and the nodal currents can be written in terms of per unit area capacitive coefficients as

$$I_i = C_{ii}\frac{dV_{IB}}{dt} + C_{ic}\frac{dV_{CB}}{dt} = \frac{dQ_i}{dt}$$
$$= -K\epsilon_0\frac{dE_I}{dt}, \qquad (5)$$
$$I_c = C_{ci}\frac{dV_{IB}}{dt} + C_{cc}\frac{dV_{CB}}{dt} = \frac{dQc}{dt}$$
$$= K\epsilon_0\frac{dE_C}{dt}. \qquad (6)$$

but from (3) and (4)

$$\frac{dE_I}{dt} = \frac{dE_C}{dt} = \frac{1}{W}\left[\frac{dV_{CB}}{dt} - \frac{dV_{IB}}{dt}\right]. \qquad (7)$$

Using this equation in (5) and (6),

$$C_{ii} = C_{cc} = -C_{ic} = -C_{ci} = \frac{K\epsilon_0}{W}. \qquad (8)$$

These capacitive coefficients can be associated with an equivalent circuit, consisting of a per unit area feedthrough capacitance, $C=K\epsilon_0/W$, between interaction and collector electrodes. This simple result is directly apparent, since after all, on an ac basis, a flux increase on the interaction electrode appears as an equal flux decrease on the collector electrode.

Just before the two depletion regions interact, each electrode has capacitance to the source electrode, but no capacitance between them. As soon as the two depletion regions touch, the situation reverses and there is no ac capacitance to the source electrode, but a fixed feedthrough ac capacitance. This situation can be clarified by considering the case depicted in Fig. 3, with V_{CB} held constant at a value corresponding to depletion to $x=W/2$. Fig. 3 shows the aforementioned variation of the per unit area capacitive coefficients as a function of the interaction electrode bias voltage V_{IB}, with V_{CB} held constant at the above specified value. If the collector bias voltage is smaller than for depletion to $x=W/2$, the C_{cb} value of Fig. 3 would be larger and would remain the same to a larger value of V_{IB} before dropping to zero. Thus, by changing the interaction electrode bias voltage a small amount, a rather large discontinuity in collector junction capacitance can be obtained. The minimum capacitance as seen at the collector will depend upon the impedance of the interaction electrode bias source. If this impedance is zero, the minimum per unit area collector capacitance will be $K\epsilon_0/W$, whereas if the impedance is infinite, the minimum per unit area collector capacitance will be zero.

The device as described should have interesting possibilities in parametric type amplifier circuits and in bistable circuits. In addition, many new types of RF circuits should be possible by making use of the on-off coupling capacitance as associated with C_{ic}.

The preceding description of the device operation has been idealized for ease of presentation. In a practical device the characteristics cited will be modified by the resistance of the undepleted semiconductor; this resistance will be particularly significant just before the two depletion regions interact. The presence of this resistance will degrade the capacitances by introducing losses, will cause rounding off of the sharp corners depicted in Fig. 3, and will introduce coupling between input and output, even when C_{ic} equals zero. A practical device will also have edge effects associated with depleted regions which do not interact. Edge effects will cause C_{cb} and C_{ib} to be larger than zero even after the depletion regions interact. Current flowing in the undepleted semiconductor between the junctions will produce a "pinch-off" effect as in a field-effect transistor.

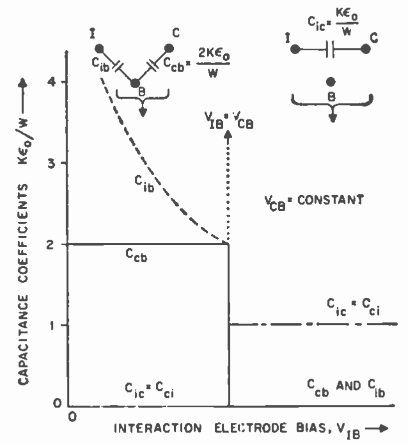

Fig. 3—Variation of capacitive coefficients with interaction electrode bias voltage.

The possibility of a three-terminal variable capacitance device was first noted by the writer several years ago while with the RCA Laboratories, Princeton, N. J. The analysis presented here is the result of discussions with D. Donald of Ford Scientific Laboratory, who analyzed the device along a somewhat different line, and J. Moll, of Stanford University.

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A High-Performance X-Band Parametric Amplifier*

A degenerate X-band parametric amplifier has been developed which realizes unusually[1,2] large gain-bandwidth products along with good noise figure performance. Voltage gain-bandwidth products as large as 10,000 Mc at high but stable gains and noise figures as low as 0.77 DSB have been obtained. The amplifier utilizes a diffused-junction, gallium-arsenide diode operated near its series resonant frequency in a coaxial circuit configuration. The self-resonant frequency of the diode is the principal resonant circuit of the amplifier. A number of matching elements were spaced along the coaxial line to provide wide-band impedance matching to the diode.

An argon gas discharge lamp was utilized as the noise source, and noise figure measurements were made using the precision IF attenuator method. The accuracy of the measurement is believed to be ±0.35 db. Fig. 1 shows some typical gain-frequency responses of the amplifier. The shape of the curves indicates that the resonant network associated with the diode junction capacitance is more complex than a simple "single-tuned" circuit.

Table I shows performance data obtained with various conditions of circuit adjustment. The center of the amplifier pass band (which corresponds to one-half pump frequency) for these measurements was 8.62 Gc. The gain given is the double-channel gain, which includes the idler output power, the single-channel gain being approximately 3 db less. The gain-bandwidth products given are based on the double-channel gain.

Noise-figure measurements were made at several spots within the amplifier pass band, and no appreciable change in noise figure was noticed. In making noise figure measurements at reduced temperatures, the entire amplifier assembly containing the GaAs diode was submerged in liquid nitrogen. While the actual junction temperature was not measured, the diode base made good thermal contact to the amplifier body and should have reached liquid nitrogen temperature.

The last column of Table I gives theoretical noise figures calculated from (1) given by Kotzebue[3] for the double sideband noise figure[4] of an optimum-degenerate amplifier in terms of the Q of the diode at operating bias and a capacitance change coefficient γ.

$$F = 1 + \frac{T_d}{T_0}\left[\frac{1}{\gamma Q} + \frac{1}{(\gamma Q)^2}\right]. \qquad (1)$$

A value of γ of 0.2 was assumed as a rough estimate based on the diode bias and

* Received by the IRE, August 22, 1960.
[1] B. C. DeLoach and W. M. Sharpless, "X-band parametric amplifier noise figures," PROC. IRE, vol. 47, p. 2115; December, 1959.
[2] R. D. Weglein and F. Keywell, "A Low-Noise X-Band Parametric Amplifier Using a Silicon Mesa Diode," presented at PGMTT Natl. Symp., San Diego, Calif.; May 9–11, 1960.

[3] K. L. Kotzebue, "Optimum noise performance of parametric amplifiers," PROC. IRE, vol. 48, pp. 1324–1325; July, 1960.
[4] It is well recognized that the double-channel noise figure of the degenerate-parametric amplifier is the appropriate noise figure for radiometer applications. Perhaps not generally appreciated, however, is the fact that the degenerate amplifier can be utilized in certain radar applications so as to realize the DSB noise figure. An excellent example is the noncoherent pulse radar using a degenerate parametric preamplifier, which is pumped at very nearly (but not necessarily precisely) twice the transmitted frequency. The explanation is that, like the radiometer case, signal information enters simultaneously both the signal and idler channels. While the output of the amplifier consists of the overlapping signal and idler spectrums, this presents no difficulty in many applications due to the post-detection time constants employed.

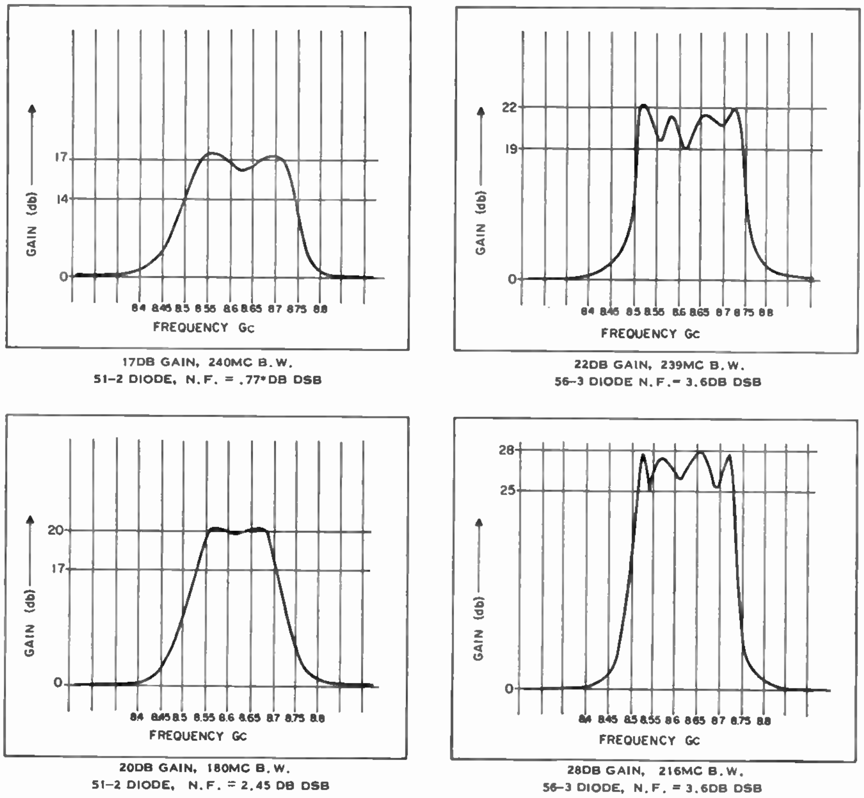

Fig. 1.

TABLE I

Diode	Gain (DSB)	Bandwidth (Mc)	$G^{1/2}B$ (DSB)	Over-All measured Noise Factor (DSB)	Computed Noise Factor (Amp. alone)	Theoretical Noise Factor Assuming $\gamma=0.2$
56-3 $f_c=129$ Gc At −2 volts Bias	15	377	2120	3.8 db	3.58 db	3.1 db
	20	260	2600	3.7 db	3.56 db	3.1 db
	22	239	3080	3.7 db	3.6 db	3.1 db
	25	231	4100	3.7 db	3.6 db	3.1 db
	28	216	5400	3.7 db	3.6 db	3.1 db
	39	117	10,300	not measured	—	3.1 db
51-2 $f_c=168$ Gc At −2 volts Bias	15	297	1676	2.65 db	2.38 db	1.6 db
	20	180	1800	2.6 db	2.45 db	1.6 db
	23	132	1870	2.3 db	2.2 db	1.6 db
	15	300	1670	1.4 db*	1.15 db	0.5 db
	17	240	1700	1.1 db*	0.77 db	0.5 db

* Cooled to liquid nitrogen temperature.

pump-voltage swing utilized. In all cases, the calculated noise figure for the amplifier alone, based on the measured values, was higher than the theoretical value. This could be due in part to poor assumptions for γ and diode Q at the operating conditions; however, it is believed that part of the difference is due to circuit loss. A low-pass filter used between the circulator and the amplifier had an insertion loss of about 0.1 db, which was included as part of the amplifier structure and not subtracted from the amplifier noise figure. In addition, it is believed that the tuning elements used to shape the band-pass curve introduced some loss.

Larger bandwidths than those given in Table I were obtained; however, it was not possible with the present circuit to restrict the ripple to 3 db. Examples of some of these adjustments were 510 Mc bandwidth with 12-db gain but gain peaks going to over 20 db; 450 Mc bandwidth with 18-db gain but gain peaks to 30 db; 680 Mc bandwidth with 9-db gain but gain peaks going to over 20 db. These measurements were made with 56-3 diode and yielded an over-all noise figure of approximately 3.7 db DSB. Such excessive gain peaks would undoubtedly be objectionable in most applications; however, they could probably be tolerated in some radiometer applications.

The performance of this amplifier should by no means be considered optimum. Further improvements in the circuit configuration should result in both wider bandwidths and lower noise figures, using the quality of diodes already available.

The author wishes to thank Dr. L. Blackwell and T. Hyltin for helpful suggestions, and is especially indebted to R. Biard, J. Eddleston, G. Pittman, and L. Wetterau of the Semiconductor Components Division, who developed and tested the GaAs diodes.

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WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should, for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table.

The characteristics of the USFS, and its relation to time scales such as ET and UT2, were described previously,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. Also they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary This new system was commenced on January 1, 1960. A retardation time adjustment of 20 msec was made on December 16, 1959; another retardation adjustment of 5 msec was made at 0000 UT on January 1, 1961.[2]

WWV Frequency
With Respect to U. S. Frequency Standard

1960 November 1600 UT	Parts in 10^{10}†
1	−148
2	−147
3	−147
4	−147
5	−147
6	−147
7	−147
8	−147
9	−148
10	−148
11	−147
12	−147
13	−147
14‡	−148
15	−152
16	−152
17	−152
18	−152
19	−152
20	−152
21	−152
22‡	−151
23	−152
24	−152
25	−152
26	−152
27	−152
28	−152
29	−152
30	−152

† A minus sign indicates that the broadcast frequency was low.

‡ The method of averaging is such that an adjustment of frequency appears on the day it is made. The frequency was decreased 4×10^{-10} on November 14, 1960, and decreased 1×10^{-10} on November 22, 1960.

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Boulder, Colo.

* Received by the IRE, December 27, 1960.

[1] "National Standards of Time and Frequency in the United States," Proc. IRE, vol. 48, pp. 105–106; January, 1960.

[2] In the preceding issue of Proc. IRE, p. 379, this date should have read January 1, 1961, instead of January 1, 1960.

Binary Error Rates in Fast Fading FDM-FM*

In several recent papers[1-3] binary error rate predictions are obtained for a frequency-division-multiplexed FM (FDM-FM) system in which the received carrier experiences Rayleigh fading sufficiently slow for the received carrier power to be considered constant over the duration of each binary waveform transmission. Though the slow fade model is often an excellent first approximation to overhorizon FDM-FM design, it is noted that in microwave tropospheric scatter communications, deep fades frequently occur having a duration of the order of 20 msec,[4] a figure which is comparable to the waveform duration used in many data systems. This note discusses the general effects on FDM-FM system error rate expected because of fast fading.[5]

We first discuss the direct distortion of intelligence due to carrier fading, *i.e.* the effects of fading in the absence of background noise. It is generally accepted that in tropospheric scatter the transmission medium effects the transmitted signal as a random time varying linear filter. We write the transmitted FM modulated carrier in its complex form, $E(t)=\exp\{i[2\pi f_ct+\theta(t)]\}$, where $\theta(t)$ represents the modulation and f_c the carrier frequency. Then the received carrier will appear in the form $V(t)=|V(t)|\exp\{i[2\pi f_ct+\theta(t)+n(t)]\}$ where $|V(t)|$ and $n(t)$ are random quantities introduced by the fading medium. $|V(t)|$ and $n(t)$ are independent of the modulation.

In the event that $V(t)$ is presented to the FM discriminator in the absence of background noise, the signal output from the FM limiter-discriminator will be proportional to the derivative of the phase exponent, $\theta'(t)+n'(t)$. Hence the direct effects of carrier fading $n'(t)$ are purely additive within the baseband, and further analysis may be carried out without regard to the actual modulation $\theta(t)$.

We assume no modulation, $[\theta(t)=0]$, and the received signal becomes, $V(t)=|V(t)|\exp\{i[2\pi f_ct+n(t)]\}$. Using a model which is often applied in explaining tropospheric scatter, it can be argued that $V(t)$ will have the properties of narrow band Gaussian noise with a bandwidth about equal to carrier fading rate.

Lawson and Uhlenbeck[6] evaluate the spectrum of an FM discriminator in response to Gaussian noise. Briefly, if the fading spectrum is confined to a narrow band near zero frequency, then so also will the "fading noise" $n'(t)$, which appears at the discriminator output. Since it is common practice in tropospheric scatter FM design to leave several empty channels (typically a band 4–12 kc wide) at the bottom of the baseband, and carrier fading rate is generally of the order of a few cycles per second, we conclude that the amount of power in the noise signal $n'(t)$ which falls within the information bearing channels will be negligible.

Even in those systems in which the direct effects of carrier fading are not negligible (the baseband intelligence extends down to near zero frequency), it is nevertheless seen that the direct effects of fading are still additive. For analytical purposes we therefore lump this "fading noise" with the rest of the background noise in the channel and conclude that unlike the more direct systems for data transmission (such as SSB), in FDM-FM the individual data signals within the baseband may be considered to maintain phase coherence during the entire transmission time of each binary pulse.

We next consider the additional effects of fading when the FM carrier is received in a background of thermal noise. We consider square-wave FSK operation with noncoherent matched-filter detection. We denote the instantaneous power of the data signal at the discriminator output by $S(t)$ and the noise power density in the vicinity of the data channel by $N(t)$. Because of the fading input both $S(t)$ and $N(t)$ will fluctuate with time. The signal component of the matched filter output will be deterministic with an energy proportional to $U_s=\{\int_0^T S(t)^{1/2}dt\}^2$. Assuming white noise input, the matched filter noise output will obey complex Gaussian statistics, and will have a mean square magnitude of $U_n=\int_0^T N(t)dt$. The probability of error is then provided according to the usual formulas for noncoherent detection,[7] $E=\frac{1}{2}\exp\{-\frac{1}{2}U_s/U_n\}$.

As a general indication of the performance to be expected when fading occurs within a transmitted binary waveform, we consider the limiting case in which fading is very rapid and the transmitted binary waveforms are very long. In this limit it is expected that U_s and U_n will each approach their long term average over fading. Denoting these averages by the bar notation $\overline{U}_s$ and $\overline{U}_n$ and substituting above, one obtains $E=\frac{1}{2}\exp\{-\frac{1}{2}\overline{U}_s/\overline{U}_n\}$.

The important thing here is that $\overline{U}_s/\overline{U}_n$ is linearly related to carrier-noise ratio C/N. Hence error rate decreases exponentially with increasing average carrier-noise ratio. This contrasts sharply with the slow fade case[8] in which the error rate for a system without diversity decreases only linearly with increasing C/N.

Even without evaluating the coefficient of proportionality for the dependence of C/N in the above formula, it should be obvious from the exponential form that for sufficiently large C/N, the fast fading system will always perform better than the corresponding slow fade system. Indeed, evaluation of this coefficient and comparison with the slow fade case indicates that for most, if not all, systems of design interest error rate is always lowest for the fast fade case.

An intuitive explanation of these results is obtained by analogy to the oft-discussed concept of time diversity. It is recognized that for data systems of interest, binary errors are rare events which occur only when received signal strength remains very small over an entire transmitted waveform. Any significant change in fading level of the signal within the transmitted waveform operates to assure that for a fraction of the pulse at least, some signal is always received. In the FM case, phase distortions introduced by the medium are not passed on to the baseband. Hence, it is possible to use pulse lengths many fade intervals long without introducing the deleterious effects of phase interference in the matched filter. This is in sharp contrast to the case of SSB, for example, where fading induced phase distortions are passed into the data channel; when long pulses are used, one accordingly finds that adjacent segments of the received signal effectively cancel. Hence in FM, fast fading operates to "chop-up" the long deep fades and assure sufficient signal strength for a proper mark-space decision with each bit transmission.

Although the results quoted apply specifically for the case of very high fading rates, it seems reasonable to expect the approach to the limit to be monotonic, and that the qualitative features of the results will remain valid even for cases of moderately fast fading. Thus we argue that the principle effect of fast fading is to produce a net decrease in system error rate. The principle results derived in the papers by DuCastel and Magnen,[1] Barrow,[2] and Johansen[3] for slow fading are thus expected to provide an important upper bound on system error rate in both the fast and moderately fast fade cases.

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* Received by the IRE, September 8, 1960. This note is a partial amplification of a study originally performed in 1958 at Hermes Electronics Co., Cambridge, Mass.

[1] F. DuCastel and J. P. Magnen, "Etude de la qualité télégraphique dans les faisceaux Hertziens transhorizon," *Annales des Télécommunications*, vol. 14, pp. 93–103; March–April, 1959.

[2] B. B. Barrow, "Error probabilities for telegraph signals transmitted on a fading FM carrier," Proc. IRE, vol. 48, pp. 1613–1629; September, 1960.

[3] D. E. Johansen, "Binary error rates in fading FDM-FM communications," submitted to the IRE Trans. on Communications Systems.

[4] A. D. Watt, *et al.*, "Performance of Some Radio Systems in the Presence of Thermal and Atmospheric Noise," Natl. Bur. of Standards, Boulder, Colo., Rept. No. 5088, Fig. 4; June 12, 1957.

[5] It is also noted that an interesting method for obtaining rough estimates of error rates in fast fading FDM-FM appears in the appendix of footnote 1.

[6] J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 24, pp. 369–374; 1950.

[7] S. Reiger, "Error probabilities of binary data transmission systems in the presence of random noise," 1953 IRE Convention Record, pt. 8, p. 72.

[8] Johansen, *op. cit.*, Fig. 2.

Capacity of Continuous Information Channel Corrected for Time or Phase Jitter*

To obtain the information capacity of a continuous communication channel, we observe that we are concerned with extracting error-free information from a signal waveform at some specified rate. If P is the signal power and N the noise power, then

$$\left(1+\frac{P}{N}\right)^{1/2}$$

* Received by the IRE, August 26, 1960. Based on work supported by ITT Communication Systems, Inc., Garden State Plaza, Paramus, N. J.

represents the number of distinguishable signal levels at a sampling point. The number of bits of information associated with this number of distinguishable levels is

$$\log_2 \left(1 + \frac{P}{N}\right)^{1/2}.$$

According to Shannon's Sampling Theorem, any signal can be reconstructed by sampling the original waveform at a rate $2W_s$, where W_s is the highest frequency component in the signal. Thus, if the amount of information associated with a given sampling point is

$$\log_2 \left(1 + \frac{P}{N}\right)^{1/2},$$

then information can be transmitted at a rate of

$$2W_s \log_2 \left(1 + \frac{P}{N}\right)^{1/2}$$

bits per unit time, where it is noted that $1/2W_s$ represents the interval of time between sampling instants. This information rate is precisely the relation obtained by Shannon for the capacity of a noisy channel.

We consider now the role played by time jitter. The Shannon relation implies unique determination of information at each sampling instant. This cannot be assumed if the waveform is jittering in time or phase. If there is jitter between sampling instants, a part of the successive sample values may represent a repetition of previous sample values and will thus constitute redundant information. Accordingly, sampling intervals should now be spaced at a greater distance apart than $1/2W_s$ in order to insure that information will not be redundant. Let us say that this additional time is Δt. Then the corrected sampling interval should be

$$\left(\frac{1}{2W_s} + \Delta t\right)$$

and we may write

$$\left(\frac{1}{2W_s} + \Delta t\right) = \frac{1}{2W_e} \qquad (1)$$

where W_e is the effective signal bandwidth, allowance being made for time jitter. It thus follows that

$$W_e = \frac{W_s}{1 + 2\Delta t W_s}. \qquad (2)$$

Hence, the corrected information capacity of the channel is

$$C_e = \frac{2W_s}{(1 + 2\Delta t W_s)} \log \left(1 + \frac{P'}{N'}\right)^{1/2}, \qquad (3)$$

noting that the original signal and noise powers have been changed to P' and N' to accord with the corrected bandwidth.

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Note on the Visualization of Impedance Transformations by Means of Three-Dimensional Plastic Sphere Models*

It has been shown in a recent note[1] and in a more complete form in another publication[2] that the Cayley-Klein representation of three-dimensional hyperbolic space is a useful tool in studying various problems in electrical engineering. However, a disadvantage with the use of this representation is the difficulty in visualizing the different geometric operations performed in three dimensions. A simple way of trying to remedy this situation is to build three-dimensional models. The author, therefore, has built a set of models of the Cayley-Klein representation of three-dimensional hyperbolic space consisting of plastic spheres pierced by differently colored straight wires. The members of the set represent studies of several technical problems. These selected problems, simple in nature, are, for example, impedance and reflection coefficient transformations through reciprocal lossy and lossless two-port networks, and cascades of two-port networks of the same type.

Fig. 1 shows the transformation of an arbitrary impedance or reflection coefficient through an ideal transformer. The ideal transformer is represented by the vertical heavy line and the two parallel horizontal heavy lines. The vertical line is the "axis" that cuts the surface of the sphere in two points which, by stereographic mapping, correspond to the fixed points in the xy impedance plane or the yz reflection coefficient plane (Smith chart). The horizontal lines,

Fig. 1—Model of the Cayley-Klein representation of three-dimensional hyperbolic space showing the transformation of an impedance or a reflection coefficient through an ideal transformer.

which are perpendicular to the axis, in the general case they are perpendicular in a non-Euclidean manner, can be selected arbitrarily as long as the non-Euclidean distance between them is kept constant. This distance is, in this case, related to the transformation ratio of the ideal transformer. An arbitrary impedance or reflection coefficient represented by the arbitrary point P on the surface of the sphere is transformed by two consecutive non-Euclidean reflections in the horizontal perpendiculars into the point P' which represents the transformed impedance or reflection coefficient[2].

It may be mentioned that the plastic spheres which consist of two hemispheres can be obtained from companies that manufacture models of the terrestrial globe. The straight lines which consist of elastic bands are covered by plastic tubes, "spaghetti," in different colors.

The author hopes that models of the Cayley-Klein representation of three-dimensional hyperbolic space will prove to be useful in electrical engineering courses. A more complete presentation of the different models built will be given in the near future.

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* Received by the IRE, September 21, 1960.

[1] E. F. Bolinder, "Radio engineering use of the Cayley-Klein model of three-dimensional hyperbolic space," PROC. IRE, vol. 46, pp. 1650–1651; September, 1958.

[2] E. F. Bolinder, "Impedance and Power Transformations by the Isometric Circle Method and Non-Euclidean Hyperbolic Geometry," Res. Lab. of Electronics, M.I.T., Cambridge, Mass., Rept. No. 312; June 14, 1957.

A Hollow-Beam Focusing System*

Previous hollow-beam focusing schemes such as Harris Flow[1] and Brillouin Flow as described by Samuel[2] ideally require the charge density in the electron beam to vary as

$$[1 + (r_a/r)^4]$$

where r_a is the inner radius of the beam. The fulfillment of this condition imposes very severe gun design problems. Practically, people have ignored the requirement, with some degradation in the focusing of the beam.

Both Harris and Samuel require a pole piece within the cathode of the hollow beam. This is structurally unattractive, and it is often difficult to keep the temperature of the ferromagnetic material below the Curie point, due to the heat radiated from the cathode. The presence of the pole piece also places restrictions on the electron gun design.

The basis of the present method is the ordinary solid-beam Brillouin focusing.[3–5] It

* Received by the IRE, August 11, 1960.

[1] L. A. Harris, "Axially symmetric electron beams and magnetic-field systems," PROC. IRE, vol. 40, pp. 700–709; June, 1952.

[2] A. L. Samuel, "On the theory of axially symmetrical electron beams in an axial magnetic field," PROC. IRE, vol. 37, pp. 1252–1258; November, 1949.

[3] L. Brillouin, "A theorem of Lamor and its importance for electrons in magnetic fields," *Phys. Rev.*, vol. 67, p. 260; April, 1945.

[4] C. C. Wang, "Electron beams in axially symmetrical electric and magnetic fields," PROC. IRE, vol. 38, pp. 135–148; February, 1950.

[5] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Co., Inc., Princeton, N. J., 2nd ed., p. 152; 1954.

is proposed that the central core of a solid beam of electrons be removed and replaced by a central cylindrical conductor which carries exactly the amount of negative charge which the electrons had carried. Thus the charge density in the remaining beam can be uniform and there is no requirement for flux to link the cathode. In other words, we can allow the innermost electrons to cross magnetic flux lines in moving from the cathode into the focusing region. The resultant inward Lorentz force on them is balanced by outward centrifugal force and an outward electric force due to the charge on the central conductor. Fig. 1 shows a sketch of a practical embodiment of the idea.

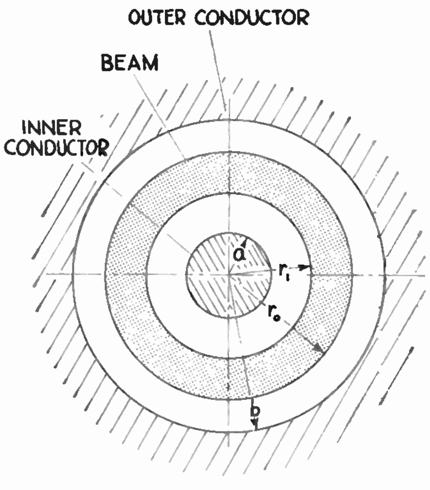

Fig. 1.

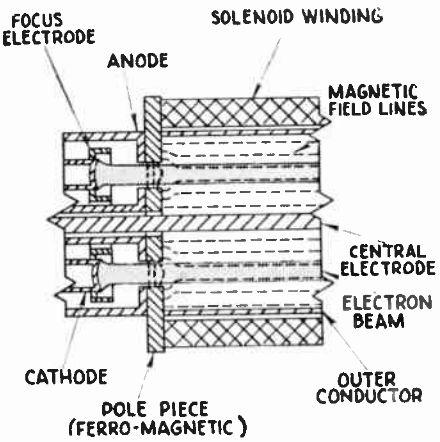

Fig. 2.

It is quite easy to derive the design equations for this system. Fig. 2 shows the nomenclature used. To find the magnetic field required, we note that the current in the equivalent solid beam is $\pi r_0^2 J_0$, where J_0 is the current density in amperes per square meter, and r_0 is the outer radius of the beam in meters. Pierce's (9.26)[5] can then be written

$$B = \sqrt{\frac{J_0\sqrt{2}}{\eta^{3/2}\epsilon V_0^{1/2}}}\ \text{webers/square meter.} \quad (1)$$

Here V_0 is the potential which corresponds to the axial velocity of the electrons. It is the potential on the axis of the equivalent solid beam.

The next requirement is to find the voltage between the center and outer conductors. We note that the charge per unit length on the inner conductor must be

$$-\pi\frac{r_i^2 J_0}{u_0}.$$

Here u_0 is the axial velocity of the electrons, and is given by

$$u_0 = \sqrt{2\eta V_0}. \quad (2)$$

Then, using Gauss's law, and integrating, we find the voltage rise between the inner electrode and outer electrode is

$$V_{ab} = \frac{J_0}{2\epsilon u_0}\left[r_i^2 \ln\frac{r_i}{a} + \frac{r_0^2 - r_i^2}{2} + r_0^2 \ln\frac{b}{r_0}\right]. \quad (3)$$

The final design equation needed is a relation between V_0 and V_b, the potential of the outer electrode relative to the cathode. By analogy with the equivalent solid beam[6]

$$V_{r_0} = V_0 + \frac{\eta B^2 r_0^2}{8}, \quad (4)$$

where V_{r_0} is the potential at the outer edge of the beam.

Adding the potential rise from radius r_0 to the outer conductor [the last term of (3)], and simplifying, we get

$$V_b = V_0 + \frac{J_0 r_0^2}{\eta^{1/2}\epsilon V_0^{1/2}}\frac{\sqrt{2}}{8}\left(1 + 2\ln\frac{b}{r_0}\right). \quad (5)$$

The problem of entrance conditions has not been considered. Generally, the requirements will be similar to those in the solid-beam case.

The focusing system has another advantage. With relatively thin beams, the focusing can be made stiffer by increasing both the axial magnetic field and the voltage between inner and outer conductors. Extra stiffness is desirable in high-power tubes, where the spreading forces due to bunching of the beam can be considerable.

The major disadvantage of this focusing system is the relatively large magnetic field required.

A note on perveance limitations appropriate to both this method of focusing and solid-beam Brillouin focusing is given in the Appendix.

Appendix

A Note of Perveance Limitations

Pierce[7] gives the following equation relating the current, the magnetic field, the beam radius and the potential at the edge of the beam for a solid, Brillouin-flow beam.

$$I = \frac{8\pi\epsilon\eta^{1/2}}{\sqrt{2}}\frac{\eta B^2 a^2}{8}\left(V_a - \frac{\eta B^2 a^2}{8}\right)^{1/2}. \quad (6)$$

[6] *Ibid.*, (9.25).
[7] *Ibid.*, (9.30).

Here a is beam radius and V_a is the potential at the edge of the beam. Since

$$V_0 = V_a - \frac{\eta B^2 a^2}{8} \quad (7)$$

(6) can be rewritten

$$I = \frac{8\pi\epsilon\eta^{1/2}}{\sqrt{2}}(V_a - V_0)V_0^{1/2}. \quad (8)$$

If V_a is fixed, and V_0 is varied, there is clearly a maximum current, which occurs when

$$V_0 = \tfrac{1}{3}V_a. \quad (9)$$

This leads to the usual statement that the maximum perveance attainable in a Brillouin focused solid beam is 25.4×10^{-6}.

If, however, we regard V_0 as fixed, there is no maximum current. If perveance is defined as $I/V_a^{3/2}$, there is a maximum perveance identical to that mentioned above, even though we regard V_0 as fixed and V_a as variable.

On the other hand, if we define perveance as $I/V_0^{3/2}$, the expression for perveance is

$$K = \frac{I}{V_0^{3/2}} = \frac{8\pi\epsilon\eta^{1/2}}{\sqrt{2}}\frac{V_a - V_0}{V_0} \quad (10)$$

and it is clear that we can raise the perveance as high as we like, if we raise V_a (and the magnetic field).

The penalty to be paid for this, of course, is that a great deal of the energy of the electrons is put into rotation around the axis. For TWT's and klystrons, (10) is a perfectly acceptable definition of beam (not gun) perveance since the axial velocity of the electrons is governed by V_0 and not V_a.

Stated another way, it is possible to pass arbitrarily large currents, at a given electron velocity, through a tube of fixed radius if we are willing to operate the tube at a high enough potential and supply enough magnetic field.

These statements become untrue when the total electron velocity (the vector sum of the axial and tangential velocities) is great enough so that relativistic effects are important.

It is easy to deduce the equivalent relations for the hollow beam described in the body of this note.

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Endfire Antennas*

In a letter published in a recent Proceedings,[1] A. F. Wickersham, Jr., described some VHF broad-band endfire antennas using tapered structures. In connection with this, we would like to mention another type of high gain endfire antenna

* Received by the IRE, September 8, 1960.
[1] A. F. Wickersham, Jr., "Recent developments in very broad-band endfire arrays," Proc. IRE, vol. 48, pp. 794–795; April, 1960.

called "Saucisson Antenna"[2] which was developed in our laboratory. It has a very interesting performance and at the same time is simple, light and cheap, especially in the VHF range.

It consists essentially of a two-wire line surrounded by a helix (Fig. 1). The two-wire line is excited by a broad-band balun fed with a 50-ohm coaxial cable. The two-wire mode is gradually changed by the action of the helix in a mode capable of radiating in the axial direction of the antenna.

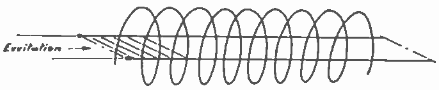

Fig. 1—Saucisson antenna for horizontal polarization.

Although the structure is uniform (with the exception of a short length of the two-wire line entering the helix) and consequently easy to build, high gain and broad-band performances are obtained. Thus, a scale model five feet long working in the S band (Fig. 2) between 2000 and 4000 Mc conserved a gain between 18 and 20 db in the whole range with secondary lobes lower than 20 db. A typical diagram of this antenna is in Fig. 3.

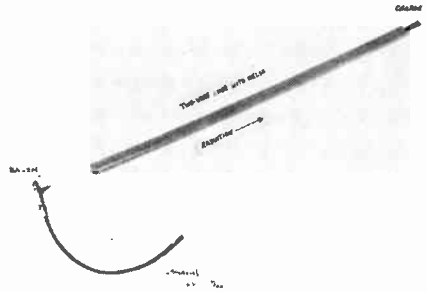

Fig. 2—Saucisson antenna.

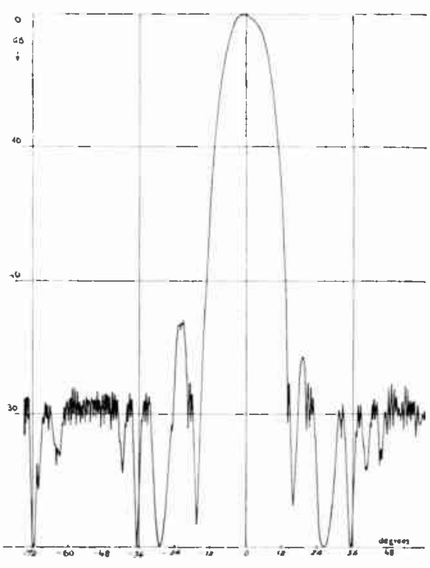

Fig. 3—Saucisson antenna.

[2] French Patent No. 1160874; November 21, 1956.

The polarization is linear and is in the plane of the two-wire line. If another line perpendicular to the first two-wire line is placed in the helix, the polarizations are perpendicular and two decoupled radiating structures are obtained. This can be very advantageous in duplex operation.

Research on this type of structure is continuing and the influence of tapering and modulation of the structure is being studied.

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Beam Interception and Limiting Gain in Adler Tubes*

Bridges and Ashkin[1] have recently described an electron beam quadrupole parametric amplifier of the Adler type[2] operating in the microwave region at 4140 Mc. In the reported experimental findings of these authors, tube saturation occurs at a pump power corresponding to approximately 20-db net gain of signal plus idler, due to beam interception in the quadrupole pump structure. As stated by the authors, amplification of random noise orbits of the electrons leaving the cathode is a possible cause of the observed current interception.

Although we agree that beam expansion due to amplification of remaining noise components takes place, it is questionable whether this effect is sufficiently large to account fully for the observed beam interception. It is the purpose of this letter to point out some beam spread phenomena that have hitherto escaped notice, from which the experimentally observed current interception can be explained on entirely different grounds. It will be shown that even an unmodulated beam of finite diameter and entirely free of noise is subject to expansion in a quadrupole pump field. This effect is of second order in the pump voltage; for this reason it is not predicted from the usual first order perturbation theory, according to which the gain in db of a fast cyclotron wave is proportional to pump voltage and the synchronous waves are left unperturbed. The individual electrons of the synchronous waves have no transverse velocities, the trajectories being parallel to the axis. Thus, the electron orbits of an unmodulated beam of finite diameter are identical to those of a pure synchronous wave. Therefore, the question of beam expansion is intimately related to the amplification of the synchronous waves. In this connection it is significant that a perturbation theory of second order in the pump voltage also reveals that the synchronous waves are amplified in the quadrupole structure and, which is more important, the synchronous waves excite cyclotron waves due to cross-coupling terms of second order in the pump voltage. Once the cyclotron waves are excited near the input region of the quadrupole structure, the further amplification is of first order in the pump voltage. The subsequent analysis of the magnitude of the effect shows excellent quantitative agreement with experimental findings.[1]

Although discussion of noise is beyond the scope of this letter, it should be pointed out that the described cross-couplings between waves introduce additional fast-wave noise components which seem to be comparable in magnitude to those considered previously.[3]

Due to the mathematical difficulties of a wave analysis beyond first order perturbations, we have investigated the beam spread phenomena in a quadrupole field using methods different from the ordinary wave approach. We have evaluated the cross-coupling between the synchronous waves and the cyclotron waves from an approximate solution of the following system of linear equations describing the electron orbits in the quadrupole field under the assumption of zero space-charge:

$$\ddot{x} + \omega_c\dot{y} = -\omega_c^2 Cx \sin(\omega_p t + \phi)$$
$$\ddot{y} - \omega_c\dot{x} = \omega_c^2 Cy \sin(\omega_p t + \phi) \qquad (1)$$

where ω_c is the cyclotron frequency, ω_p is the pump frequency, ϕ is the pump phase, and the factor C is proportional to the pump voltage V:

$$C = 2\frac{e}{m}\frac{V}{\omega_c^2 a^2}. \qquad (2)$$

$2a$ being the plate spacing in the quadrupole structure.

For $\omega_p = 2\omega_c$ and the initial conditions $x = x_0$, $y = y_0$, $\dot{x} = \dot{y} = 0$, an approximate solution of (1) can readily be found by setting $x = x_0$, $y = y_0$ in the expressions on the right-hand side of the equations. It is found that the electron orbits, which are initially parallel to the axis, start spiraling around $x = x_0$, $y = y_0$ with an initial radius approximately given by $r_1 = 2Cr/3$ where $r = \sqrt{x_0^2 + y_0^2}$ is the original radius of the circular beam. These orbits correspond exactly to the electron orbits of cyclotron waves; hence amplification takes place in the quadrupole structure. The rate at which a particular electron orbit increases its radius depends on the pump phase and the initial position of the electron relative to the quadrupole structure. For the most unfavorable combination, the instantaneous radius is given by $r_1 \exp(C\omega_c t/2)$. Parts of the beam will therefore expand to a radius R given by

$$\frac{R - r}{r} = \frac{2}{3} C \exp(C\omega_c t/2). \qquad (3)$$

Expressed in terms of the gain and the number of cyclotron periods n, the equation

* Received by the IRE, May 31, 1960; revised manuscript received, June 27, 1960.

[1] T. J. Bridges and A. Ashkin, "A microwave Adler tube," Proc. IRE, vol. 48, pp. 361–363; March, 1960.

[2] R. Adler, G. Hrbek, and G. Wade, "A low-noise electron-beam parametric amplifier," Proc. IRE, vol. 46, pp. 1756–1757; October, 1958.

[3] C. P. Lea-Wilson, "Some possible causes of noise in Adler tubes," Proc. IRE, vol. 48, pp. 255–256; February, 1960.

can be written

$$n\frac{R-r}{r}=0.035(G+3)10^{G/20}, \qquad (4)$$

where G is the power gain in db of signal plus idler under the assumption that the pump is not phase-locked to the signal.

The above expression is confirmed by an exact solution of the differential (1) on an analog computer. In Fig. 1 are shown, as examples of the results from the computer, orbits of two electrons entering the quadrupole structure simultaneously with zero transverse velocities at the positions corresponding to maximum respective minimum expansion. The curves shown in Fig. 2 bring out the close agreement between the exact results from the computer and those obtained from the approximate expression (4).

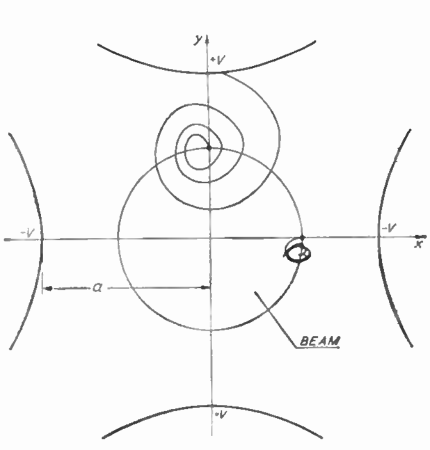
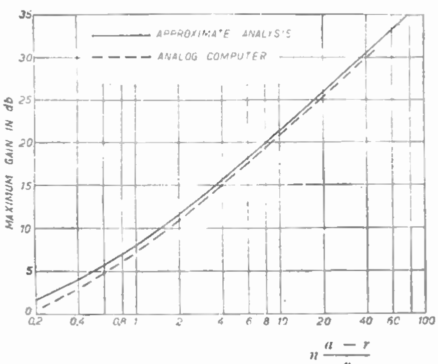


Fig. 1—Orbits of two electrons with zero initial transverse velocities. The pump voltage corresponds to $C=0.2$, or approximately 5 db gain per cyclotron period. The pump phase ϕ is zero.



$n\frac{a-r}{r}$

Fig. 2—Universal curves for beam interception in a quadrupole structure with plate spacing $2a$ and initial beam diameter $2r$. The curves show the maximum gain that can be obtained with no current intercepted as a function of the number of cyclotron revolutions

The analysis done here also leads to the obvious conclusion that current interception can be avoided or minimized without reducing the gain by using a relatively long quadrupole structure and a correspondingly smaller pump voltage.

The pump cavity used by Bridges and Ashkin is 0.125 inch long, corresponding to 5.3 cyclotron periods at 17.5 volts beam voltage. The plate spacing is 0.043 inch and the original beam diameter 0.015 inch. From the curve in Fig. 2 we find that beam interception is expected to begin at a pump voltage corresponding to a gain of 21 db, which is in excellent agreement with the experimental observations reported by these authors.

On the other hand, Adler, Hrbek, and Wade[2] have reported considerably higher gain than the saturation value predicted from the present analysis. We have no explanation for the discrepancy other than that it may possibly be due to differences in beam focusing or to stray fields near the input region of the quadrupole structure such that the initial conditions are different from those assumed in the present analysis in which we have assumed an electron beam model with electron trajectories originally parallel to the static magnetic field and no stray fields in the pump structure. For beams with initial transverse velocity components, such as Brillouin focused beams, space-charge forces play an important role for the equilibrium of the beam and thus for the details of the electron orbits in the quadrupole field. In order to obtain definite answers concerning expansion of beams based on focusing schemes in which space-charge forces are involved it appears that further theoretical and experimental investigations are required.

In conclusion we would like to suggest an experiment from which definite evidence concerning the main cause of the observed beam expansion can be obtained using a tube with a split quadrupole structure operating either with the full quadrupole length or with some fraction of the full length and a correspondingly higher pump voltage. In this way the gain and, therefore, the beam spread due to amplified noise can be kept constant, whereas the beam expansion due to the phenomena considered here will increase proportional to the pump voltage.

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Rise-Time Measurements in MgO Cold Cathode Diodes*

Recent investigations of cold electron emission from MgO and related substances have been directed both toward an understanding of the fundamental mechanisms[1,2] and toward application of this phenomenon in vacuum tubes.[3,4] The present work presents some measurements which are pertinent to both investigations. Using diodes of simple geometry, with MgO cold cathodes, current rise times upon application of a voltage step have been measured as a function of the initial current and the magnitude of the step itself.

The cathodes were prepared as previously described[1] and our typical diode configuration appears in Fig. 1. Electrons from the tungsten filament are used to start the diode, after which the filament is turned off and grounded. Anode No. 1 is the collector mesh, anode No. 2 is grounded, and the cathode metal sleeve is grounded through a low impedance logarithmic amplifier. A 1000 Ω resistor is usually placed in series with the collector mesh and the voltage source, consisting of a regulated supply and a step generator in series. Values of the resistance, between 10 and 10K Ω have little effect on the rise times.

For a diode as shown in Fig. 1, whose cathode area is approximately 1 cm², typical rise time data are shown in Fig. 2.

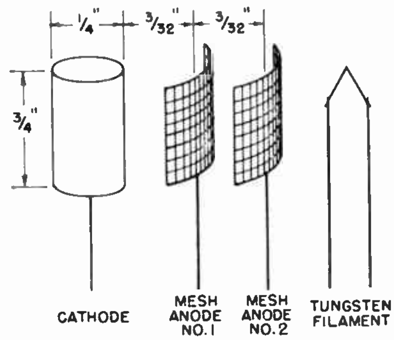

Fig. 1.

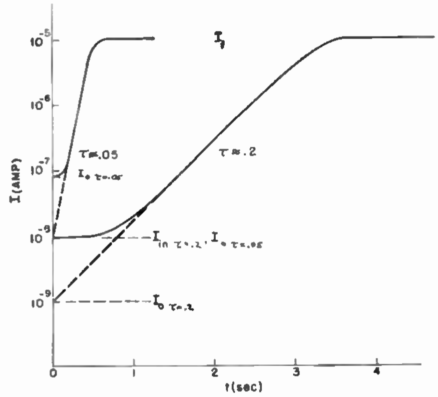

Fig. 2.

It has been found possible to fit such data with the relation[5]

$$I=I_{in}+I_0e^{t/\tau}\left(1+\frac{I_0}{I_f}e^{t/\tau}\right)^{-1}, \qquad (1)$$

where

I_0 = the extrapolated value of I at $t=0$,
I_{in} = the initial equilibrium current,

* Received by the IRE, July 27, 1960.

[1] D. Dobischek, H. Jacobs, and J. Freely, "The mechanism of self-sustained electron emission from magnesium oxide," *Phys. Rev.*, vol. 91, pp. 804–812; August, 1953.

[2] M. I. Elinson and D. V. Zernov, "The mechanism of electron emission from thin dielectric layers under the action of a strong electric field (malter effect)," *Radio tekhronika i elecktronika*, vol. 2, pp. 75–85; January, 1957.

[3] A. M. Skellett, B. G. Firth and D. W. Mayer, "The magnesium oxide cold cathode and its application in vacuum tubes," Proc. IRE, vol. 47, pp. 1704–1712; October, 1959.

[4] B. G. Firth, *et al.*, "Utilization of Self-Sustained Electron Emission Investigation," U. S. Army Signal Corps Res. Dev. Lab., Contract No. DA36–039-Sc-78081; September, 1959.

[5] Eq. (1) is derived as follows: The initial rise is exponential, of the form $I=I_0e^{t/\tau}$. Multiplying the right-hand term by a damping factor $(1-I/I_f)$ leads to (1). I_{in} is an additive constant.

I_f = the final equilibrium current,
t = time in seconds, and
τ = an empirical relaxation time.

The value of τ depends inversely on both the initial and final currents according to the relation

$$\tau = K/I_{in}I_f, \qquad (2)$$

where K is a constant equal to about 10^{-14} second A^2. The relation is valid for I_{in} between 10^{-8} A and 10^{-5} A and for I_f between 10^{-5} A and 10^{-3} A.

The rise time is thus determined by the initial steady-state current and the final current, along with geometrical factors for the thickness of the cathode layer and its area. The effect of the latter two has not yet been measured. Since the value of I is given by the relation[3]

$$I = I_1 e^{v/v_0}, \qquad (3)$$

where v is the collector potential and I_1 and v_0 are constants, and since it has been shown that the cathode surface potential is proportional to the collector potential in the ranges of current of similar experiments,[1] the rise time is thus controlled by the rate of change of the potential of the MgO cathode surface.

For fast rise times, and, therefore, high frequency response with these cold cathode diodes, the surface potential must be changed rapidly. This requires either high operating current densities and/or large collector voltage swings.

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Stripline Y-Circulators for the 100 to 400 Mc Region*

Stripline Y-circulators have been investigated by several workers.[1] All of the published work, however, is restricted to frequencies above 400 Mc. This note reports successful development efforts on such circulators in the frequency region below 400 Mc. Typical isolation and loss characteristics of Y-circulators operating at frequencies of 150, 280, and 375 Mc are presented in Fig. 1, for constant and variable applied magnetic field. A low M_s, MnMg aluminate ferrite such as Trans-Tech 414 is used in the form of circular disks with a constant thickness of 0.405 inch (per disk).

A center conductor in the form of a clover leaf (Fig. 2, III) has been chosen to obtain an optimum bandwidth for the case where a fixed magnetic field is applied. Fig. 2 gives the total bandwidth obtained with three different circulator structures, and shows cross-sectional views of the structures.

* Received by the IRE, August 15, 1960. This work was supported by Contract NObsr-77602, Electronics Div., Bureau of Ships.
[1] S. Yoshida, "Strip-line Y-circulator," PROC. IRE, vol. 48, pp. 1337–1338; July, 1960.

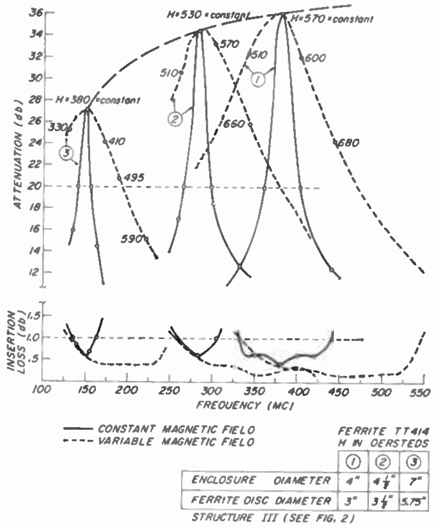

Fig. 1—Low-frequency circulator characteristics; frequency vs attenuation measured at two adjacent ports.

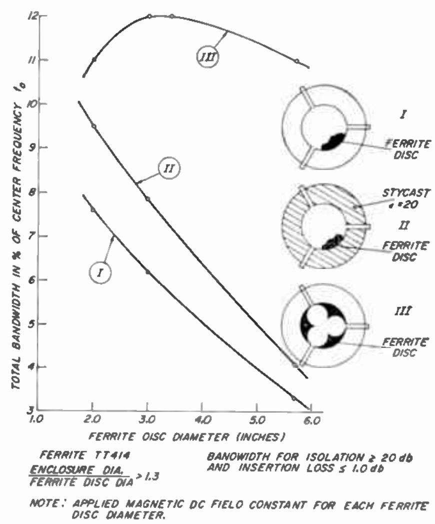

Fig. 2—Total bandwidth in per cent of center frequency f_0 for 3 different circulator structures with varying ferrite disk diameter.

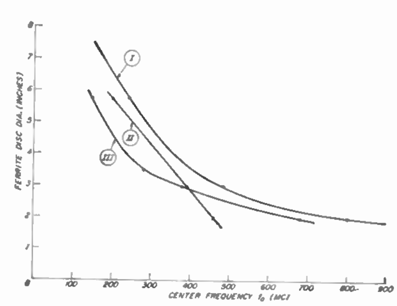

Fig. 3—Center frequency f_0 vs ferrite disk diameter for 3 circulator structures.

Fig. 3 shows the center frequency vs ferrite disk diameter relationship. The center frequency f_0 is the frequency at which maximum isolation is obtained with an applied magnetic field H_0. The loss and isolation characteristics are symmetrical, with respect to the center frequency, for the case of a constant applied field. If a variable magnetic field is used, the device operates over a larger bandwidth, which is asymmetrical with respect to the "center" frequency. Fig. 4 shows the bandwidth obtained as a function of the disk diameter for the "clover leaf" structure (Fig. 2, III). The bandwidth is given in ±per cent of the center frequency and plotted for isolations ≥20 db at a loss ≤1 db for a constant and a variable applied magnetic field. The 12-db isolation bandwidth is given for the variable field case only.

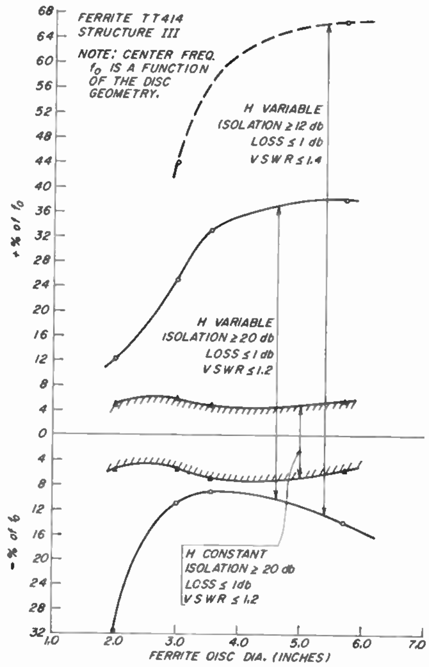

Fig. 4—Bandwidth characteristics of low-frequency circulators.

It is of importance to note that the diameter of the enclosure should not be chosen too small. The following diameter relationship should be maintained to assure optimum circulator performance:

$$\frac{\text{enclosure diameter}}{\text{ferrite disk diameter}} \geq 1.3.$$

The thickness of the ferrite disk is of some importance also, since the demagnetizing factor is determined by this dimension at a given disk diameter. The demagnetizing factor has to be chosen in such a way as to allow the applied magnetic field to saturate the ferrite disks. Saturation of the ferrite disks is necessary to operate the device at low insertion loss. The device is operated close to resonance, since nonreciprocal effects of appreciable magnitude can be observed. In summary, the following characteristics can be regarded as typical: frequency range for stripline Y-circulators 100 to 1000 Mc with maximum isolation in the order of 25 to 40 db at a minimum insertion loss of 0.6 db or less. The bandwidth for 20-db isolation at a loss of 1 db or less lies between 9 to 12 per cent for a constant and 40 to 85 per cent for a variable magnetic field. At low frequencies, the isolation can be maintained to 40 db at the expense of bandwidth. Structures I and

II (in Fig. 2) exhibit higher isolations at the center frequency; however, the bandwidth decreases rapidly with increasing ferrite disk diameter necessary for very low-frequency operation.

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Limitations of the AND-OR to Majority-Logic Conversion Technique*

Lindaman[1] described a method for the conversion of logical networks containing AND and OR functions to logical networks containing only majority decision elements. By using the conversion method, it was demonstrated that a savings in majority decision elements could be obtained. The conversion method, described by Lindaman, was restricted to networks having

1) no more than three inputs,
2) no more than three inputs per element, and
3) no storage function.

In addition to the above restrictions, the conversion method is limited to only 10 of the 256 possible switching functions of three variables.

The method is limited to those logical equations in which, after one application of the expansion theorem[2] on any one variable, *e.g.*, x, the residues of x and $\bar{x}$ must have a certain form. The form required is the x residue must be the hindrance function of the $\bar{x}$ residue, or vice versa. The hindrance function is obtained from the transmission function by interchanging the AND and OR operations in the logical equation. The 10 functions of three variables which have this property are the sum equation, the inverted sum equation and the eight permutations of the variables of symmetry of the symmetric function $S_{2.3}(X, Y, Z)$.

The 10 applicable functions are listed below:

$$\text{Sum} = XYZ + X\bar{Y}\bar{Z} + \bar{X}Y\bar{Z} + \bar{X}\bar{Y}Z \quad (1)$$

Inverted sum

$$= \bar{X}\bar{Y}\bar{Z} + \bar{X}YZ + X\bar{Y}Z + XY\bar{Z} \quad (2)$$

$$S_{2.3}(X, Y, Z) = XY\bar{Z} + XYZ + \bar{X}YZ + X\bar{Y}Z \quad (3)$$

$$S_{2,3}(\bar{X}, \bar{Y}, \bar{Z}) = \bar{X}\bar{Y}\bar{Z} + \bar{X}\bar{Y}Z + \bar{X}Y\bar{Z} + X\bar{Y}\bar{Z} \quad (4)$$

$$S_{2,3}(\bar{X}, Y, Z) = \bar{X}Y\bar{Z} + \bar{X}YZ + \bar{X}\bar{Y}Z + XYZ \quad (5)$$

$$S_{2,3}(X, \bar{Y}, \bar{Z}) = \bar{X}\bar{Y}\bar{Z} + XY\bar{Z} + X\bar{Y}\bar{Z} + X\bar{Y}Z \quad (6)$$

$$S_{2.3}(X, \bar{Y}, Z) = \bar{X}\bar{Y}Z + XYZ + X\bar{Y}Z + X\bar{Y}\bar{Z} \quad (7)$$

$$S_{2.3}(\bar{X}, Y, \bar{Z}) = \bar{X}\bar{Y}\bar{Z} + \bar{X}Y\bar{Z} + \bar{X}YZ + XY\bar{Z} \quad (8)$$

$$S_{2.3}(X, Y, \bar{Z}) = \bar{X}Y\bar{Z} + XY\bar{Z} + XYZ + X\bar{Y}\bar{Z} \quad (9)$$

$$S_{2.3}(\bar{X}, \bar{Y}, Z) = \bar{X}\bar{Y}\bar{Z} + \bar{X}\bar{Y}Z + \bar{X}YZ + X\bar{Y}Z. \quad (10)$$

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* Received by the IRE, May 9, 1960.

[1] R. Lindaman, "A new concept in computing," Proc. IRE (Correspondence), vol. 48, p. 257; February, 1960.

[2] Expansion theorem: $f(X, Y, Z) = Xf(1, Y, Z) + \bar{X}f(0, Y, Z)$.

Correction to "Negative *L* and *C* in Solid State Masers"*

The last two sentences of the correspondence under the above title, which appeared on page 1157 of the June, 1960, issue of Proceedings, are incorrect, as several readers have pointed out privately.

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* Received by the IRE, October 26, 1960.

Discussion on Tunable Maser Hybrids*

The recent article on masers by Arams and Okwit[1] shows a multiple branch 3-db 90° hybrid in strip line in Fig. 10(b). This type of hybrid has some advantages over the more conventional quarter wave capacity coupled transmission line hybrid.[2,3]

While the multiple branch coupler is somewhat larger than the capacity coupled type, the tolerances required are easier to maintain since the narrow gap between the two lines is not needed. This also means that the high power capacity of the multiple branch hybrid is also likely to be greater than that for the capacity coupled type. The resistive dissipation loss of the multiple branch coupler is also due less to the decreased field intensity.

It might be noted that this multiple branch hybrid is similar in performance to the "rat race" type, Fig. 10(a), when an extra quarter wavelength of line is added to one output of the rat race at the band center. However, the quarter-wavelength of line which would be required in this application is much more frequency sensitive than the 90° phase shift difference obtained in the multiple branch hybrid.

The normalized impedance of the main line, the auxiliary line, and the connecting branch lines, and the frequency characteristics to be expected have been available for several years.[4,5]

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* Received by the IRE, September 2, 1960.

[1] F. R. Arams and S. Okwit, "Packaged tunable L band maser system," Proc. IRE, vol. 48, p. 871; May, 1960.

[2] W. L. Firestone, "Analysis of transmission line directional couplers," Proc. IRE, vol. 22, pp. 1686–1692; November, 1954.

[3] J. K. Shimizu and E. M. T. Jones, "Coupled transmission-line directional couplers," IRE Trans. on Microwave Theory and Techniques, vol. MTT-6, p. 404; October, 1958.

[4] J. Reed and G. J. Wheeler, "A method of analysis of symmetrical four-port networks," IRE Trans. on Microwave Theory and Techniques, vol. MTT-4, pp. 246–252; October, 1956.

[5] L. Young, "Branch guide couplers," *Proc. NEC*, vol. 12, pp. 723–732; 1956.

10.8-KMC Germanium Tunnel Diode*

Tunnel diodes offer considerable promise as versatile high-frequency oscillators, and as wide-band amplifiers or down-converter elements for microwave use. Investigations on germanium low time-constant junctions and low inductance-mount configurations led to the construction of tunnel diodes oscillating at fundamental frequencies up to 10.8 kMc.

Negative resistance *p-n* junctions were fabricated by forming degenerate *n*-type regions by means of an As alloy on degenerate *p*-type germanium doped with Al to 6×10^{19}–8×10^{19} atoms/cc. The alloying was done at approximately 500°C using a rapid heating and cooling cycle of a few seconds. The diodes were soldered in the 0.25 mμh low inductance metal-ceramic mount developed at General Telephone and Electronics Laboratories, Bayside, N. Y. (Type D4115 Sylvania tunnel diode).

Table I gives various measurements of diode parameters taken at several stages of junction etching. The junction was reduced from approximately 3 mils to less than 0.2 mils. I_p, V_p are the current and voltage

* Received by the IRE, September 30, 1960.

TABLE I

I_p ma	I_v ma	V_p mv	V_v mv	R_s ohm	I_p/I_v	k ma/$\mu\mu$f	C_m $\mu\mu$f	C $\mu\mu$f	f kMc
560	220	148	260		2.5				
58	24	96	250		2.4		24		
30	11.8	90	240	.83	2.5	3.2	8	8.1	3.88
18	8	86	230	1.25	2.2	4.1	4.4	4.85	5.34
11.6	4.9	84	230	1.7	2.4	3.6	3.2	3.14	6.25
8	3.3	84	230	2.5	2.4	3.7	2.2	2.16	7.73
4.7	2	84	230	3.8	2.4	3	1.55	1.27	9.2
3.7	1.6	84	230	5	2.3			1	9.65
2.9								.784	10.8

Increase in diode oscillation frequency with reduction in size of junction. Area of junction is proportional to the peak current I_p. Data were taken on a single unit at various stages of etching.

peaks of the forward characteristic, and I_v, V_v similar parameters for the valley. R_s is the series resistance, C_m the capacitance measured at the valley, f the frequency of oscillation, and k the ratio of peak current to junction capacitance, I_p/C. This ratio is proportional to the time constant of the junction which in this diode varied between 3 and 4.1 ma/$\mu\mu$f. The measurement of capacitance is not as accurate as the measurement of peak current, therefore k was averaged to 3.7, and the calculated junction capacitances C for all the junctions were obtained from $C=I_p/3.7$.

Assuming a peak-to-valley current ratio of $I_p/I_v=2.2$ for the smallest junction, one obtains an average negative resistance $R_{AV} = V_p - V_v/I_p - I_v$ of 91 ohms. The minimum negative resistance for the same junction over a linear-voltage swing $\Delta V=0.088$ volt is 55 ohms. This diode has a rather poor current ratio, and the difference between the average and minimum negative resistances is smaller than that occurring in higher current ratio diodes. Although the current ratio of this diode was poor, other similarly constructed units had much higher current ratios; in one it was 15 and oscillated at 7.8 kMc.

The time constant for the smallest junction, based on $R=55$ ohms and $C=.784$ $\mu\mu$f, is 43×10^{-12} seconds. This value is considerably higher than the 5×10^{-12} seconds time constant reported for InSb; however, the frequency of oscillation reported for the latter was much lower than 10.8 kMc.[1]

The measurements are performed by placing these diodes into an oscillator circuit consisting of a circular cylindrical cavity, in the center of which the diode is mounted. The diameter of the cavity is just slightly larger than the diode package itself. The circuit can be looked upon as a radial transmission line loaded by the capacitance of the diode. Bias is supplied through a bypass capacitor.

In Fig. 1 we plot the experimentally observed frequencies vs peak current. This diagram also shows calculated frequencies based on the expression[2]

$$f = 1.59 \times 10^9 \sqrt{\frac{40k}{I_p} - \frac{k^2}{\Delta V^2 3.36}}.$$

The junction capacitance has been expressed here in terms of k and I_p, and the negative resistance in terms of k, ΔV and a current

[1] R. Batdorf, G. Dacey, R. Wallace and D. Walsh, "Esaki Diode in Insb," *J. Appl. Phys.*, vol. 31, pp. 613–614; March, 1960.

[2] G. Dermit, "High Frequency Power in Tunnel Diodes," (to be published).

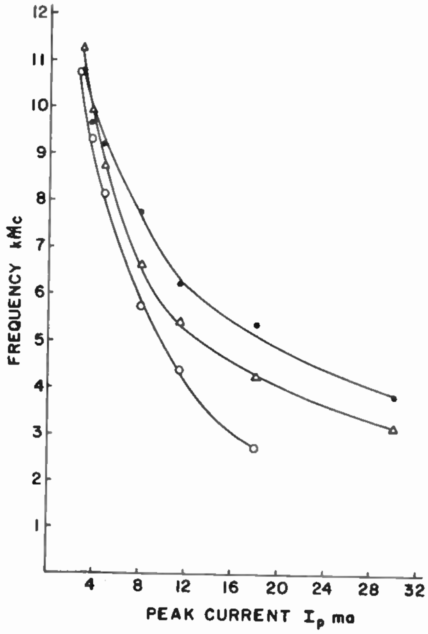

Fig. 1—Experimental and calculated frequencies of oscillation as a function of peak current. The diode peak current is proportional to the area of the junction. (● = experimental; △ = calculated, ΔV = 0.2 volt; ○ = calculated, ΔV = 0.088 volt.)

ratio of 2.2. The inductance was taken as 0.25 mμh. Calculated frequencies have been plotted for voltage swings of 0.088 and 0.2 volt. The former curve approaches the experimental value at the high-frequency side where the diode is operating near its cut-off whereas the latter is a better estimate at lower frequencies. According to these results, the voltage swing ΔV becomes smaller as the frequency approaches cut-off. With $C=0.784$ $\mu\mu$f, $R=55$ ohms and $R_s=6$ ohms, the cut-off for the smallest junction is at 11.2 kMc. When the frequency is much below cut-off, the voltage swings well into the valley region. This can be observed clearly by varying the bias. When the diode operates close to its cut-off, a small change in bias in either direction is enough to stop the oscillations, whereas for operation far below the cut-off frequency the bias can be shifted over a much wider range and into the valley region. The oscillation pattern at low frequencies is quite complex with strong harmonic content.

The power available at 10.8 kMc is estimated[2] to be 2.5 μW for an inductance of 0.25 mph. According to the same estimate this diode, if oscillating in a larger cavity at 6 kMc, will yield approximately 12 microwatts.

The contributions of J. McCarthy to the etching and measurements of these diodes is gratefully acknowledged.

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The Noise Figure of Negative-Conductance Amplifiers*

It is interesting to compare the noise figure of an electron device when used as a conventional amplifier and when used as a negative conductance amplifier. A general equivalent of a noisy amplifier can be represented as in Fig. 1. The box contains a fictitious perfect amplifier having the same voltage gain A as the real amplifier, and Z_1 is the input impedance of the real amplifier.

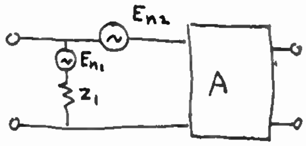

Fig. 1

The voltage E_{n1} is the rms sum of the thermal noise voltage in Z_1 and the amplifier induced noise (*e.g.*, induced grid noise). E_{n2} represents an equivalent internal source of noise assumed uncorrelated with E_{n1} (*e.g.*, shot noise). If the signal source consists of an admittance Y_s shunted by a thermal noise current I, and is coupled to the input terminals of Fig. 1 by an ideal transformer of turns ratio N, the noise figure F will be

$$F = 1 + \frac{E_{n2}^2(1 + Y_s N^2 Z_1)^2 + E_{n1}^2}{N^2 Z_1^2 I^2}. \quad (1)$$

To find the optimum value of N, equate dF/dN to zero, yielding

$$N^2 = \frac{\sqrt{E_{n1}^2 + E_{n2}^2}}{Y_s Z_1 E_{n2}}. \quad (2)$$

Substituting in (1) gives

$$F_{\text{opt}} = 1 + \frac{2Y_s}{I^2 Z_1}\left[E_{n2}\sqrt{E_{n1}^2 + E_{n2}^2} + E_{n2}^2\right]. \quad (3)$$

If now, the same amplifier is connected through a passive feedback network of gain β as in Fig. 2, the input terminals will show an admittance

$$Y_1 = \frac{1}{Z_1(1 - A\beta)} \quad (4)$$

which can have a negative real component by suitable choice of $A\beta$. The noise equivalent of Fig. 2 can be represented as a current

* Received by the IRE, August 1, 1960; revised manuscript received, August 26, 1960.

source I_n shunting Y_1. The value of I_n will be Y_1 times the open circuit voltage of Fig. 2. This voltage is given by

$$E_n^2 = A^2\beta^2 E_{n2}^2 + (1 - A\beta)^2 E_{n1}^2, \tag{5}$$

so that

$$I_n^2 = \frac{A^2\beta^2 E_{n2}^2 + (1 - A\beta)^2 E_{n1}^2}{Z_1^2(1 - A\beta)^2}. \tag{6}$$

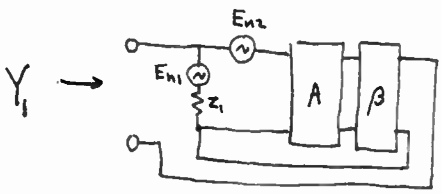

Fig. 2

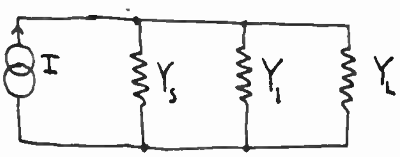

Fig. 3

Fig. 2 can be utilized as a negative conductance amplifier connected directly to the signal source (Fig. 3). Y_L is the load admittance, assumed small compared to Y_s, thus obviating the need to consider noise current introduced by Y_L. The power gain of Fig. 3 will be

$$G = \frac{4I_L^2 Y_s}{I^2 Y_L} = \frac{4Y_L Y_s}{(Y_s + Y_L + Y_1)^2}, \tag{7}$$

I_L being the current flowing in Y_L. G can be made arbitrarily large by making Y_1 sufficiently negative. The noise current I_n, and the noise current I from the source, are both amplified by the same factor; and therefore the noise figure will be

$$F' = 1 + \frac{I_n^2}{I^2}$$

$$= 1 + \frac{E_{n2}^2}{I^2 Z_1^2(1 - A\beta)^2} + \frac{E_{n1}^2}{I^2 Z_1^2}. \tag{8}$$

Solving (7) for Y_1, we have

$$Y_1 = 2\sqrt{\frac{Y_L Y_s}{G}} - Y_L - Y_s. \tag{9}$$

In general, we are interested in large values of G. Under the assumptions that $G \gg 1$ and $Y_s \gg Y_L$, (9) will simplify to

$$Y_1 \approx -Y_s. \tag{10}$$

If now, the source is coupled through an ideal transformer of turns ratio N, I will be transformed to NI and Y_s will become $N^2 Y_s$. By substitution of these conditions, along with (10) and (4), (8) becomes

$$F' = 1 + \frac{E_{n2}^2 N^2 Y_s^2}{I^2} + \frac{E_{n1}^2}{N^2 I^2 Z_1^2}. \tag{11}$$

As before, equating dF'/dN to zero yields the optimum value for N:

$$N^2 = \frac{E_{n1}}{E_{n2} Y_s Z_1}, \tag{12}$$

and substituting (12) in (11),

$$F_{\text{opt}}' = 1 + \frac{2E_{n2}E_{n1}Y_s}{Z_1 I^2}. \tag{13}$$

By comparing (13) with (3), we see that the negative conductance amplifier will show a significantly better noise figure, whenever E_{n2} is comparable in magnitude or greater than E_{n1}. This analysis possibly explains the low noise figure reported by Ishii,[1] obtained with an ordinary reflex klystron.

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[1] K. Ishii, "Noise figures of reflex klystron amplifiers," IRE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 291–294; May, 1960.

Nonuniform Sampling and Extremal Approximation*

Let a continuous function $g(t)$ be differentiable to order q as $d^q g(t)/dt^q = g^{(q)}(t)$. Locate all maxima and minima of $g^{(q)}(t)$ and interconnect these points with straight lines, assuming finite line slopes. Integrate the broken-line approximation q times; there results a smooth approximation to $g(t)$ which is continuous for all derivatives up to and including the qth. Provided a waveform is band limited and q is sufficiently large, the approximation to $g(t)$ may be made as exact as is desired.

If the broken-line approximation to $g^{(q)}(t)$ is differentiated twice, there results an impulse approximation to the same waveform which is readily described in terms of impulse values a_{kq} at times T_{kq}; a total of $q+2$ integrations then achieves the desired approximation to $g(t)$ which, in Laplace transform notation, is

$$\mathcal{L}g(t) \cong \sum_{k=1}^{\infty} \frac{a_{kq}\exp(-sT_{kq})}{s^{q+2}}.$$

The broken-line approximation to $g^{(q)}(t)$ is indicated in Fig. 1(a). The information contained in the sample points of Fig. 1(a) is identical to that included in the approximations to the same $g^{(q)}(t)$ in Figs. 1(b) and 1(c). Note, however, that the several approximations are not obtainable one from the other for the same $g^{(q)}(t)$ by linear procedures such as integration or differentiation.

It has been implied that use of ever larger values for q achieves an increasingly exact approximation for $g(t)$. This is not true, however, unless $g(t)$ is "absolutely" band limited such that spectral components are

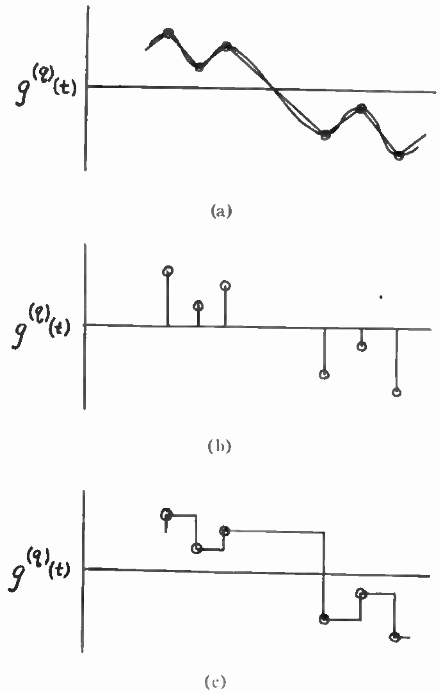

Fig. 1—Various approximations for $g^{(q)}(t) = d^q g(t)/dt^q$. (a) Broken line; (b) sample; (c) step function.

identically zero for all $f > B$ where B is band-limit frequency. If $g(t)$ is not band limited in this sense, no unique maximum frequency can be defined.

A qth derivative changes the spectrum of $g(t)$ by factor f^{2q} which enhances frequencies near the band limit B increasingly with derivative order q. In the limit, only single frequency B remains. We thus have derived the Nyquist sampling theorem: Arbitrarily exact approximation to an absolutely band-limited waveform with band-limit frequency B requires uniform sampling at a rate $2B$ samples per second. Nonuniform sampling at an average rate less than $2B$ can thus do no more than approximate a waveform.

In the narrow-band case, the simple broken-line approximation to $g(t)$ (for $q=0$) is an exact one, at least in the limit of very small bandwidth. Another kind of approximation is thus suggested for a wide-band signal; the band is broken into a number of contiguous bands which are individually narrow and the waveform in each band is approximated using $q=0$. Clearly, as the number of bands increases, the accuracy of the approximation improves. An example of band-splitting methods is the speech Vocoder (except in this case approximation is incomplete because data concerning relative phase angles in the various bands are customarily ignored).

It is evident in the narrow-band case (for $q=0$) that extremal approximation achieves estimates of equivalent amplitude and frequency modulating components of the waveform. The same thing occurs in the wide-band case, although difficulty may arise in the physical interpretation of generalized envelope and angle.

A speech waveform may be approximated to a first order as in Fig. 1 for $q=0$. Mathews reports results of such approxima-

* Received by the IRE, September 6, 1960.

tion using digital simulation.[1] He obtains a parabolic fit to sample points and thus "simulates" parabolic-segment approximation; however, the basic approximation remains a first-order one. More recently, Spogen, Shaver, and Baker[2] at University of Arizona have constructed a real-time speech system based on step-function approximation as in Fig. 1(c) in which good results are obtained in real-time without concern for the awkward "buffer" stage proposed by Mathews.

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[1] M. V. Mathews, "Extremal coding for speech transmission," IRE TRANS. ON INFORMATION THEORY, vol. IT-5, pp. 129–136; September, 1959. Mathews is concerned with sampling theory rather than with approximation and hence does not recognize that what he calls "extremal coding" is a first-order member of a general method for approximation.

[2] L. R. Spogen, H. N. Shaver, and D. E. Baker, "Speech Processing by the Selective Amplitude Sampling System," to be published in *J. Acoust. Soc. Am.*

Space-Charge Potentials in Depressed-Collector Design*

This note describes a simple, effective technique for increasing the efficiency of linear-electron-beam tubes by the use of depressed collectors. The main requirement for successful operation is that secondary electrons produced at the depressed collector do not reach the higher voltage interaction structure. We believe a purely electrostatic scheme is by far the simplest arrangement for suppression of secondary electrons. For a long time, space-charge potential has been used effectively in beam tetrodes to suppress the secondaries from the plate.

To apply this principle to a collector design, it must be realized that the general geometrical situation in a linear-beam tube is different from a tetrode, in which the transverse dimension of the beam is usually large with respect to the longitudinal dimension. In a linear-beam tube, the transverse beam dimension is relatively small compared to the longitudinal dimension traversed by the beam. Thus, collector aperture effect and the critical distances between electrodes and the surrounding boundaries must be taken into consideration for a successful collector design that will achieve definite potential minimum in front of the collector entrance.

The particular solutions of Poisson's equation ($\nabla^2 V = -\rho/\epsilon$) that allow depressed-collector potentials with effective secondary-electron trapping were used. (Fay, Samuel, and Shockley Type C.)[1] In linear-beam tubes, the electron beam enters a region bounded by two parallel planes at different potentials, separated in the axial or z direction. Certain solutions yield a potential minimum in the region between the two bounding planes, and this "well" is ideally suited for trapping secondary and reflected primary electrons from the collector surface.

To test the theory, a Pierce gun with a uniform beam emerging at point A (Fig. 1) was used. At A, the beam is parallel, all radial velocity being zero, and a focus structure maintains a parallel beam with a potential minimum in front of the collector. The collector operates at 72 per cent below anode, or at 28 per cent of beam voltage above cathode. The focus structure operates at a cathode voltage that causes the potential minimum of the beam to be 80 per cent below the anode, making the well sufficiently negative with respect to collector to trap most slow-speed secondaries for ordinary vacuum tube materials at $V_b < -500$ volts.

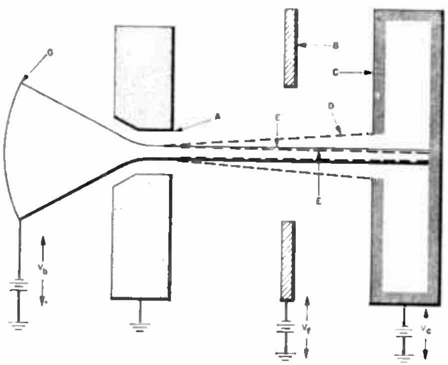

Fig. 1—Pierce gun. A = anode, B = focus structure, C = collector, D = divergent beam, E = parallel beam, F = convergent beam, and G = cathode.

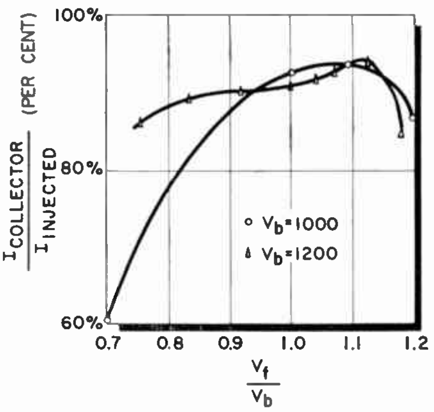

Fig. 2—Current transmission vs normalized focus voltage.

A typical result of net collector current as a function of normalized focus structure potential is given in Fig. 2, where V_b, V_c, and V_f denote beam potential, collector potential, and focus-structure potential, respectively. It is significant that the net collector beam current (primaries minus secondaries) reaches a maximum close to V_f/V_b equal to one, and that the normalized collector current is over 93 per cent with respect to injected current. At this maximum point it is believed that the beam is behaving as designed.

On either side of the maximum, net collector current drops rapidly, indicating reflected primaries or a large return of secondaries to the anode region. The large decrease in current as V_f approaches anode voltage will cause the beam to diverge slightly, leaving the possibility of the disappearance of the potential minimum. Then the trapping of secondaries is no longer possible and a decrease in net collector current must take place. On the other hand, since beam transmission maximizes at a ratio of V_f/V_b slightly larger than one, it is reasonable to assume that a slight beam convergence exists that would partially compensate for the lens effect. When V_f/V_b exceeds 1.2, a virtual cathode is created in front of the collector, reflecting most of the current to the anode.

Theoretically, when ϕ, the ratio of collector to anode potential (referred to the cathode), is less than 0.35 in Type C operation, the desirable potential minimum is unstable because of the simultaneous existence of Type B operation in which the injected current will be partially reflected. For ϕ larger than 0.35, stable single-mode Type C operation occurs if the current injected into the space does not exceed a beam perveance limit:

$$K = 2.33 \times 10^{-6}(3.2\phi + 0.8)2(A/S^2)$$

where A is the beam area and S is the equivalent diode distance for the anode operating voltage.

To design a depressed collector to operate with ϕ below 0.35, a scheme using multiple-stage electrodes (with various operating voltages other than the cathode potential alone) will be necessary. This procedure will define one of the beam boundary potentials close to the beam, so that the section closer to the collector will operate essentially within the Type C mode. We have succeeded in designing depressed collectors with ϕ as low as 0.04 with no observed instability.

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* Received by the IRE, September 8, 1960.

[1] C. E. Fay, A. L. Samuel, and W. Shockley, "On the theory of space charge between parallel plane electrodes," *Bell. Sys. Tech. J.*, vol. 17, pp. 49–79; January, 1938.

A Helix-Type Variable Capacitor*

INTRODUCTION

The present widely used types of variable capacitors obtain a continuously variable capacitance by changing the effective capacitive area between a fixed number of rotor and stator electrodes. Hence, the number of elements, or electrodes, contributing to the total capacitance is fixed and the

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range of effective area of engagement of these elements is obtained by something less than 360° rotation of the rotor.

The variable capacitor described herein is based upon a double helix relationship between the electrodes. In obtaining an incremental change in capacitance, motion takes place in three dimensions, rotational and longitudinal, as compared with two dimensional variation in conventional variable capacitors, *e.g.* rotational only.

Description

In referring to Fig. 1, the stator helix (A) is spatially fixed, while the rotor helix (B) is rotated within it in the manner of a nut and screw. As the rotor or inside helix is rotated, effective capacitive area is varied by both rotational and longitudinal movement. Full engagement of the inner and outer helices corresponds to the maximum capacitance, determined by the length of the full engagement, the inner and outer helix diameter, the shaft diameter, the lead angle of the helices and the dielectric used. Full disengagement of the inner and outer helices would theoretically result in zero capacitance. It should be noted that the rotor-stator relationship referred to above may be interchanged, that is (B) can be spatially fixed while (A) rotates and moves along the inner helix. In either case the overall length of the assembly is twice the full engagement length.

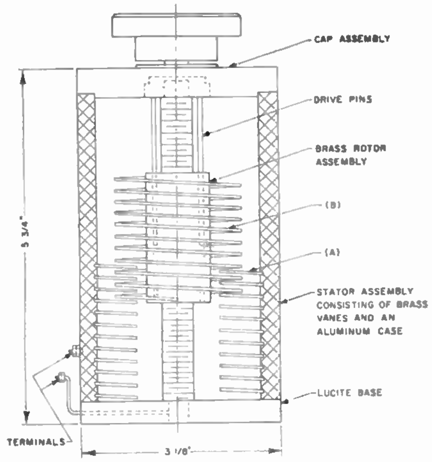

Fig. 1—Helix variable capacitor.

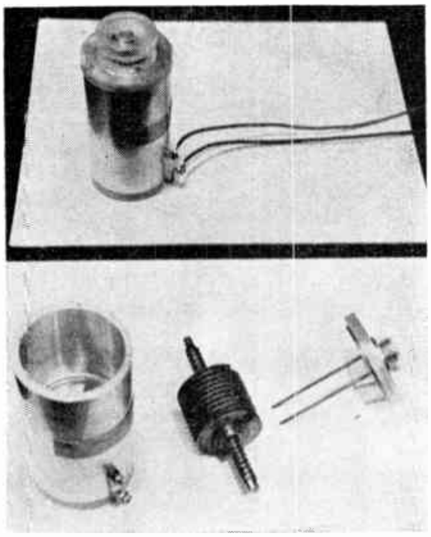

Fig. 2—Helix variable capacitor.

A prototype helix variable capacitor is shown in Fig. 2.

Applications

By applying proper additional mechanism or dielectric material, the capacitor has a number of applications for/in

1) a computer component: precision and wide range capacitance.
2) communication instruments: *i.e.*, a wide range tuner without a band switch. Also, it can be combined with a variable coil in high frequency range application.
3) a standard linear capacitor.
4) a precision wide range transducer of mechanical rotation vs electrical output (voltage, frequency, etc.).

Acknowledgment

The author wishes to thank B. W. Morris, H. Olson and C. N. Batsel for their kind assistance and support. The author also thanks Dr. I. Whiteman for his encouragement.

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Because capacitance variation is obtained by motion in three dimensions, this capacitor has many advantages in comparison with present two dimensional variable capacitors.

1) The rotation of the rotor is not limited to 180° or to 360°. Smaller capacitance increments can be obtained per revolution.
2) Maximum and minimum capacitance ratios of 10,000 or more may be obtained while present capacitors are limited to the ratio of 10 to 100.
3) A linear capacitance increment may be obtained.
4) This capacitor occupies the same or less space as present capacitors per unit capacitance.

Quartz AT-Type Filter Crystals for the Frequency Range 0.7 to 60 Mc.*

For application of quartz crystals to filters, the motional parameters of the crystal, *i.e.*, the capacitance C_1, the resistance R_1 and the shunt capacitance C_0 must be known. In addition, the crystal must be free from unwanted modes in a specified range in the vicinity of the wanted frequency.

Design data leading to suppression of unwanted modes to at least 40 db below the main mode for AT-type crystals in the frequency range 0.7 to 60 Mc are given.

The behavior of unwanted modes in thickness-shear vibrating quartz plates, *e.g.*, AT-fundamental or overtone modes, is determined by two parameters Φ_a/t and Φ_e/t, where Φ_a and Φ_e are the plate diameter and the electrode diameter, respectively, and t the thickness. The electrodes are assumed to be circular and accurately centered.

For plates vibrating at frequencies higher than 10 Mc, Φ_a/t is usually larger than 50. Unwanted modes then depend in first approximation on Φ_e/t and are practically independent of Φ_a/t.

For circular quartz plates in the range 10 to 30 Mc, AT-type, fundamental, unwanted modes are suppressed to −40 db and more below the main mode when $\Phi_e/t \cong 18$.

Triangular shaped quartz crystals[1] in the range 10 to 30 Mc, fundamental, are suitable filter crystals when $\Phi_e/t \cong 24$ or less. The performance of triangular crystals is superior when one side of the triangle is parallel to the X-axis.

Circular quartz plates in the range 30 to 60 Mc, AT-type, third overtone, are sufficiently free of unwanted modes for $\Phi_e/t = 10$ or less. The limitation for practical application is Φ_e which becomes very small at higher frequencies.

The motional capacitance C_1 follows from

$$C_1 = \Gamma \frac{a}{t} \tag{1}$$

where Γ (see "IRE Standards"[2]) is the motional capacitance constant measured in 10^{-6} $\mu\mu$f mm^{-1}, a the area of the electrode in mm, t^2 the thickness of the plate in mm and the motional capacitance C_1 in $\mu\mu$f. Γ depends slightly on Φ_e. Thickness t and frequency f are related by the frequency constant $N = f \cdot t$ when f is in Mc: for the fundamental mode $N_1 = 1.660$ Mc mm, for the third overtone $N_3 = 4.980$ Mc mm. Eq. (1) can be written as

$$C_1 = \frac{K}{f} \tag{2}$$

where

$$K = N\Gamma \left(\frac{\Phi_e}{t}\right)^2 \frac{\pi}{4} \ \mu\mu\text{f} \cdot \text{mc}. \tag{2a}$$

The values[3] for Γ corresponding to the plate size Φ_e, K and C_1 for circular and triangular plates, fundamental mode (10 to 30 Mc) using $\Phi_e/t = 18$ and 24, respectively, and for circular plates, third overtone (30 to 60 Mc) $\Phi_e/t = 10$ are shown in Table I.

For plates with a diameter-thickness ratio smaller than 60, two ranges are considered:

1) $\frac{\Phi_a}{t} = 60$ to 25, 2) $\frac{\Phi_a}{t} < 25$.

In the range $(\Phi_a/t) < 25$, the unwanted

* Received by the IRE, August 23, 1960; revised manuscript received, September 2, 1960.

[1] R. Bechmann, "High-frequency quartz filter crystals," Proc. IRE, vol. 46, pp. 617–618; March, 1958.

[2] For the definition of Γ, see the "IRE Standards on Piezoelectric Crystals: Determination of the Elastic, Piezoelectric and Dielectric Constants—The Electromechanical Coupling Factor, 1958," Proc. IRE, vol. 46, pp. 765–778; April, 1958.

[3] The accuracy of Γ for the fundamental mode is about ±5 per cent, for the third overtone, at present, about ±10 per cent.

TABLE I

DATA FOR FILTER CRYSTALS IN THE RANGE 10 TO 60 MC

f Mc	Order n	Form	Φ_e mm	Γ 10^{-6} μμf mm^{-1}	K 10^{-4} μμf Mc	C_1 10^{-4} μμf
10	1	Circ.	2.988	200	845	84.5
10	1	Triang.	3.984	190	1427	142.7
20	1	Circ.	1.494	215	908	45.4
20	1	Triang.	1.992	210	1577	78.9
30	1	Circ.	0.996	220	929	31.0
30	1	Triang.	1.328	217	1630	54.3
30	3	Circ.	1.660	21.5	84.1	2.80
40	3	Circ.	1.245	22.7	88.8	2.22
50	3	Circ.	0.996	24	93.9	1.88
60	3	Circ.	0.830	24	93.9	1.57

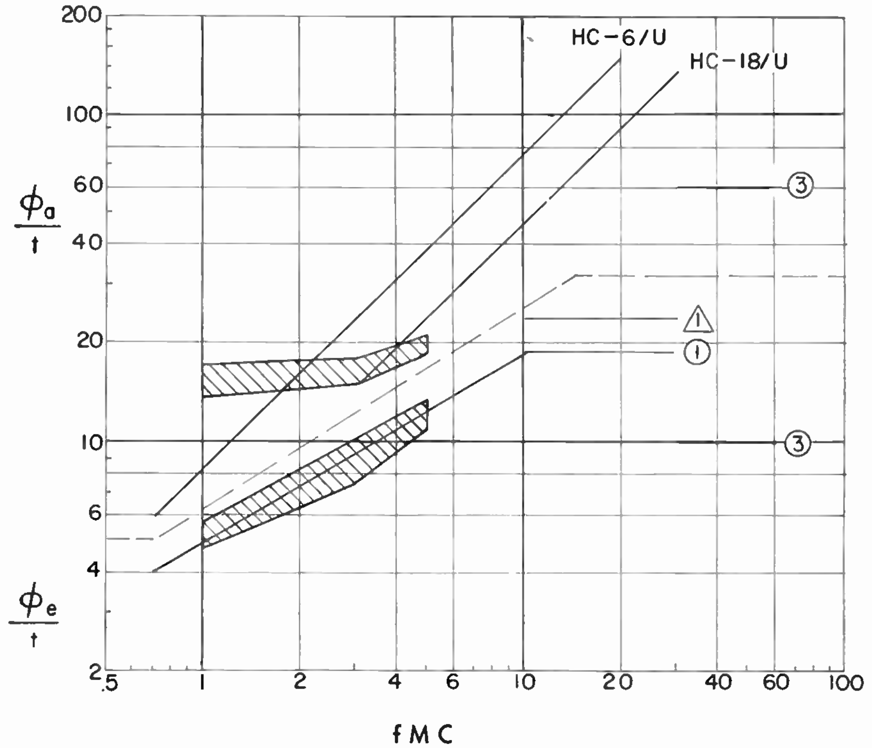

Fig. 1—Φ_a/t and Φ_e/t for fundamental and third overtone AT-type quartz crystals for optimum suppression of unwanted modes. ① ③ and △ refer to fundamental and third overtone circular plates and fundamental triangular, respectively.

modes can be suppressed by adequate bevelling of the edges of the plate.[4,5] The size of the electrodes becomes unessential. For plates having $\Phi_e/t=25$ to 60, bevelling becomes less effective with increasing Φ_a/t. The parallelism of the plate becomes more and more essential and Φ_e/t must be sufficiently small.

Small changes of the radius of the bevel give rise to essential changes in crystal performance. In the range 0.7 to 10 Mc, suitable dimensions for both ratios Φ_a/t and Φ_e/t can be found to suppress unwanted modes. In this frequency range, C_1 can be calculated from $\Gamma=210\cdot10^{-6}$ μμf mm$^{-1}\pm5$ per cent.

Fig. 1 gives a survey of Φ_a/t and Φ_e/t resulting in optimum suppression of unwanted modes. The frequency in the range 0.7 to 60 Mc is plotted against Φ_a/t (upper part of the ordinate) and Φ_e/t (lower part). Curves representing typical values for Φ_e/t are shown for plate sizes corresponding to crystal units HC-6/U and HC-18/U. In the frequency range 1 to 5 Mc, two cross-hatched areas are shown: one for Φ_a/t and the corresponding range for Φ_e/t. These areas represent dimensions used by industry. In the frequency range 0.7 to 2 Mc, very small diameter-thickness ratios can be used provided the plates are adequately bevelled.

The behavior of unwanted modes with respect to the electrode diameter can be understood by considering the distribution of the inharmonic overtones of a plate. A more comprehensive paper will follow at a later date.

This information has been obtained from measurements by the author over a long period of time. In addition, the author acknowledges with grateful appreciation, information obtained from Hermes Electronics Company, Contract Nos. DA36-039 SC-73234 and DA36-039 SC-78242, USASRDL, James Knights Company, McCoy Electronics Company, Piezo Crystal Company, Reeves-Hoffman Division, and Union Thermoelectric Corporation.

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[4] R. Bechmann, "Single response thickness-shear mode resonators using circular bevelled plates," *J. Sci. Instruments*, vol. 29, pp. 73–76; March, 1952.

[5] R. Bechmann, "Filter Crystals," *Proc. Twelfth Annual Symp. on Frequency Control*, USASRDL, Fort Monmouth, N. J., pp. 437–474; May 6–8, 1958.

Early History of Parametric Transducers*

In his paper on the early history of parametric transducers,[1] W. W. Mumford gives a list of two hundred references (papers and patents) related to actual parametric devices and to allied subjects. May I be permitted to add an additional reference which falls into the gap between 1936 and 1944 shown after Reference 24?

When working in a consultant capacity for this company I reported on June 14, 1941, on the possibility of obtaining a power gain in a reactive frequency changer. A United Kingdom patent application was filed on June 30, 1941, but did not result in the grant of U.K. Letters Patent. A corresponding Indian Application No. 28917 was filed on June 29, 1942, with a priority claim based on the United Kingdom application and was accepted on March 3, 1943. The specification of the Indian patent represents the only publication available on this work. The original United Kingdom patent appplication was made available to Government departments and I have also been informed that my original report was discussed with British government research organizations.

The Indian patent discusses a 4-terminal network in which a variable capacitor forms the series element. It derives an equivalent circuit in the form of a π-network with constant reactances and a controlled current source across the output terminals and derives the relation $W_I/W_{II}=\omega_1/\omega_2$ between input and output power from a consideration of the input resistance reflected into the primary terminals by a load at the secondary terminals. It suggests that a variable capacitor might be constructed from an inherently non-linear dielectric like rochelle salt, but it also suggests that this variable capacitor might be a plate condensor between the plates of which a stream of electrons of variable density is passing.

In an additional note circulated to my colleagues during 1941 and 1942, attention was also drawn to the analogy that exists between such a system and a synchronous generator with electrical input or output to rotor and stator (the mechanical input being the analogue of the pump power).

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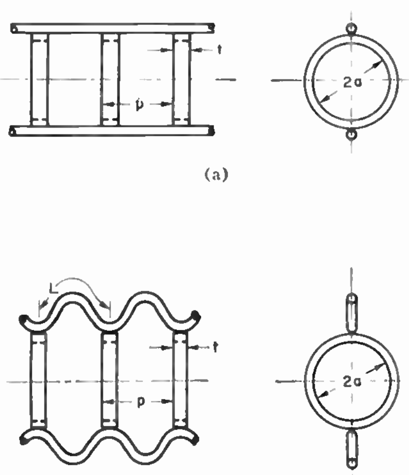
* Received by the IRE, September 23, 1960.

[1] W. W. Mumford, "Some notes on the history of parametric transducers," PROC. IRE, vol. 48, pp. 848–853; May, 1960.

An Interaction Circuit for Traveling-Wave Tubes*

This note describes a slow-wave propagating structure suitable for use in high-power traveling-wave tubes. Cold-test measurements indicate bandwidths of the order of an octave at megawatt peak power levels.

Several years ago, E. L. Chu, of Stanford University, Calif., predicted the theoretical performance of a two-wire transmission line periodically loaded with circular rings, the circuit shown in Fig. 1(a). His calculations suggested a wide-band circuit with reasonable interaction impedance, but with phase velocities too fast for use in traveling-wave tubes. More recently, as a consultant to Sperry Gyroscope Company, Great Neck, N. Y., Chu suggested that a phase-velocity reduction might be achieved by replacing the parallel-wire portion of the circuit by an inherently slower guiding structure. Following Chu's suggestion, a series of slower circuits were examined experimentally. Of this series, the most effective has been a pair of coplanar meander lines having reflection symmetry about the propagating axis [Fig. 1(b)]. In addition to making the necessary phase-velocity correction, this modification enhanced the interaction impedance while reducing the upper cutoff frequency of the pass band.



(a)

(b)

Fig. 1

Experimental results are presented in terms of normalized circuit dimensions derived from the actual circuit dimensions. L is the developed meandering wire-length per period, a is the ring radius, and t is the ring thickness measured axially. Also, k and β are the free space and axial wave numbers, respectively.

Dispersion and impedance characteristics for a series of ring-loaded meander lines are depicted in Figs. 2 and 3 which show variation with the following parameters:

$$\sigma = L/p; \quad \cot\theta = \frac{2\pi a}{p}; \quad \delta = t/p.$$

* Received by the IRE, September 30, 1960.

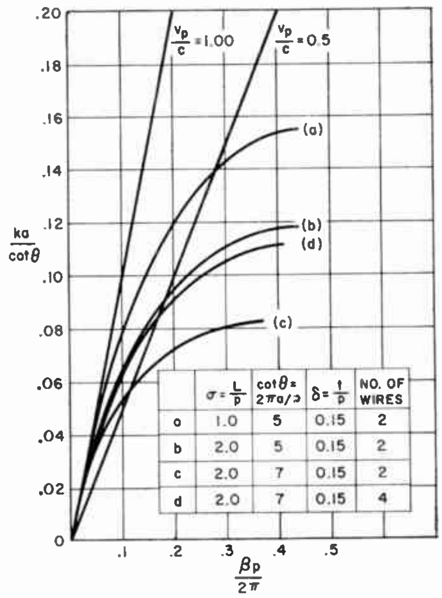


Fig. 2—Normalized dispersion curves for variations of the ring-loaded meander line.

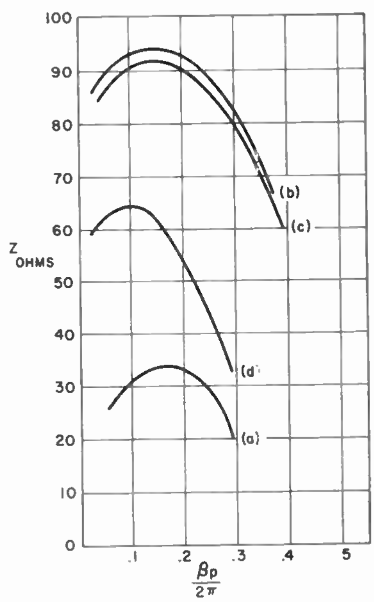


Fig. 3—Pierce impedance curves for variations of the ring-loaded meander line.

Curves a and b compare the dispersion and impedance characteristics for different values of σ, holding $\cot\theta$ and δ fixed. Note that $\sigma = 1$ represents the previously mentioned ring-loaded transmission line. Curves b and c show the variation with $\cot\theta$, keeping the other parameters fixed. In addition, it was found that dispersion increased and impedance decreased with increasing δ; therefore, the optimum ring thickness is the minimum value permitted by mechanical and thermal constraints.

The preceding curves demonstrate a circuit which has useful phase velocities between .4 and .8 times the speed of light, c, and interaction impedance approaching 100 ohms.

Dispersion curves have been normalized to preserve identical phase velocity lines. This normalization obscures the fact that the upper cutoff of ka is relatively independent of variations in σ, $\cot\theta$, and δ. This upper value of ka can be increased by the use of a greater number of meandering wires. In curves c and d the dispersion and impedance characteristics of a two-wire model are compared with a four-wire model for similar values of σ, $\cot\theta$, and δ. The four-wire model had identical meanders located azimuthally at 90-degree intervals instead of 180-degree intervals as in the two-wire model. This improvement in ka is obtained at the expense of impedance and higher phase velocity.

When considering the circuits studied, a number of other features that make them suitable for use in high-power traveling-wave tubes become apparent: 1) The circuits are free of interacting backward-wave modes within the pass band, and the lower cutoff of the next higher pass band is approximately 1.5 times the upper cutoff of the useful pass band. This implies that spurious-mode suppression may be achieved by relatively simple frequency-rejection schemes. 2) The circuit form is such that fabrication from tubing, which may then be liquid cooled, is feasible. 3) From a structural viewpoint, the circuit is somewhat self-supporting, particularly in the case of the four-wire model. For the two-wire model, ceramic rod supports may be used without drastically changing the dispersion characteristics.

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An Experiment of Ion Relaxation Oscillation in Electron Beams*

In many microwave tubes using high density electron beams a spurious modulation in the drift space has frequently been observed which is a deleterious effect caused by ions.[1] We have done some experiments on this spurious modulation using a Hernquist-type ion trapping gun as shown in Fig. 1. This experimental tube cannot produce a well-shaped high-density electron beam, but can easily produce a higher perveance electron beam than could be obtained from a Pierce-type gun. The difference between this experimental tube and Hernquvist's tube is that our tube has a rectangular cathode and produces a strip beam. The cathode area is 7.5 by 15 mm and its surface radius of curvature is 32 mm with its center located at the planar electrode G_3. Electrode G_1 is a mesh grid having about 40 per cent beam transmission and it injects an electron beam into the space surrounded by the electrodes G_1, G_2, and G_3, by applying positive

* Received by the IRE, April 5, 1960; revised manuscript received August 31, 1960.

[1] C. C. Cutler, "Spurious modulation of electron beams," Proc. IRE, vol. 44, pp. 61–64; January, 1956.

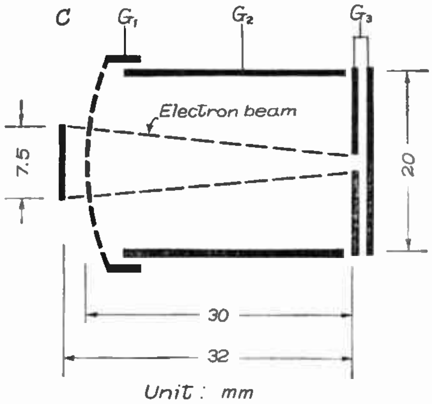

Fig. 1—Dimension of the experimental tube.

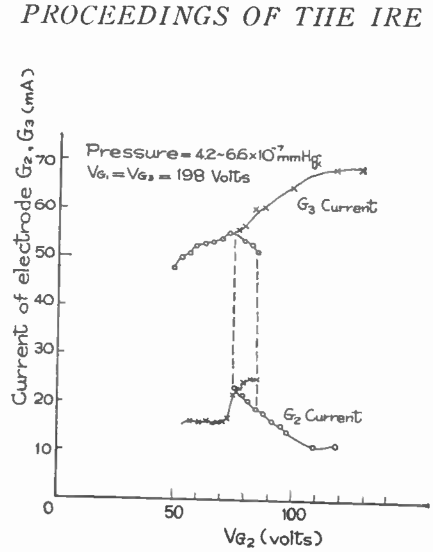

Fig. 2—Current division of electrode G_2, G_3 near a virtual cathode by V_{G2}.

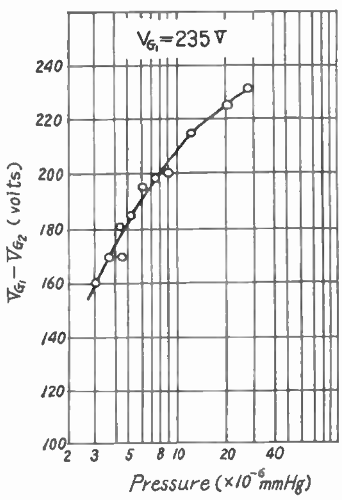

Fig. 3—Experimental curve showing a relation between pressure vs V_{G2} producing a virtual cathode under $V_{G1} = V_{G3}$, constant injection current.

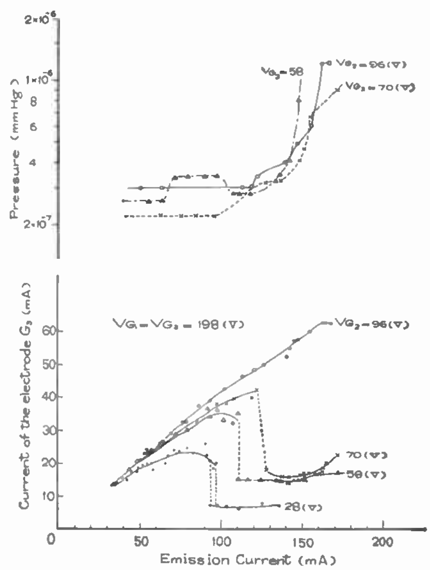

Fig. 4—Experimental relation between the current of the electrode G_3 and the emission current when V_{G2} alone is fixed, and pressure is varied.

Fig. 5—A photograph of ion relaxation oscillation.
Pressure . . . . 5.1×10⁻⁷ mm Hg.
Oscillation current, peak to peak . . . 22 ma.
Oscillation voltage . . . . 0.75 v.
Oscillation frequency . . . . 3 kc.

pulses. The inside surface of the electrodes was sprayed with carbon powder in order to prevent secondary emission. The diode which exists between the cathode and the accelerating grid G_1 has a perveance of 75 microperves. When the cathode emission is very low the space-charge effect of the injected electrons is so small that the beam tends to converge within the slit centered on electrode G_3.

As the space charge effect is increased, the beam spread is larger, and considerable current begins to flow to the plate having a slit in the electrode G_3. A potential well due to space charge is formed in the electron beam and the higher the beam current density, the deeper the potential well becomes. Even under high vacuum conditions, Hernquvist[2] has shown that considerable ions which were caused by collision were accumulated in this potential well and neutralized the negative space charge of the electron beam. This was also shown to be true in our experiments by means of measuring each current of two plates of the electrode G_3 under various conditions. When V_{G2}($V_{G1,2,3}$ is the voltage of the cathode to the electrode $G_{1,2,3}$) was gradually decreased in this tube, a virtual cathode was abruptly produced. When the injected current and the pressure were kept constant (V_{G1} was set equal to V_{G3}) and the potential on G_2 was varied, the current division between the electrodes G_2 and G_3 was observed as shown in Fig. 2.

In Fig. 2 it may be seen that the right side of the current discontinuity is a stable beam without a virtual cathode. On the left side of the current this continuity is not a stable beam and possesses a virtual cathode as indicated by the fact that spurious modulations occur in the beam current. At pressures greater than 10^{-5} mm of mercury these spurious modulations are caused mainly by ion plasma oscillations;[3] and at pressures less than 10^{-6} these spurious modulations are caused by relaxation-type oscillations. Between 10^{-5} and 10^{-6} mm of mercury, both types of instability are observed simultaneously. The characteristics as shown in Fig. 2 are sensitively dependent upon the pressure. When the pressure alone is varied the variation of the discontinuity points associated with V_{G2} is shown in Fig. 3. This discontinuity moves higher in voltage $V_{G1}-V_{G2}$, and becomes more indefinite, as the pressure is increased. The relation between the current of the electrode G_3 and the emission current for the case when V_{G2} alone is fixed is shown in Fig. 4. Current flowing to the electrode G_3 is abruptly intercepted by a virtual cathode caused at a certain emission current. The spurious modulation was observed in the right side of the discontinuity in this figure. By a fine adjustment of voltage V_{G2} near the discontinuity it was noted that the relaxation oscillations could be increased until their amplitude was nearly equal to the difference between the discontinuity currents. For example, one of the oscillation figures is shown in Fig. 5. Frequencies of relaxation oscillation are very sensitive to pressure but are reproducible. From the above experimental results it is seen that there is an oscillation mechanism excited by ions under critical perveance[4] in the electron beams. In electron beams which are dense enough so as to have virtual cathodes, ion oscillations cannot be eliminated without the achievement of an ultra-high Vacuum during the operation.

Valuable discussion with Professor Wada and Dr. Yamanaka are gratefully acknowledged.

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[3] W. W. Peterson and H. Puthoff, "A theoretical study of ion plasma oscillations," IRE Trans. on Electron Devices, vol. ED-6, pp. 372–377; October, 1959.

[4] J. C. Twombly, "Shot noise amplification in beams beyond critical perveance," 1957 IRE WESCON Convention Record, pt. 3, pp. 156–162.

Can the Social Sciences Be Made Exact?*

The recent article by Berkner[1] has prodded us to comment on some of the

* Received by the IRE, August 18, 1960.

[1] L. V. Berkner, "Can the social sciences be made exact?" Proc. IRE, vol. 48, pp. 1376–1380; August, 1960.

topics he treats and his methods of treatment. It is in the hope of stirring further consideration and interest from the community in the activities he considered that we write this letter. In this respect we are aware, also, that a survey article, of necessity, cannot contain excessive detail. However, we feel that further scope should have been given to his "speech-made-paper."

The author cites certain of the pioneering works of Hartley, Shannon and Wiener, and observes, quite rightly, that these constitute a well-defined mathematical science of considerable generality. Then he observes that information is useless unless it is transmitted in an assimilatable form. This suggests to us that, in his brevity, he has not made it clear that there are at least three definitions of the word "information." There is a vague sort to be found in dictionaries. There is the measure of information proposed by Shannon[2] and Wiener.[3] They define "information" in the sense that they tell, pragmatically speaking, how to measure a certain entity which is a primitive concept to be treated syntactically and not to be decomposed further. And, third, there is the "semantic information" in the sense of Bar-Hillel[4] as based upon certain works of Carnap.[5] This is a measure of an entity associated with propositions of "language systems."

The author suggests that since "information" (the second sort in our description) is so well-understood in the mathematical sense, it should be applied to the social sciences more briskly, as the social sciences are badly in need of quantification of their data. He deplores the condition that one cannot seem to identify the "independent variables and parameters" of human and animal behavior.

We should like to question the principle that all disciplines, and in particular the social sciences, may be usefully constructed on the norm of the more classical aspects of the physical sciences. Psychology, once a subdivision of philosophy, separated from its parent by acknowledging the efficacy of mechanistic science and its methods. To what extent is the "amazing" resemblance between mechanistic circuits and man, viewed by the conventional psychologist, attributable only to this cause. If we choose to view all entities mechanistically and incidentally wish to study man, why should we be surprised at man's mechanistic nature?

In studies of social processes, it seems to us more meaningful than present-day schemes either to begin to modify what is meant by mathematical proof or to hunt for even richer mathematical treasures.

We point out that, for some time, an ever increasing use of principles of social science has been made in engineering. The statistical methods of the physical sciences come, by analogy, from the statistical methods of the older sciences of politics and economics. We also call attention to the study of gross interaction processes in the engineering sense of Pask.[6] Here, the consideration of "independent variables" of relations between local processes is taken to be irrelevant and meaningless. Here one destroys the concept of independent variables from within the theory rather than from without the theory. Further research should be devoted to devising more adequate logico-mathematical techniques for treating such processes.

The author brings out an excellent point in his discussion dealing with the recognition of talented students by average teachers. Exams ask of the student "What is the *answer* to such-and-such question?" But the more intelligent student is used to asking of himself "What *question* should I be asking myself?"

Finally, we wish to point out that it seems reasonable that communication engineers should engage in even more serious study of the effects of high-speed mass communications. For the first time in history, man has become swamped in a huge bulk of communication that must go largely uninterpreted. In the "olden days," people worried because certain crucial messages might fail to arrive on time for them to be acted upon. Now we must worry about so many messages arriving that we might not *discover* significant messages in time to act on them. Now certain communication channels are violating, legally, the First Amendment to our Constitution.

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[2] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. J.*, vol. 27, pp. 379–423, July, 1948; pp. 623–656, October, 1948.
[3] N. Wiener, "Cybernetics," The Technology Press, Mass. Inst. Tech., Cambridge, Mass.; 1948.
[4] Y. Bar-Hillel, "Semantic information," *Brit. J. Phil. Sci.*, vol. 4, pp. 147–157; August, 1953.
[5] R. Carnap, "Logical Foundations of Probability," University of Chicago Press, Chicago, Ill.; 1950.
[6] M. C. Yovits and S. Cameron, Eds., "Self-Organizing Systems," Pergamon Press, New York, N. Y.; 1960.

Author's Comment[7]

The comments of Mullin and Zopf are well taken and one cannot disagree in the main. Certainly, the statistical processes are powerful tools in probing the behavior of so complex a creature as man. Thus, thermodynamic processes describe a phenomenon precisely without knowing much about the individual molecules. But the study of the molecule in its relation to others can provide a richness of comprehension of the phenomenon that leads to successive and more complete generalizations. In this sense, the mechanistic approach may have much to offer. It is, of course, necessary to start from the premise that in consonance with natural science, man's behavior can be described as a consequence of precedent and generally definable events, and that no Aristotelian "final cause" is involved.

The whole problem of assimilation of mass high-speed communication pointed out by Mullin and Zopf is a serious one. The problem is not simply failure to interpret the large bulk of communication, but also the training, rules, and criteria under which interpretation is made. In this, credibility plays a major role, involving both the available facts underlying each view, and the experience and skill of the interpreter in objective evaluation of the data. In addition, the time required for the brain to fully assimilate the deep meaning of data is probably much longer than the time available to formulate the reply. Here the speed of the sailing ship in the days of Ben Franklin may have had advantages as well as disadvantages.

Here, the clear division of data into "tactical" and "strategic" has some merit. Tactical data require decision under an imposed time limit required by the situation. Strategic data impose no fixed time limit for decision, but may be ignored in favor of tactical data for this very reason. Consequently, the really important strategic decisions requiring accumulation of extensive data and their thorough assimilation may never be made. Perhaps the analysis proposed by Mullin and Zopf can lead to analytical methods for sorting and handling data more completely and effectively.

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[7] Received by the IRE, September 8, 1960.

Phase Stability of Oscillators*

INTRODUCTION

The frequency stability of an oscillator must always be defined with reference to measurement time. In general, we can distinguish the principal determinant of frequency instability for each of three measurement interval magnitudes. This is illustrated in Fig. 1.

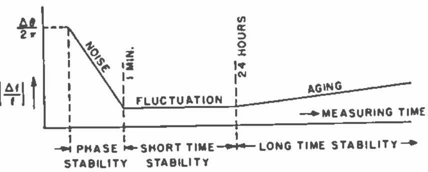

Fig. 1

1) *Phase stability:* The frequency stability measured in a very short time interval depends mainly upon the phase instabilities of the oscillator. The fluctuation of the operating parameters and the aging is small compared with this effect. The measuring time starts with one period (see Fig. 2). Because of the phase instability $\Delta\theta$, the measurement has an uncertainty, often called short term stability.

$$\frac{\Delta f}{f} = \frac{\Delta\theta}{2\pi} \cdot \frac{1}{N}, \tag{1}$$

where $\Delta\theta/2\pi = S_\theta =$ "phase stability," and $N =$ number of periods used for the measurement.

* Received by the IRE, June 17, 1960; revised manuscript received, September 28, 1960.

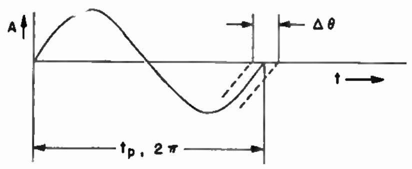

Fig. 2

2) *Short-time stability:* For a measuring time longer than approximately one minute, the fluctuations of the operating parameters are the main cause of instability.

3) *Long-time stability:* For a measuring time greater than 24 hours, the aging of the frequency-determining elements is responsible for the variation in frequency.

Measuring Systems

After a short discussion of theory, two systems using a period measurement and one system using a frequency discriminator are described.

A. Theory

The definition of frequency is given as the number of cycles divided by the time:

$$f = \frac{N}{t_m}. \tag{2}$$

We have only to measure the time t_m required for N periods to determine the frequency. Because of the phase jitter, the time t_m has an uncertainty,

$$\Delta t_m = \frac{\Delta\theta}{2\pi} \cdot t_f. \tag{3}$$

Here t_f is the period of the measured frequency.

This time uncertainty can be expressed as a frequency uncertainty,

$$\Delta f = \frac{df}{dt_m} \Delta t_m = -\frac{N}{t_m^2} \Delta t_m. \tag{4}$$

Assuming that Δt_m is only caused by the phase jitter, we combine (2), (3), and (4) and have

$$\frac{\Delta f}{f} = -\frac{\Delta\theta}{2\pi}\frac{t_f}{t_m} = -\frac{\Delta\theta}{2\pi}\frac{1}{t_m \cdot f}. \tag{5}$$

Interpreting (5) we can say that the uncertainty of a frequency measurement is given by the phase instability of the signal source, the measuring time t_m, and the signal frequency f. The phase stability

$$S_\theta = \frac{\Delta\theta}{2\pi}, \tag{6}$$

being independent of measuring time and frequency, is a figure of merit for an oscillator.

B. Conversion System

We often measure the frequency stability of an oscillator with a system of the type shown in Fig. 3. The frequency f_1 with the phase excursion $\Delta\theta_1$ of the oscillator under test is multiplied to nf_1 and the phase excursion to $n \cdot \Delta\theta_1$. This frequency is mixed down to f_D with the remaining phase excursion $\Delta\theta_1$. The phase stability is given by

$$S_\theta = \frac{\Delta\theta_1}{2\pi} = \frac{\Delta f_1}{f_1} \cdot t_m \cdot f_1 = \frac{f_D}{n} \Delta t_m. \tag{7}$$

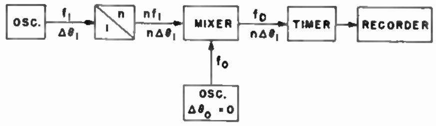

Fig. 3

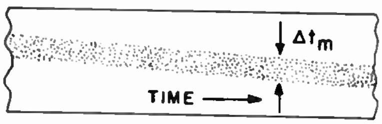

Fig. 4

Fig. 4 is a schematic plot of a stability recording. The resolution of this sytem depends upon the time unit used to measure Δt_m and the difference frequency at the signal frequency $\Delta f = f_D/n$. The time needed for the measurement is proportional to $1/n$ for the same resolution.

With a time unit of 10^{-7} seconds and $f_D = 10$ cps, $n=2$; the resolution is $5 \cdot 10^{-7}$. The advantage of this system is the high resolution. The disadvantage is that the phase stability of the second signal source has to be much better than that of the tested oscillator.

C. Direct System

Measuring the phase stability directly by means of a period measurement with a timer is useful only at audio frequencies. In this case, the phase stability is given by

$$S_\theta = f_{\text{signal}} \cdot \Delta t_m. \tag{8}$$

With the time unit of 10^{-7} seconds, we have a resolution of 10^{-1} at a signal frequency of 1 Mc and of 10^{-4} at 1 kc.

D. Discriminator System

Another system to measure the phase stability is outlined in Fig. 5. A phase modulated signal with a peak-to-peak deviation $\Delta\theta_1$ and a modulation frequency f_m has a peak-to-peak frequency deviation of

$$\Delta f = \frac{\Delta\theta}{2\pi} \cdot f_m. \tag{9}$$

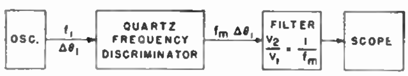

Fig. 5

The phase stability of a crystal oscillator is in the order of 10^{-4}. The frequency range of f_m is from 1 to 1000 cps. With this value we find that our discriminator has to detect frequency deviations from

$$\Delta f = 10^{-4} \text{ to } 10^{-1} \text{ cps}. \tag{10}$$

Relating this frequency deviation to a discriminator frequency of 10 Mc, the quartz crystal frequency, gives us a relative deviation of 10^{-11} to 10^{-8}. The stability of the crystal representing the discriminator has to be better than this deviation. Instabilities caused by the active components of the oscillator network are now omitted.

Results

Evaluations of measurements (1), (2), and (3) on high stability oscillators showed an average phase stability of

$0.5 \cdot 10^{-4}$ for 0.1 Mc crystals,
$1 \cdot 10^{-4}$ for 1 Mc crystals, and
$5 \cdot 10^{-4}$ for 5 Mc crystals.

The phase stability is apparently related to the crystal bandwidth.

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[3] K. H. Sann, Diamond Ordnance Fuze Labs., unpublished data.

Failure Character of Receiving Tubes Due to Deterioration*

Failures of electronic receiving tubes, like the failure of other electronic components, are classified into two categories: chance failures and wear-out failures. The lifetime distribution of chance failures can be expressed by a logarithmic function. Although the Weibull[1] and other distribution functions have been proposed and do agree closely with the wear-out failures, these functions do not explain why they agree with the wear-out characters of electron tubes.

There exists one method of preventive maintenance which we have been using to check the mutual conductance (G_m) and the plate current (I_p) of all tubes after certain fixed interval of operation and for replacing all tubes which fail under the specifications. However, we are faced with the problems of deciding how to select the interval and what standard we should take to discard the tubes.

Furthermore, it is necessary to compare the advantages of the above-mentioned method of maintenance with the others, such as making repairs only at a failure occurrence or replacing all tubes after a fixed period of operations.

In order to solve such problems we must know not only the failure distribution of the tubes, but also both the reason why this distribution appears and how the failure is characteristic of the tubes.

The following assumptions about the wear-out characteristics of receiving tubes will be made:

* Received by the IRE, July 5, 1960.
[1] G. C. Dalman and J. H. Kao, "Determination of vaccum-tube catastrophic and wear-out failure properties from life-testing data," *Proc. 4th Natl. Convention on Tube Techniques*, pp. 245–261; September, 1958.

1) The failure characteristic curve is convex and decreases monotonously.
2) The distribution of the initial mutual conductances of the tubes is normal.
3) The distribution of one of the parameters of the curve is normal.

Let us consider the following equation which agrees with the assumptions stated above.

$$Y(t) = \frac{2\alpha}{\pi} \tan^{-1} \frac{t}{t_0} \qquad (1)$$

where $Y_{(t)}$ is the percentage decrease of G_m

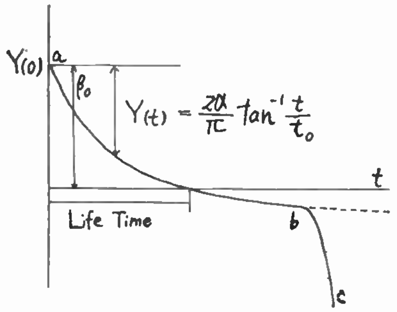

Fig. 1

(Fig. 1). If the distribution of $Y_{(0)}$ and α of (1) are normal, and their standard deviations are σ_1 and σ_2 respectively, we can obtain the following failure density distribution function:

$$f(t) = \frac{1}{\sqrt{2\pi}\sqrt{\sigma_1^2 + \sigma_{2t}^2}} \exp\left\{-\frac{2\alpha_0}{\pi^2(\sigma_1^2+\sigma_{2t}^2)} \cdot \left(\tan^{-1}\frac{t}{t_0} - \tan^{-1}\frac{T}{t_0}\right)^2\right\}. \qquad (2)$$

where

$$\sigma_{2t} = \frac{2\sigma_2}{\pi} \tan^{-1} \frac{t_0}{t}$$

$$\tan^{-1} \frac{T}{t_0} = \frac{\pi\beta_0}{2\alpha_0} \stackrel{\Delta}{=} \theta_0,$$

β_0 is the value of G_m by which we decide whether the tubes are to be discarded or not.

Curves plotted from data obtained from some TV broadcast stations are shown in Figs. 2 and 3. From these curves we can see that (1) gives a first order approximation of the wear-out characteristic of a receiving tube.

However, since the values of (1) converge to a finite limit as t becomes infinite, $Y_{(t)}$ does not fall below $Y = \beta_0$ for some values of α and $Y_{(0)}$. In these cases we obtain the result, which is inconsistent with experience, that some electron tubes have an infinite life time. But we can explain this inconsistency as follows: because (1) closely agrees with the data, "Eq. (1) is realised in any values of time $t(t<\tau)$, and hereafter quickly runs down as depicted by the line segment b–c. The time represented by point b is essentially random." (See Fig. 1.)

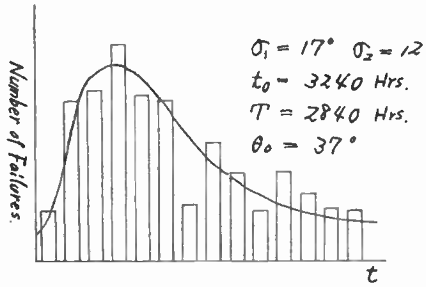

Fig. 2

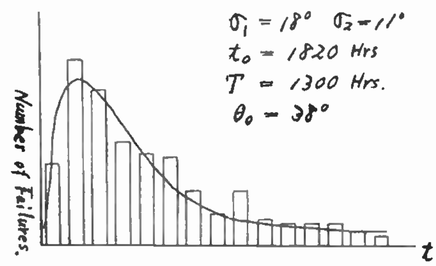

Fig. 3

In (2), σ_1, σ_2 and α_0 are invariable inherent for electron tubes and θ_0 is a variable which varies proportional to β_0.

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History of the Cross Antenna*

The advent of shortwave international communication made really directive radio circuits feasible. Foster[1] mathematically examined a variety of wire configurations for these purposes. One of the exercises involved two lines, each of 16 dipoles in form of a cross. The resultant computed polar pattern is shown in Fig. 1 which is a reproduction of his Fig. 12. This figure is for both lines of dipoles in phase. If the phase of one line is reversed, the instantaneous polarity of one coordinate reverses, canceling the center lobe. All the side lobe structure remains with merely a change of polarity. Thus, as the two lines of dipoles are phase-reverse switched, the center lobe appears and disappears as if by magic at the switching rate, while the side-lobe structure remains fixed. Consequently, the energy present at the switching frequency will be proportional only to the energy accepted by the solid angle of the center lobe. Foster does not mention the above switching technique, nor was any such array constructed.

Southworth,[2] apparently unaware of the above, independently considered the same configuration "about 1941" for aircraft an-

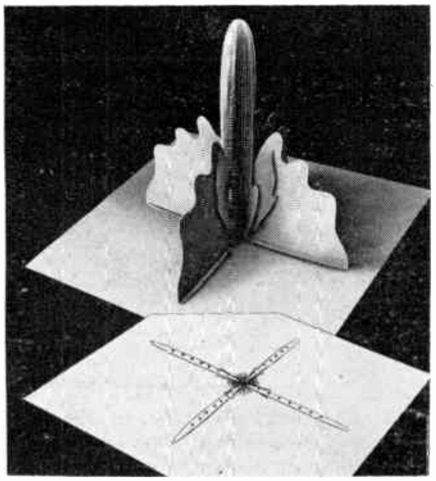

Fig. 1—Model of the space characteristic for an array of 32 antennas located along the diagonals of a square, with a separation of one-half wave length between adjacent antennas in each diagonal, and with zero phase difference.

tennas with low head resistance. In this case, the arms of "my cross consisted of two perpendicular Vees (Hansen horns) made up of leaky-pipe radiators." His method of connecting the arms to "a circular waveguide carrying a circularly-polarized dominant wave of a frequency of 3000 Mc" would produce circularly polarized waves in space. The above reverse-phase switching concept seems to have been overlooked. Notes and sketches were set down but nothing was built.

Ryle[3] writes that "by using aerials of equal area, one of which has a large aperture in the north-south plane, whilst the other has a large aperture in the east-west plane, it becomes possible to reduce the effective solid angle for reception without increasing the total area of the system." The main subject of this paper is a reverse-phase switching interferometer. Obviously, the potential ability of the system is understood. Unfortunately, it seems that no mathematical study was undertaken and again nothing was built

The modern effective use of the cross antenna has its origin in the independent inspiration of Mills,[4,5] who first built and demonstrated the important advantages of this system to radio astronomy. Essentially, the system is a method of exchanging excess sensitivity for improved angular resolution. At wavelengths longer than a meter, where the celestial objects are of sufficient intensity, this antenna system has no rivals. Several very large ones are now under construction to assist in the study of our sidereal universe. If any reader knows of other early examples of this art, I'll be pleased to learn the details. The optical analog is crossed slits. Does anyone have a reference?

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New York 17, N. Y.

* Received by the IRE, September 26, 1960.

[1] R. M. Foster, "Directive diagrams of antenna arrays," *Bell Sys. Tech. J.*, vol. 5, pp. 292–307; April, 1926.

[2] G. C. Southworth, private correspondence, 1956, 1957, 1960.

[3] M. Ryle, "A new radio interferometer," *Proc. Royal Soc. (London) A*, vol. 211, pp. 351–375; March 6, 1952.

[4] B. Y. Mills and A. G. Little, "A high resolution aerial system of a new type," *Australian J. Phys.*, vol. 6, pp. 272–278; September, 1953.

[5] B. Y. Mills, *et al.*, "A high resolution radio telescope for use at 3.5 meters," Proc. IRE, vol. 46, pp. 67–84; January, 1958.

A Note on the Relationship Between the Geomagnetic and the Geographic Axes*

It is known[1] that, to a very rough approximation, the magnetic field around the earth is similar to one that would be produced by a magnetic dipole near the center of the earth or by a uniformly magnetized sphere coinciding approximately with the earth's surface. The position of the magnetic poles with respect to the apparent geographic poles is:

North Magnetic Pole
70°5′ North Latitude 96°46′ West Longitude
South Magnetic Pole
72°46′ South Latitude 155°16′ East Longitude.

It is also well known[2] that the direction of the earth's axis of rotation in space (*i.e.*, geographic axis) undergoes a great, though slowly proceeding, periodic change. The position of the north or south geographic pole on the celestial sphere moves slowly around in a circle, the angular diameter of which subtended at the earth's center is 46°54′16″, and completes a circuit in 25,800 years. This conical motion of the axis is due to gravitational action of the sun and of the moon on that belt of matter around the earth's equator.

* Received by the IRE, October 4, 1960.
[1] G. P. Harnwell, "Principles of Electricity and Electromagnetism," McGraw-Hill Book Co., Inc., New York, N. Y., 380 pp.; 1938.
[2] A. Gray, "Gyrostatics and Rotational Motion," The Macmillan Co., New York, N. Y., pp. 11–12; 1918.

The similarity[3] of the earth's magnetic field to the field of a sphere or a centered dipole is usually assumed to imply that the cause of the field must be planetary, affecting the whole earth, and leaving only the superposed irregularities to be explained as regional effects. Less unanimity exists as to whether or not the obliquity[3] of 11.5° (a difference when compared to Harnwell,[1] and supported by other authorities[4]) between the earth's axis of rotation (*i.e.*, geographic axis) and the geomagnetic axis is of fundamental significance. Many authors[3] consider that the angle is sufficiently small to warrant the assumption of a physical connection between the rotation and the magnetization of the earth.

The aim of this note is to suggest that there is a definite relationship between the geomagnetic and the geographic axes. Because of the precession of the earth's axis of rotation over a circle of a radius of about 23°, subtended at the center of the earth, we see that the geomagnetic axis is within this circle of precession. Assuming that there is some residual induced magnetization in the earth, the magnetic moment of the earth will not be identical in its direction with the axis of rotation. It will be the vectorial sum of all the previous residual induced magnetizations due to the earth's rotation in the previous era, plus the induced magnetization, due to the rotation at present. This theory will require a liquid core inside the earth, which will be magnetized and will not follow the geographic axis in its precession.

[3] S. Chapman and J. Bartels, "Geomagnetism," Oxford University Press, Oxford, Eng., vol. 2, pp. 701–710; 1940.
[4] M. G. Morgan, "The nature of the ionosphere—an IGY objective," Proc. IRE, vol. 47, pp. 131–132; February, 1959.

The fact that the geomagnetic axis is within the circle of precession of the axis of rotation of the earth shows that:

1) The magnetic moment of the earth is due to the earth's rotation around its axis now and in the previous era.
2) There is a residual induced magnetization in the earth: It is in a liquid core which does not follow the axis of rotation.

This might explain the discrepancy of factor of 10^8 or 10^{10} in previous theories[3] of gyromagnetic effect and rotating electric charges. This might also explain the fact that the geomagnetic axis does not pass through the center of the earth, since the residual magnetism does not have to be of equal value and distribution in the north and the south hemispheres.

It has been pointed out to the author by Dr. Gravitt, Midwest Research Institute, that an additional requirement of the above theory is that the decay time of the residual induced magnetization should *not* be small compared with the period of precession of the axis of rotation.

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Contributors

Semi J. Begun (SM'46–F'52) was born on December 2, 1905, in Danzig, Germany. He received the M.S. degree in 1929 and the Ph.D. degree in 1933, both from the Institute of Technology, Berlin.


S. J. Begun

He came to the United States in 1935 and joined the Guided Radio Corporation, where he worked on the circuit design of seaborne equipment for safe, ship-wide transmission of messages from the ship's bridge. He has also done a great deal of work on telephone circuits and problems of electroacoustics. His research contributions to the original Brush Development Company, now Brush Instruments, Division of Clevite Corporation, Cleveland, Ohio, helped in the production of an inexpensive, high-quality phonograph pickup with a permanent point stylus and many magnetic recording instruments. He also participated in the first U. S. development work attempted on powdered magnetic recording media. He is presently Director of Marketing for the Clevite Corporation, and after more than 25 years of experience in the field of magnetic recording, holds approximately 50 U. S. patents in this and related fields. His wartime contributions on special devices earned him a Presidential Certificate of Merit.

Dr. Begun is a Fellow of the Acoustical Society of America. He has published many papers on magnetic recording and a book entitled "Magnetic Recording."

❖

David I. Breitzer (A'48–M'55) was born in New York, N. Y., on January 11, 1917. He received the B.E.E. and the M.E.E. degrees from the College of the City of New York, N. Y., in 1938.

In 1939, he was employed as a junior naval architect in the U. S. Navy Department. From 1939 to 1941, he was employed by the New York City Board of Transportation as a junior electrical engineer. From 1941 to 1945, he worked on the Panama Canal as an electrical engineer. In 1946 and 1947, he was employed by the Celanese Corporation of America as an electrical designer. In 1947 he joined Airborne Instruments Laboratory, Inc., Mineola, N. Y., and has since been engaged in the mathematical analysis of systems and problems connected with radar and guided missiles, which includes the evaluation of signal-to-noise ratios, the solution of electromagnetic field boundary value problems, the analysis of transients in lumped and distributed circuits network analysis, and the solution of varied types of stochastic problems. In the field of radiation detection, he has analyzed the operating characteristics of a microwave ionization chamber, and has contributed to the solution of several low-noise circuit problems.


D. I. Breitzer

Jack E. Bridges (A'50–M'55–SM'56) was born on January 6, 1925, in Denver, Colo. He received the B.S.E.E. degree with honors from the University of Colorado, Boulder, in 1945. After completion of active duty with the U. S. Navy, he obtained the M.S. degree from the University of Colorado in 1947.

J. E. Bridges

From 1947 to 1948, he was an instructor of electrical engineering at Iowa State College, Ames, and from there he went on to the Zenith Radio Corporation, Chicago, Ill., where he was a research engineer until 1955. From 1955 to 1956 he was engaged as head of the color television engineering section at Magnavox Corporation, Fort Wayne, Ind., and since 1956 he has been the chief electrical engineer in the research department of Warwick Manufacturing Corporation, Chicago, Ill.

Mr. Bridges is a member of Tau Beta Pi and Eta Kappa Nu.

❖

D. E. Edmondson was born in Dallas, Tex., on September 6, 1925. He received the B.S. degree in mechanical engineering and the M.S. degree in mathematics from Southern Methodist University, Dallas, in 1945 and 1948, respectively. After graduate work at the University of Chicago in 1949 and 1950, he taught at S.M.U. and in 1951, he received a fellowship to the California Institute of Technology, Pasadena, where he received the Ph.D. degree in mathematics in 1954.

D. E. Edmondson

In 1954 he joined Tulane University as a research instructor where he did research on abstract and topological lattices. In 1955, he returned again to S.M.U., where he became an associate professor in 1958. While at S.M.U., he was active in the graduate mathematics program, served in several educational experiment programs, and conducted research in pure and applied mathematics as a consultant with Texas Instruments, Dallas. Presently, he is an associate professor of mathematics at The University of Texas, Austin.

Dr. Edmondson is a member of Sigma Tau, Sigma Xi, the Mathematical Association of America, American Mathematical Society, and S.I.A.M.

❖

Martin M. Freundlich (A'38–SM'45) was born in Goerlitz, Germany, on November 23, 1905. He received the Dipl. Eng. and the Dr.Eng. degree in 1929 and 1933, respectively, from the Technical University of Berlin, Berlin, Germany.

M. M. Freundlich

From 1929–1934, he worked with Prof. A. Matthias and Dr. M. Knoll at the High Tension Laboratory of the Technical University of Berlin, engaged in research on demountable high-speed cathode-ray oscillographs and on circuits for recording high-speed phenomena. In 1935, he acted as research assistant to Prof. S. Parker-Smith of the Royal Technical Institute, Glasgow, Scotland. He was associated with Pye Radio Ltd., Cambridge, England, conducting research on cathode-ray tubes for television from 1935–1936. From 1936–1944 and 1945–1949, he was in charge of tube research for the Columbia Broadcasting System, New York, N. Y. He built cathode-ray tubes for early black and white and color television experiments and conducted research on various electronic color television systems. During 1944 and 1945, he was with North American Philips Company, Dobbs Ferry, N. Y., developing cathode-ray tubes for radar and television. He joined Airborne Instruments Laboratory, Deer Park, N. Y., in 1949, where he has worked on storage tubes and microwave triodes, among various other problems. He is presently in charge of the Space Research Laboratory of Space Technology and Research Department of AIL, where he studies the behavior of materials under space conditions and investigates lubrication in space.

Dr. Freundlich is a member of the American Physical Society, the American Vacuum Society, and the International Organization for Vacuum Science and Technology.

❖

John B. Gehman (M'45) was born on July 29, 1916, in Pa. He received the B.S.E.E. degree from Villanova University, Villanova, Pa., in 1939.

J. B. Gehman

He joined Radio Corporation of America, in 1939, doing work in the Sound Engineering, Advanced Development Engineering and Aviation Electronic Engineering Departments. He was a group leader in the Advanced Development and Aviation Sections. During World War II, he was responsible for the development of a nonmetallic land mine detector. He left RCA in 1954 to join Clevite Research Center, Cleveland, Ohio, as head of the Advanced Development Department doing work on semiconductor devices and devices utilizing electromagnetic techniques. He then became Head of the Electronics Department. He joined Thompson Ramo Wooldridge (TAPCO Group), Cleveland, in 1958, heading Advanced Electronic Development in the fields of control circuitry, RF video telemetry links, and electronic display devices.

❖

J. Kenneth Lewis (A'52–M'47–SM'58) was born on December 29, 1920, in Washington, D. C. He received the B.S.E.E. degree from Kansas State University, Manhattan, in 1942.

J. K. Lewis

He was employed from 1942 to 1952 with the Acoustical Branch, Bureau of Ships, Navy Department, Washington, D. C., where he designed various audio communications systems for Naval vessels and operations, including loudspeakers and intercommunications system; shipboard motion picture equipment; portable high-powered systems for amphibious operations; submarine communications systems; and sound recording and reproducing systems. From 1952 to the present, he has been employed with the Department of Defense as Chief of the Research and Development Division which is working on data storage systems.

He established the first government specifications defining characteristics of magnetic tape, and is currently designing mechanized tape test systems.

Mr. Lewis is a member of the SMPTE and is a Registered Professional Electrical and Electronic Engineer in the District of Columbia.

❖

Chesley H. Looney, Jr. (S'50–A'51–M'55–SM'60) was born in Newburg, Mo., on April 7, 1928. He received the B.S.E.E. degree from the University of Kansas, Lawrence, in 1950, and did graduate work at the University of Maryland, College Park.

C. H. Looney, Jr.

During 1950, Mr. Looney was employed by the Midwest Engineering and Development Co., Kansas City, Mo., as an electronics engineer. In 1951 he joined the scientific staff of the Naval Research Laboratory, Washington, D. C., where he worked in the fields of automatic control, analog computation, and electronic pulse circuitry, and where, in 1956, he joined Project Vanguard and was transferred to the National Aeronautics and Space Administration (NASA), with Project Vanguard, in 1958. He has been responsible for satellite tracking instrumentation design, especially in the fields of data handling, phase angle measurement, and time standards. He is Head of the Time Measurement Branch of Goddard Space Flight Center, NASA, Washington, D. C.

Mr. Looney is a member of Tau Beta Pi.

❖

J. Ross Macdonald (S'44–A'48–SM'54–F'59) was born in Savannah, Ga., on February 27, 1923. He received the B.A. degree in physics from Williams College, Williamstown, Mass., and the B.S. and S.M. degrees in electrical engineering from the Massachusetts Institute of Technology, Cambridge, in 1944 and 1947, respectively. He was awarded a Rhodes Scholarship from Massachusetts for study at Oxford University, England, where in 1950 he received the D. Phil. degree in physics for theoretical and experimental work in ferromagnetic phenomena.


J. R. Macdonald

After war service he returned to M.I.T. for further study and to carry out research on storage tubes for the M.I.T. digital computer project.

In 1950, he joined the Physics Department of the Armour Research Foundation and there carried out and directed work in theoretical and experimental physics until 1952. He then spent a year's leave of absence at the Argonne National Laboratory of the A.E.C. working on solid-state physics problems. He is presently Director of the Solid State Physics Research Department at Texas Instruments Incorporated, Dallas. In addition, he is also serving as clinical associate professor of medical electronics at Southwestern Medical School of the University of Texas, Dallas.

Dr. Macdonald is a member of Phi Beta Kappa and Sigma Xi, and is a Fellow of the American Physical Society and of the AAAS.

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Sanford A. Meltzer was born on October 22, 1930, in Syracuse, N. Y. He received the A.B. and M.S. degrees in physics from Syracuse University in 1950 and 1952, respectively.

Following his graduation he joined the General Electric Company's Physics Program. From 1953 to 1955 he was a member of the Communication Theory Group of the General Electric Advanced Electronics Center at Cornell University, Ithaca, N. Y. From 1955 to 1957 he was associated with the Advanced Development Section of Radio Corporation of America, Camden, N. J. In August, 1957 he joined the technical staff of the Hughes Aircraft Company, Culver City, Calif., where he currently holds the position of head of the Signal Processing and Performance Studies Group in the Advanced Development Laboratory.


S. A. Meltzer

Mr. Meltzer is a member of the American Physical Society, Sigma Xi, RESA, Sigma Pi Sigma, and Pi Mu Epsilon.

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William W. Mumford (A'30–SM'46–F'52), for a photograph and biography, please see page 959 of the May, 1960, issue of PROCEEDINGS.

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Marshall C. Pease III (M'47–SM'51) was born on July 30, 1920, in New York, N. Y. He received the B.S. degree in chemistry from Yale University, New Haven, Conn., in 1940, and the M.A. degree in physical chemistry from Princeton University, Princeton, N. J., in 1943. From 1943 to 1945 he was a research associate at the Radio Research Laboratory at Harvard University, Cambridge, Mass., working in the field of microwave tube applications and testing. He attended Pre-Radar School at Harvard in 1943. In 1945 he entered the U. S. Navy with a prior commission as Ensign. He attended Radar School at Massachusetts Institute of Technology, Cambridge, and served on active duty until 1947.


M. C. Pease

From 1947 to April, 1960, when he joined the staff of Stanford Research Institute, Menlo Park, Calif., he was employed by Sylvania Electric Products Co. in Woburn, Mass., and Mountain View, Calif., as a tube engineer and in various supervisory and managerial capacities in connection with the engineering and advanced development of microwave and special-purpose tubes. His final task was as senior engineering specialist, doing theoretical investigation on special problems in electron dynamics and parametric circuit theory. At the Institute, he is working on an analysis of electron-beam parameter amplifiers and also assisting on a project aimed at the development of new computer logic devices. He has published articles in technical journals in the fields of microwave circuit theory and of electron flow theory and holds several patents in the field of microwave electron devices.

Mr. Pease is a member of Phi Beta Kappa, the Scientific Research Society of America, and the AAAS. He is an Associate Editor of the *Microwave Journal*.

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Jerome J. Suran (A'52–SM'55) was born in New York, N. Y., on January 11, 1926. After having served for three years with the U. S. Army during World War II, he received the B.S.E.E. degree from Columbia University, New York, N. Y., in 1949, and continued graduate studies there and at the Illinois Institute of Technology, Chicago, Ill.


J. J. Suran

From 1949 to 1952 he was employed in the field of control systems design and development by J. W. Meaker and Co., New York, N. Y., and in the field of FM research and development by Motorola, Inc., Chicago, Ill. Since 1952 he has been active in the area of solid-state circuits, and is currently manager of the Advanced Circuits Component of the Electronics Laboratory, General Electric Company, Syracuse, N. Y.

Mr. Suran is co-author of "Principles of Transistor Circuits" and "Transistor Circuit Engineering." He is a member of the AIEE and the Research Society of America.

❖

Samuel Thaler (A'56) was born on June 17, 1917, in Brooklyn, N. Y. He received the B.E.E. degree from the College of the City of New York in 1940.


S. Thaler

From 1940 to 1956 he worked for agencies of the United States government on a variety of projects. In 1956 he became a member of the technical staff of the Hughes Aircraft Company, Culver City, Calif., where he was engaged in design and performance studies of Doppler radar systems. In March, 1960, he joined the West Coast Missile and Surface Radar Division of RCA, Van Nuys, Calif., where he is now a principal member of the systems engineering staff.

Mr. Thaler is a member of the AIEE.

Books

Self-Organizing Systems, Marshall C. Yovits and Scott Cameron, Eds.

Published (1960) by Pergamon Press, Inc., 122 E. 55 St., N. Y. 22, N. Y. 322 pages+ix pages+bibliography by chapter. Illus. 5⅜×8⅜. $8.50.

Consisting, as it does, of the proceedings of a conference attended by leading scientists in the field of self-adaptive systems, this book necessarily contains a great deal of stimulating material. There is an implicit assumption, however, in many of the articles that the reader is thoroughly acquainted with the background material and that, hence, preliminary explanations and definitions are not required. Written from this viewpoint, the book retains as its audience the small group of scientists for whose benefit the conference was originally intended. It loses or unnecessarily handicaps the much larger potential audience of scientists and engineers who might have found the book a useful guide into the deeper concepts of an important new field.

F. J. Weyl, in the opening talk of the conference, points out that the concept of organization emerging is a communication and information theoretical one. The first four papers, under the topic of "Perception of the Environment," give general evidence of the validity of Weyl's observation without developing the point explicitly.

Rosenblatt's paper on the generalization capabilities of the cross-coupled Perceptron is significant, reflecting as it does the recent rigorous mathematical study with which Rosenblatt has been involved at Cornell University. In the same series of papers, von Foerster discusses self-organizing systems from the thermodynamics viewpoint and develops briefly a topic which will undoubtedly be the subject of continued investigation. In particular, it is hoped by this reviewer that work along the lines indicated by von Foerster will provide some criterion of the abstraction process that will avoid equating it simply to the discarding of information.

The second group, entitled "Effects of Environmental Feedback," consists of three papers by Auerbach, Goldman, and Bishop, respectively. They demonstrate very clearly the value of the application, to the study of biological systems, of knowledge derived from information and communication theory. Auerbach points out the application of this knowledge to the embryonic system in process of organization. Goldman emphasizes the servo control aspects of homeostasis. Bishop stresses the importance of feedback processes in brain functions.

Newell, Shaw, and Simon discuss the possibility of learning in the general problem solver in the first paper of the third group. The group title is "Learning in Finite Automata" and includes, in addition to the foregoing paper, a paper each by psychologists Milner and Campbell. The three papers are related only in a very general way.

The last group, "Structure of Self-Organizing Systems," includes papers by Pask, McCulloch, and Burks, respectively. Pask discusses the problem of analyzing and controlling self-organizing systems. McCulloch reports on his recent work on infallible networks of fallible neurons and probabilistic logic. His paper is concise and a reading of some of the reference literature is required to appreciate it, for McCulloch, of course, is a leader in the field with a long and continuous history of significant contributions. Burks concerns himself with deterministic automata and considers Turing machines and self-reproducing machines. It is clear that a complete theory of deterministic machines necessarily illuminates the subject of self-organizing machines as well.

Most, but not all, of the papers have been mentioned in the preceding paragraphs. The depth and breadth of the papers indicate how very difficult it might be for any editor to provide interpretive links between the various sections of the text. It may well be that the field is not yet ready for tutorial or interpretive generalizations except on restricted topics. The inclusion of the discussions held at the conference is of some help.

The book is recommended to the professional scientist with some background in the field.

Note: Those readers particularly interested in the Rosenblatt paper will note that the third rule on page 75 should read as follows:

> "All connection weights W_{iu} which originate from a unit a_i which was active at time $(t-1)$ gain an increment Δw."

There are, in addition, a number of typographical errors which should, however, cause the serious reader no difficulty.

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Wave Generation and Shaping, by Leonard Strauss

Published (1960) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 506 pages+9 index pages+xvi pages+bibliography by chapter. Illus. 6¼×9¼. $12.50.

Professor Strauss has written an excellent book on wave generation and shaping. The principal feature of the book is the unified approach to practically all wave shaping circuits obtained by emphasizing a piecewise linear representation combined with break point analysis. The book is practical in the sense that actual circuit design is the goal, along with a thorough understanding of the operation of the various circuits covered. This book is not a reference book but is intended to be studied from the beginning, since the analysis of later circuits depends heavily on a thorough understanding of the method. However, once the method is mastered, the analysis of a large class of circuits becomes relatively simple.

The book is divided into five sections to emphasize the unity of the ideas presented. The first section is the basic and most important section and covers diode wave shaping techniques, diode gates, and triode, transistor, and pentode models and circuits. The section section is devoted to timing and contains chapters on the principles of sweeps, vacuum tube voltage sweeps, transistor sweeps, and linear current sweeps. The third section is on switching and contains two chapters on multivibrators, an excellent chapter on negative-resistance switching circuits, and a chapter on the blocking oscillator. The fourth section is a brief introductory chapter on magnetic and dielectric devices as memory and switching elements. The fifth and last section is devoted to oscillations and contains a chapter on almost sinusoidal oscillations and a chapter on negative-resistance oscillations.

Although the book is intended as a text for seniors and first-year graduate students, the very clear exposition is well suited to engineers who find themselves in need of additional education in the topics covered. Some of the chapters, such as the one on memory, are clearly introductory, and if the book has a defect, it is in the somewhat superficial coverage in this chapter. But the large body of the information is well chosen, clearly presented, and adequately covered. The illustrations are well drawn. I recommend this book as a worthy addition to an engineer's bookshelf—particularly if wave shaping and generation circuits are not his specialty—and as a text as Professor Strauss has intended.

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Electronic Maintainability, Vol. 3, F. L. Ankenbrandt, Ed.

Published (1960) by Engineering Publishers, Division of the AC Book Co., Inc., Elizabeth, N. J. 312 pages+viii pages. Illus. 6×9. $10.00.

This book reports the Third Electronic Industries Association Conference on Maintainability of Electronic Equipment, where the modern aspects of this important problem were discussed in their many phases, both theoretical and practical. Although the conference was concerned mainly with military equipment, solutions to maintenance problems occasioned by large computer installations and by airline equipment complexities provided additional contributions. It becomes evident that designers must realize that improved maintainability can often be improved with little or no added cost by providing accessibility, integral test aids, simpler servicing instructions. Then the estimated price of keeping electronic systems operating, which will often total several times the original installation, will come down. Rules for setting up service routines, check-outs, test facilities, spare parts availability and training programs are among the discussions presented. This volume is a useful reference in this important field.

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An Introduction to Transistor Circuits, by E. H. Cooke-Yarborough

Published (1960) by Interscience Publishers, Inc., 250 Fifth Ave., N. Y. 1, N. Y. 150 pages+6 index pages+xii pages+bibliography by chapter. Illus. 5⅜×9. $3.50.

This monograph is based on a series of lectures given by the author, one of the pioneers of the transistor circuit art in the United Kingdom, in 1955 at University College, London. The author states in the introduction that "a good deal of material has since been added, but it would be idle to pretend that every aspect of transistor circuits has been covered." This is undoubtedly so.

The first chapter presents a qualitative description of physical phenomena contributing to transistor action and of semiconductor device properties in general. Low frequency amplification, behavior of various low frequency amplifier types, and some aspects of high frequency amplification are given in the next two chapters. The remaining half of the volume deals with pulse circuits and their applications, and includes the discussion of a variety of multivibrators and of some counting, specialized switching, logic, and gating circuits. The treatment of the subjects covered is intentionally qualitative rather than quantitative and is colored, in several instances, by excellent "intuitive" explanations of physical phenomena and of circuit behavior.

The book reflects the author's considerable and thorough understanding of transistor behavior. While it should not be considered as representative of today's state of the art, it is still of historic interest and could be interesting reading for those who desire to gain a cursory and general understanding of transistor circuit principles.

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Manual of Mathematical Physics, by Paul I. Richards

Published (1960) by Pergamon Press, Inc., 122 E. 55 St., N. Y. 22, N. Y. 459 pages+25 index pages+xi pages. Illus. 6¾×10. $17.50.

"The Manual of Mathematical Physics," by Paul I. Richards, is divided into two almost equal parts, one covering physics and the other mathematics. Part I deals with mechanics of particles, rigid bodies and continuous media; thermodynamics; electromagnetic theory, including optics, circuit theory and communication theory; relativity, special and general; quantum mechanics, its formulation and methods, important specific results, and high energy quantum mechanics; statistical physics, including simple kinetic theory, thermodynamic equilibrium, and transport phenomena.

Part II covers algebra, differentiation and integration, including Stieltjes integrals, delta functions and Schwartz's theory of distributions; infinite series; vector analysis; determinants and matrices; functions of a complex variable; integral transforms; ordinary differential equations; partial differential equations, including Laplace's, Poisson's, diffusion and wave equations as well as the transport equation; integral equations, of Wiener-Hopf type, Fredholm's, Volterra's and equations of the "first kind"; variational problems; linear programming; unitary spaces; eigenvalue problems; perturbation theory; probability and game theory; tensor analysis; group theory.

The treatment is highly condensed and the reference material is by no means elementary (although a few simple but very important formulas are included). For this reason some readers will find the manual difficult to use. On the other hand the brevity and elimination of material readily available elsewhere enabled the author to present an amazingly large number of more advanced formulas and subjects for the benefit of those who once learned them but because of subsequent specialization need to refresh their memories. There are omissions, of course, but no manual of reasonable size can be complete.

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Analysis and Design of Feedback Control Systems, 2nd ed., by George J. Thaler and Robert G. Brown

Published (1960) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 528 pages+8 index pages+xiii pages+bibliography by chapter+111 appendix pages. Illus. 6¼×9¼. $14.50.

These two authors published the first edition of this work in 1953. The appearance of a second edition of any technical textbook testifies not only to the vigor of the authors, but also to the success of the first edition. One may assume without much doubt that the second edition of this particular book will be as successful as the first. It is written chiefly as an introduction to the subject of feedback control systems. The treatment is not excessively sophisticated, but is very thorough, and without doubt should be satisfactory for a good percentage of the schools in the country which present one or two courses in servomechanisms or feedback control systems. While the authors have introduced sampled-data systems and phase plane analysis, they have decided not to discuss probability theory or random processes; thus no statistical theory is presented.

The new edition has eliminated the word "servomechanism" from the title. There has been a sufficient revision of the work to warrant the change. The first two chapters, which introduce feedback systems and give a cursory inspection of the Laplace Transform technique of solving linear differential equations, have been revised only slightly. The third chapter, which describes equations of mechanical and electrical systems, with a slight bit of hydraulic and thermal systems material as well, has been altered substantially by employing analog-computer representations of the various systems in addition to what was presented previously.

The meat of the course is in Chapters 4–7. The first of these is a conventional, but thorough treatment of a second-order system with various types of control. The second deals with transfer functions and block diagrams. The third introduces the Bode diagram and the root locus. The fourth considers stability on the basis of the Routh and Nyquist criteria and the effect of the various types of loop compensation on the stability. Fig. 7-6 is particularly commendable for a presentation of the characteristics of fifteen common types of open-loop transfer functions with Bode, Nichols, Nyquist and root locus plots for each.

Chapters 8 and 9 deal with design specifications, the compromises with reality which component shortcomings necessitate, and three well-worked-out design problems. The tenth chapter is a departure from the well established ways of books on feedback because it devotes 45 pages to a method published by Mitrovic in 1959 of analyzing the characteristic equation of a closed loop to determine the number of roots which have a damping ratio in excess of a specified value. This method also permits the evaluation of real roots. Only time will tell whether this much space is properly devoted to Mitrovic's method in a book of this type.

The concluding four chapters are all new. They cover sampled-data servosystems, phase plane analysis, describing functions and relay servomechanisms. On the whole they cover the subject adequately and well. There are errors present in these chapters as one usually finds in the first edition of a textbook. For example, Fig. 11-21, which shows the spectrum of a sampled sinusoidal wave, omits half the components. Equation 11-66 is very confusing until one realizes that two separate equations have been erroneously compressed on the page. This is, however, carping about what one should normally expect in newly published material. The book is, as a whole, a work well done.

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Electromagnetic Energy Transmission and Radiation, by Richard B. Adler, Lan Jen Chu, and Robert Fano

Published (1960) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 612 pages+9 index pages+xvii pages. Illus. 6×9¼. $14.50.

As stated by the authors, this book "treats electromagnetic waves and oscillations in one, two and three space dimensions, using time-domain, frequency-domain and energy points of view." The subject matter of the book forms a senior-level course in the "core curriculum" required of all electrical engineering students at the Massachusetts Institute of Technology. When he takes this course, the student is assumed to have had two and a half years of calculus and more than a year of college-level experience with electromagnetic fields, including one term of quasi-statics which is based upon parts of a companion volume, "Electromagnetic Fields, Energy and Forces," by the same authors.

Although the content of the book is fairly conventional, the subject matter being that which is given somewhere in nearly all electrical engineering curricula, the treatment is rather different from that given in most texts dealing with this field. The ten chapters give a quite comprehensive coverage of the following topics: "Lumped-Circuit and Field Concepts," "Quasi-Static Fields and Distributed Circuits," "Steady-State Waves on Lossless Transmission Lines," "Transient Waves on Lossless Transmission Lines," "Traveling Waves on Dissipative Transmission Lines," "Natural Oscillations," "Standing Waves and Resonance," "Plane Waves in Lossless Media," "Plane Waves in Dissipative Media," "Transverse Electromagnetic Waves," and "Elements of Radiation."

Possibly the outstanding characteristic of this text is the authoritative treatment of the subject matter. Written by experts, it devotes unusual attention to discussion of the fine points, and to answering questions which might be raised in the mind of the student. Indeed, this very thoroughness may discourage its use in some quarters as a first text in the subject. To keep the book within bounds for a senior course the authors have wisely put the more elaborate developments in asterisked sections which may be omitted without destroying the continuity of the presentation. A feature of interest to the teacher and serious student is the 262 problems which accompany the text. This wealth of well-thought-out problems points up the extensive use the text has already received in the form of preliminary classroom notes.

"Electromagnetic Energy Transmission and Radiation" is a well-written, authoritative book that should prove stimulating to teacher and student alike.

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Analog and Digital Computer Technology, by Norman R. Scott

Published (1960) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 509 pages+8 index pages+xi pages+4 appendix pages. Illus. 6×9¼. $12.75.

This book is intended as a text for an introductory course for electrical engineers who need a good foundation for more advanced study in the computer field. Generally lucid writing, an excellent bibliography, numerous problems, and a good choice of topics should make the book a joy to use as a classroom text or for home study.

The book is divided into two almost completely independent parts. The first part, dealing with the dc electronic differential analyzer, occupies 160 pages. The remaining 355 pages are devoted to the general purpose digital computer. The portion devoted to the analog computer contains an introductory chapter describing the dc differential analyzer in general terms, followed by chapters discussing techniques of setting up problems, computer representation of nonlinear functions, and computing amplifiers. Similarly, the digital computer is described in general terms in an introductory chapter and subsequent chapters are devoted to problem setup, number systems and codes, mathematical logic and switching networks, arithmetic and control logic, circuit mechanization of logic, and memory techniques. There seems to be no special reason for combining these books under one cover unless it is that there is not enough material for a one-semester three-unit course in analog computers. (There is more than enough material for such a course in digital computers.) The potential user of computers will find little to help him categorize problems as being suitable for one or the other type of computer. On the other hand, the potential designer may get some help in deciding which is his true love.

A feature of the book which appealed particularly to this reviewer is the realistic treatment of the subject matter not always found in elementary texts. For example, considerable space is devoted to the problem of extraneous roots introduced by the analog computer setup, and the practical problems of potentiometer loading, scaling, setting of initial conditions, and amplifier drift are realistically treated. Similar examples from the digital computer portion of the book include discussions of round-off errors, scaling, and circuit design for tolerances in component values and power supply voltages. (It is only fair to point out that some important problems such as diode recovery and leakage are omitted.)

The chapter on programming seems to be well suited to the purpose of the text and includes a section on modern approaches to automatic programming and error location.

In summary, this text should be useful for those interested in acquiring a good background for computer design. One might hope to acquire familiarity with many of the tools of the designer, and limited insight into the design process by conscientious home study of the book.

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Waves in Layered Media, by Leonid M. Brekhovskikh

Published (1960) by Academic Press, Inc., Publishers, 111 Fifth Ave., N. Y. 3, N. Y. 543 pages+8 index pages+xi pages+10 bibliography pages. Illus. 6¼×9¼. $16.00.

This book is succinctly stated by the author to be a systematic exposition of the propagation of elastic and electromagnetic waves in layered media. The term "layered media" is used here for media whose properties vary in only one coordinate direction, and thus covers stratified or step-wise homogeneous media and inhomogeneous media whose properties are continuous functions of position. The work is divided into six chapters: 1) "Plane Waves in Layers," 2) "Some Applications of the Theory of Plane Wave Propagation in Layered Media," 3) "Plane Waves in Layered-Inhomogeneous Media," 4) "Reflection and Refraction of Spherical Waves," 5) "Wave Propagation in Layers," and 6) "The Field of a Concentrated Source in a Layered-Inhomogeneous Medium."

The coverage is as comprehensive in scope as one can profitably deal with in a single volume. The reader is taken from the most elementary problems of reflection and refraction, to be found in any basic text, to more complex and interesting problems, such as interference filters, surface wave phenomena, ducting, normal mode theory for treating propagation in a semi-infinite inhomogeneous media, high frequency asymptotic behavior (ray theory approximations) of solutions for propagation in continuously inhomogeneous media. While treatment is based largely on Russian work, in which the author's own contributions are prominent, it also takes cognizance of contributions to the field from other countries. The bibliography itself is a very significant contribution.

As may be expected of a volume in an Applied Mathematics series, the exposition is replete with a variety of mathematical techniques. The emphasis, however, is not merely on formal solutions but also on the reading out from the solutions their physical content and practical implications. The combined treatment of scalar (acoustical) waves and vector (electromagnetic and elastic) waves is a particular feature of the work. Certain unifying concepts such as wave impedance and impedance boundary conditions are introduced, and by juxtaposition of the material, the reader is presented with a unified picture of certain classes of wave phenomena.

The book is certainly to be recommended to anyone studying problems of wave propagation seriously. It will serve more as a reference work than as a text for the usual classroom type, of course. In view of the current fashions in mathematics, physics, and engineering curricula, it is not likely that interdisciplinary aspects of the book will do more for a given reader than generate a sense of kinship among the curricula. This is not a criticism of the book but of our various curricula.

The text reads very smoothly and both the translator and the translation editor are to be complimented for the job they have done in making the work available to English-speaking workers.

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Coupled Mode and Parametric Electronics, by William H. Louisell

Published (1960) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 226 pages+4 index pages+xv pages+38 appendix pages. Illus. 6×9. $11.50.

As knowledge increases, those of us who have grown up with a particular field of research begin to despair of the newcomer ever being able to assimilate the seemingly immense quantity of facts painfully wrung from many years of experimentation and theorizing. What we tend to forget is the occasional appearance of books like this which, by a process of generalization, simplify and make coherent the bits and pieces which we ourselves have learned unconnected, item by item.

The book treats the fields of microwave electron tubes and parametric amplifiers from the unifying point of view of coupling of modes. It starts by discussing the modes of excitation on simple mechanical and electrical circuits, then treats the coupling of two pendulums and takes up the coupling of propagating modes. All of this will one day be in every undergraduate engineering curriculum. The discussion then turns to the nature of the space charge waves and cyclotron waves which exist on electron beams and how they behave in the presence of a slow wave structure. Here are the coupled mode treatments of the traveling-wave tube, the backward-wave oscillator and the Kompfner dip, which were formerly available only in their original journal article form. Next, the parametric or time-varying coupling of modes is taken up, and again much of the material is gathered together here for the first time. The author then goes on to discuss the physical principles involved in the semiconductor diode and ferrite versions of the parametric amplifier, to present some of the experimental results, and sandwiched in between, to discuss in considerable detail the space charge wave and cyclotron wave parametric amplifiers. The book is aimed at the first year graduate student.

The appearance of this book is timely, for the coupled mode treatment of electron tubes in text form is long overdue and the material on parametric devices is of great current interest. This reviewer has few quarrels with the work, the major ones being those of nomenclature (the naming of a and a^* as separate normal modes and the designation of active and passive coupling) and a disagreement with the statements that the nonlinear capacitance is equivalent to the time-varying capacitance and that noise can not be removed from the slow waves on an electron beam.

Dr. Louisell, who has been intimately concerned with the development of the fields about which he is writing, is to be congratulated on his timely and well-written book. The publishers, however, are to be chastised for the exhorbitant price they have put on this slim volume.

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The Arc Discharge, by H. de B. Knight

Published (1960) by Chapman and Hall, Ltd., 37 Essex St., London W.C. 2, Eng. 407 pages+8 index pages+xix pages+23 appendix pages. Illus. $5\frac{5}{8}\times8\frac{3}{4}$. 63s.

This up-to-date book is concerned with gas discharge tubes, specifically ignitrons, thyratrons, and excitrons, for the control of electrical power. The book is intended primarily for application engineers and users of arc discharge tubes in power circuits. It appears to this writer that the author has accomplished his task very well. Students of plasma physics should find this book invaluable as an introduction to gas discharge tubes, their capabilities and limitations, and the circuits in which these tubes can be used. Mr. H. de B. Knight has spent his industrial career specializing in the design of these tubes and their application in power circuits.

Part I of this interesting book gives the principles and general characteristics of arc discharge tubes. While this part of the book has limited interest to the experienced engineer who has used gas tubes in power circuits, it will be of really great benefit to the novice in this field to acquaint him with the thyratron, ignitron and excitron tubes.

In Part II, (Chapters 4–6) the author discusses the physical processes in the arc. Here the author brings together very effectively various arc processes and relates them to the arc discharge. Gas clean up, sputtering, arc-back, deionization, arc drop, cathode spots are a few of the subjects covered in this section.

Part III, entitled "The Valves in Use," which consists of Chapters 7–10, explains the ratings, circuits and maintenance of the tubes. It will quickly be recognized to represent an important contribution to the literature. Chapter 7 considers the arc discharge as a circuit component. Special emphasis is given to the conditions various circuits place on the tubes. Many circuit illustrations are presented and evaluated with respect to transient voltages, commutation, and various tube parameters. For instance, ignitrons as switches in high peak current capacitor discharge service are considered in this chapter. Chapter 8 discusses grid and ignitor control circuits.

Chapter 9 is entitled "Ratings and Specifications and Their Interpretation." The object of this chapter is to give the reader a feeling for the ratings, which give tube capabilities only under conditions of operation most commonly experienced. The author does an excellent job of interpreting the meaning of the published ratings on these tubes. In Chapter 10, a summary is given of the important points in the operation and maintenance of the arc discharge tubes and what action to take in event of failure.

Mr. H. de B. Knight has written a very valuable reference book for tube engineers. He is to be commended for bringing to the attention of circuit designers and scientists the capabilities of present day gas discharge tubes in power circuits. The book is recommended for all workers involved in power control and electrical energy transfer.

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RECENT BOOKS

Bartee, Thomas C., *Digital Computer Fundamentals.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $6.50. Material covered may be understood readily with a prior knowledge of only elementary algebra and basic electronics. Suited for use as a textbook.

Bibbero, Robert J., *Dictionary of Automatic Control.* Reinhold Publishing Corp., 430 Park Ave., N. Y. 22, N. Y. $6.00.

Corben, H. C., and Stehle, Philip, *Classical Mechanics, 2nd ed.* John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $12.00. Falls into two parts—the development and use of Lagrange's equations and the development and use of transformation theory. New material included in this edition.

Draper, Charles S., Wrigley, Walter, and Hovorka, John, *Inertial Guidance.* Pergamon Press, Inc., 122 E. 55 St., N. Y. 22, N. Y. $6.50. International series on Aeronautical Sciences and Space Flight Division VII, (Astronautics) vol. 3. A treatise on the physical principles and engineering methods underlying the navigation and control of vehicles solely by means of signals from sensors that depend only on the inertial properties of matter for their operation.

Forsythe, George E., and Wasow, Wolfgang R., *Finite-Difference Methods for Partial Equations.* John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $11.50. Deals exclusively with the finite-difference method of solving partial differential equations. Presupposes a knowledge of advanced calculus and some knowledge of matrix theory, but not of the theory of partial differential equations.

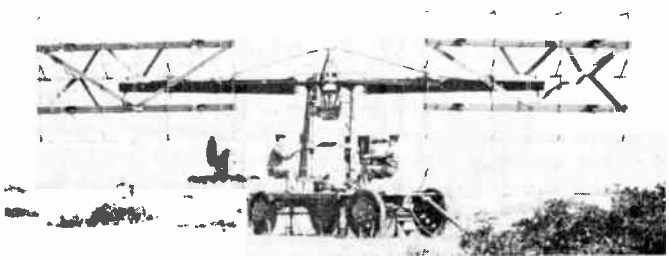
International Dictionary of Applied Mathematics, The. D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. $25.00. Terms and methods of application of mathematics to thirty-two fields of science and engineering, with multilingual indexes in German, Russian, French, and Spanish.

O'Connor, J. R. and Smiltens, J., Eds., *Silicon Carbide—A High Temperature Semiconductor.* Pergamon Press, Inc., 122 E. 55 St., N. Y. 22, N. Y. $12.50. Proceedings of the Conference on Silicon Carbide, sponsored by the Electronics Research Directorate, U. S. Air Force Cambridge Research Center, Bedford, Mass., held in Boston, Mass., April 2–3, 1959.

Ungar, A. Ed., *Proceedings of the 1959 Computer Applications Symposium.* Armour Research Foundation of the Illinois Institute of Technology, 10 W. 35 St., Chicago 16, Ill. $3.00.

Scanning the Transactions

Radar, history and the Signal Corps. Those who are interested in history may be interested in the photograph below which shows the Signal Corps' 1938 model of its famous SCR-268 radar. This initial model was designed to help position and aim searchlights for spotting aircraft. Those who are interested in history will also be interested to know (if they don't already) that the Signal Corps celebrated its 100th anniversary last year. And those who are interested in the past, present and future will surely be interested in the special Signal Corps Centennial Issue published by the PGMIL TRANSACTIONS last October. Its 35 papers, in presenting a survey of the many recent Signal Corps achievements in communications and electronics, have in effect provided an excellent picture of the current state of this many-faceted art, two samples of which appear elsewhere on this page. (Photo from: A. L. Vieweger, "Radar in the Signal Corps," IRE TRANS. ON MILITARY ELECTRONICS, October, 1960.)



The magic word today is "new power sources." Many new types of electrical power sources will be available to power the electronic equipment of the space age. Some are already available for practical applications; others are approaching the final stage of development. Where do we stand today? Chemical systems (batteries and fuel cells) are best for short-life applications. For long-life operation, solar energy systems show up most favorably where power requirements are low or moderate. Where both long life and high power are required, nuclear reactor systems appear to be the most promising if the problem of radiation hazards can be successfully overcome. Any consideration of power sources must also involve a discussion of two intimately related topics, energy conversion and energy storage. The two most promising thermal energy conversion devices are the thermoelectric and the thermionic converter. Although the thermionic converter is not presently as technologically advanced as the thermoelectric converter, its higher conversion efficiency makes it especially promising. As for energy storage, this can take many forms. Energy can be stored as heat, but this method is not very efficient. Energy can also be stored mechanically (flywheels, springs, etc.) but the weight-to-energy ratio is too high. Energy can also be stored for long periods by exciting solid crystals with X rays and gamma rays. This new approach is extremely interesting, but at present its weight-to-energy ratio is too high and efficiency too low. Batteries are still the best method, but storage in the form of bulk chemicals may ultimately be the most effective means of energy storage. The chemical compounds can be derived from the prime source of energy by a number of methods, for example, dissociation by thermal energy, dissociation by electrolysis, nucleo and photogalvanic effects. The regenerative fuel cell and photogalvanic systems are two examples of such storage systems. (David Linden, "Advanced power sources for communication electronics," IRE TRANS. ON MILITARY ELECTRONICS, October, 1960.)

TV aspect ratio: Is it time for a change? The television picture seen by the camera and transmitted over the air has a width to height ratio of 4 to 3, a ratio adopted from the pre-wide-screen movie era. Yet it has been more than a decade since television receivers have used a matching 4:3 viewing screen. The urge to enlarge the vertical dimension of the picture led first to the use of round screens and later to the present 5:4 aspect ratio. Although the 5:4 ratio has been heartily endorsed both by furniture stylists and home viewers as more attractive than the 4:3 ratio, it results in a loss of part of the transmitted picture at the sides and corners, waste of transmission time, and overdriving of sweep circuit components. The present 17-inch tubes don't show 16.3 per cent of the transmitted picture, while 19-inch tubes lose 12.5 per cent of the picture. Even if the corners of the screen were fully squared off, 6.3 per cent of the picture would still be lost. The question arises, why not change the aspect ratio of the transmitted picture from 4:3 to 5:4 so as to match present viewing screens? Only 5.4 per cent of the picture would be lost on a present-day 19-inch tube, and with fully squared corners there would be no loss at all. Moreover, the picture would be brighter and clearer, halftones would have better contrast, manufacturing costs would be reduced, sets would run cooler, and components would have longer life. It isn't likely that the proposed change would be adopted in the immediate future, but perhaps it is time to begin considering it. (W. D. Schuster and C. E. Torsch, "Benefits of a new aspect ratio for television," IRE TRANS. ON BROADCAST AND TELEVISION RECEIVERS, November, 1960.)

Many designers are looking toward electromechanical devices in their search for simpler and more reliable components which can perform complex functions. Much has already been accomplished by exploiting such phenomena as ferroelectricity of ceramics, magneto-restriction of ferrites, stable high Q's of mechanical resonators, and the low propagation velocity of sound through solids. The end is by no means in sight. Electromechanical filters, which have been given a considerable boost by the advent of single sideband communications, are now available over a range of 100 cps to 10 Mc. Ultrasonic delay lines already play an important role today in providing large delays in a small simple package. Ferroelectric power transformers without a core or coil are definite possibilities. It is expected that we will soon have available a new form of nonreciprocal circuit element for use at low frequencies, namely, an electromechanical gyrator, formed by mechanically coupling a magnetostrictive driving transducer to a ferroelectric ceramic output transducer. With the continued trend towards smaller and more complex equipment, this is a field which offers many interesting possibilities for the future. (E. Gikow, *et al.*, "Functional circuits through acoustic devices," IRE TRANS. ON MILITARY ELECTRONICS, October, 1960.)

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	IRE Members	Libraries and Colleges	Non Members
Broadcast and Television Receivers	BTR-6, No. 3	$2.25	$3.25	$4.50
Component Parts	CP-7, No. 4	2.25	3.25	4.50
Electron Devices	ED-7, No. 4	2.25	3.25	4.50
Electronic Computers	EC-9, No. 4	2.25	3.25	4.50
Military Electronics	MIL-4, No. 4	5.70	8.55	11.40
Nuclear Science	NS-7, No. 4	2.25	3.25	4.50

Broadcast and Television Receivers

Vol. BTR-6, No. 3, November, 1960

Professional Group on Broadcast and Television Receivers Administrative Committee (p. 1)

Minutes of PGBTR Administrative Committee Meeting, November 1, 1960 (p. 2)

Professional Group on Broadcast and Television Receivers Administrative Committee for 1961, 1962, 1963 (p. 4)

The Effect on Color TV of Dr. Edwin H. Land's Color Experiments—Charles J. Hirsch (p. 5)

Benefits of a New Aspect Ratio for Television—W. D. Schuster and C. E. Torsch (p. 13)

Phase Splitter Video Amplifier for Transistor TV—Zbigniew Wiencek (p. 18)

Severe requirements have to be met by the transistor as a video output amplifier. Of the few known types of video amplification—single transistor, push-pull, doubler, and phase splitter—the phase splitter, presented in this paper, represents the advantage of consuming least power, lower voltage and power transistor ratings, higher input impedance, high voltage gain, and no phase shift problems.

Common-emitter *p-n-p* transistor and common-base *n-p-n* transistor are arranged in a simple phase inverting configuration to drive the cathode and grid of the picture tube. Approximately one volt (*p-p*) of the single-ended input signal is adequate for the system. Equivalent to 50 volts (*p-p*) of the signal is easily available for each electrode of the picture tube.

Equivalent circuit, voltage gain, input impedance, thermal stability and bandwidth limitations are analyzed for this configuration. Discussion and comparison of all the video output systems is conducted in the conclusion.

A Television Stereophonic System—Robert B. Dome (p. 25)

The method of compatible stereophony described is especially designed for use with the aural channel of a television broadcast system. The make-up of the multiplex signal fed to the aural transmitter includes the use of a high fidelity 50–15,000 cps $L+R$ signal for the principal modulation. The $L-R$ information is transmitted by double sideband suppressed subcarrier together with a novel pilot frequency arrangement for use at the receiver for regenerating the subcarrier required for $L-R$ detection. Only one triode is required in the receiver demultiplexer. The stereo audio system is completed by the customary dual audio amplifiers and loudspeakers.

A Compatible Stereophonic System for the AM Broadcast Band—J. Avins (p. 30)

Stereophonic broadcasting presents a variety of problems, including compatibility, out-of-band radiation, stereo performance, distortion, signal-to-noise, receiver tuning, and receiver cost. Possible system solutions can be classified as either "multiplicative" or "additive," depending upon whether the quadrature sidebands representative of the stereo information $(L-R)$ are modulated by the envelope information $(L+R)$, or are essentially independent of the envelope information, respectively. The preferred AM-FM system uses a pre-emphasized $(L-R)$ signal to frequency modulate the carrier which is then amplitude modulated with $(L+R)$. Stereophonic field test transmissions over Radio Station WNBC have demonstrated compatibility with existing monophonic receivers and excellent reception on stereophonic receivers of simple design.

Power Sources for Low Heater Power Tubes in Transistorized Receivers—B. M. Soltoff and C. H. Tyson (p. 38)

A Transistorized AM-FM Receiver Using MADT Transistors—J. Heinchon and T. C. Lawson (p. 42)

A New Concept in Citizen's Band Equipment—H. L. Richardson, Jr. (p. 47)

A Nuvistor, Low-Noise VHF Tuner—G. C. Hermeling (p. 52)

Factors in the Circuit Operation of Transistorized Horizontal Deflection—R. W. Ahrons (p. 57)

Horizontal deflection circuits using germanium transistors can produce the deflection energy and the high voltage required to operate kinescopes. Theoretical and experimental evidence is presented to show that circuits using transistors specifically designed for deflection service can perform this function with greater efficiency and with fewer components than their counterpart: the vacuum-tube circuit. This paper discusses the factors which the engineer must consider for optimum circuit design.

The factors which most limit the performance of a transistor in a deflection circuit are the peak-to-peak operating voltage and the peak-to-peak current; the product of these factors is a measure of deflection energy. Although a transistor may be capable of producing the deflection required, additional considerations enter which may limit its performance in deflection circuits. These are 1) linearity of the deflection, 2) ringing during the trace period, 3) driving power, 4) heat dissipation, 5) symmetry of the transistor characteristics, and 6) economics of designing low voltage dc power supplies. An example is used to demonstrate the relationship between these factors and the compromises necessary to achieve an optimum deflection circuit design.

PGBTR at the IRE Convention, March 20–23, 1961 (p. 80)

Chicago Spring Conference, 1961 (p. 80)

Component Parts

Vol. CP-7, No. 4, December, 1960

Information for Authors (p. 115)

Who's Who in PGCT (p. 116)

Reversal of a Loaded Ferromagnetic Core—N. Cushman (p. 117)

This paper discusses the performance of a toroidal square-loop ferromagnetic core as a circuit element. The analysis takes account of two different physical processes in the core: a reversible one which contributes to the fast initial rise of the output signal, and an irreversible one, tentatively described as domain rotation, which is mainly responsible for switching the core. A differential equation is set up which governs the entire physical system, including the external circuit, and leads to a calculation of the output signal and its dependence on the secondary loading. Core parameters determined experimentally for an open-circuited core are used to predict the output signal when the core is loaded, and close agreement is found with measured values.

Capacitor Calibration by Step-Up Methods—Thomas L. Zapf (p. 124)

Step-calibration methods are used in many physical laboratories for the extension of measurements to quantities far removed from the magnitude of greatest accuracy at which absolute determinations are made. The excellent precision of repetitive substitution procedures is exploited by step-up or step-down methods to extend measurements to higher or lower magnitudes without serious degradation of accuracy. The application of step-up techniques to the calibration of variable air capacitors is described in this paper as a practical example of the method.

Design Characteristics of Nonvented Nickel-Cadmium Cells—Leonard M. Krugman (p. 129)

The properties and operating characteristics of nonvented, rechargeable, nickel-cadmium cells are determined and illustrated. The characteristics shown can be used to predict the performance of available sizes for circuit applications where a permanent, rechargeable dc power source is desired.

Contributors (p. 132)

Annual Index 1960 (follows p. 132)

Electron Devices

Vol. ED-7, No. 4, October, 1960

Cathode-Ray Switch Tubes with 100 and 256 Elements—R. Kalibjian and G. F. Smith (p. 189)

Switch tubes with 100 and 256 elements have been developed for data-processing applications. Simple space-charge and deflection-defocusing considerations indicate that in order to obtain maximum current output, the tube should be as small as possible. However, mechanical considerations and finite size of the beam crossover limit the amount of reduction. Aluminum targets are evaporated directly on the ground surface of the multielement header. The 100-element tube is $2\frac{1}{4}$ inches in diameter, 9 inches long, and provides a maximum target current of 3.0 ma at a modest 2-kv accelerating potential with reasonable deflection sensitivity. The 256-element tube has the same diameter, is 11 inches long, and provides 0.75 ma of output current at 2 kv.

The Grid Lens Effect in Convergent Pierce Guns—D. W. Shipley (p. 195)

In a conventional Pierce-type gun, the anode aperture causes a potential reduction in the cathode-anode region from the ideal Langmuir potential distribution. For low-voltage gating of the electron beam, a mesh grid of spherical shape (conforming to an equipotential surface) is used in front of the cathode. When this grid is operated at the Langmuir potential depicted by its relative position, there is a difference in the potential gradients on its two sides. This difference causes a lens action at each mesh element which results in a displacement of the actual electron trajectory from the ideal laminar trajectory in the region beyond the anode. A means for calculating these displacements as a function of distance along the axis is developed. As the grid lenses are divergent, the images of the mesh elements in any plane beyond the anode are larger than those for ideal laminar flow, resulting in a current density distribution which differs from that of the ideal beam. A means of calculating the current density profile by summing the effects of the grid lenses is devised and the method is applied to a sample gun design to illustrate the effect on the current density distribution.

Magnetically Tuned Klystrons for Wide-Band Frequency Modulation Applications—G. R. Jones and J. C. Cacheris (p. 206)

This paper describes the application of a ferrite-tuned X-band klystron to produce very wide-band frequency modulation with exceptionally low accompanying power variation. By perturbation techniques, a theoretical analysis of the magnetically tuned klystron is made in terms of the static and dynamic intrinsic properties of the ferrite. The observed operating characteristics show good agreement with the predicted behavior.

The design of a practical magnetically tuned X-band klystron for frequency modulation application is discussed in detail. Designs which included a fixed bias field transverse to the tuning field allowed frequency deviation over a 600-Mc bandwidth to be obtained with very good linearity at modulation frequencies below 500 cps. The change in power was less than ± 6 per cent. Another model permitted modulation frequency components up to 50 kc but tuning bandwidth was reduced to 300 Mc.

A Four-Terminal *P-N-P-N* Switching Device—Melvin Klein (p. 214)

A four-terminal germanium *p-n-p-n* switching transistor is described. In an appropriate circuit the device has useful switching gain for both the turn-on and turn-off operations. A pulse of one polarity turns the switch on, and a pulse of the opposite polarity turns it off. In the absence of triggers, the state is maintained. Equivalence to a pair of complementary conventional transistors in a direct-connected, common-base emitter-follower regenerative circuit is shown. The collector junction is common to both transistor sections so that saturation is simultaneous. Saturation current is kept small in one section, and the switch is turned off by withdrawing the small saturation current. Expressions are derived for switching gain, and requirements on device parameters are obtained. Common-emitter gain for one transistor section must exceed the ratio of load current to off-trigger current. Device structure is described. A combination of alloy, diffusion, and post-alloy diffusion techniques is used for fabrication. One transistor section with a thin base has high common-emitter gain, and the other has moderate gain. Switching performance data are given. A load current of 20 ma at 20 volts is switched with 2-ma pulses. Trigger pulse durations of the order of the switching time are required. Switching times of about 100 nsec are obtained.

High Frequency Effects of the Potential Minimum on Noise—J. R. Whinnery (p. 218)

A physical model of the compensation phenomenon in space-charge reduced shot noise is discussed in some detail and applied first to a planar diode of infinite spacing short-circuited to ac effects. Analysis of both the time response and frequency response shows that parameters of the potential minimum are very important to the degree of noise compensation or overcompensation, and for certain parameters the internal feedback mechanism of returning charges may cause a self-oscillation without external stimuli, or in other words a building up to saturation levels of noise phenomena in these frequency ranges. For other parameters of the minimum there appear to be very complete compensation phenomena possible, but practical parameters for this were not found in the infinite spacing case.

The infinite spaced short-circuited diode gives results in exact agreement with the frequency domain analysis of the open-circuited diode as analyzed by Siegman and Watkins. Some additional approximations may be made in this analysis for qualitative explorations, particularly important being the consideration of only a single velocity class of noise perturbation.

The short-circuited diode of finite spacing is analyzed in both the frequency domain and time domain, and an approximate correction to the above results found as a function of total transit angle across the diode. In particular the factor produces appreciable decreases in the reduction factor in the vicinity of certain transit angles across the diode, and as the Tien dips occur in these general ranges, the finite diode spacing is believed to be a strong candidate for explanation of these dips.

The Ubitron, a High-Power Traveling-Wave Tube Based on a Periodic Beam Interaction in Unloaded Waveguide—R. M. Phillips (p. 231)

The Ubitron is a high-power traveling-wave tube which makes use of the interaction between a magnetically undulated periodic electron beam and the TE_{01} mode in unloaded waveguide. The electron-wave interaction exhibits the same type of first-order axial beam bunching characteristic of the conventional slow-wave traveling-wave tube; hence, it can be used in place of conventional O-type interaction in extended interaction klystrons and electron accelerators, as well as traveling-wave tubes. Experimental results are presented for the simplest physical embodiment of the Ubitron, which consists of an undulated pencil beam in a rectangular waveguide. Two of the unique features of this tube are very broad interaction bandwidth which results from the absence of a dispersive slow-wave circuit, and variable interaction phase velocity—hence, variable saturation power level.

Among the physical embodiments of the Ubitron are a number of higher-order mode waveguide and beam configurations. These include plane, coaxial, and circular waveguides, all supporting the TE_{01} mode, interacting with magnetically undulated sheet, hollow and cylindrical beams, respectively. The advantage of these configurations, which have not yet been tested experimentally, is that they provide a very large interaction area for beam placement. This property, plus the fact that the peak interacting field is far from the waveguide walls, makes the Ubitron an interesting prospect for high-power millimeter wave amplification.

Photoeffect on Diffused *P-N* Junctions with Integral Field Gradients—A. G. Jordan and A. G. Milnes (p. 242)

A detailed analysis of a diffused junction photodiode is presented in which the illumination, monochromatic or broad-band, is applied to the diffused face. The electric field produced by the impurity distribution, assumed exponential, assists the transport and collection of minority carriers created by photons absorbed in the graded region.

The theoretical study covers both the steady-state and the transient response, and takes into account the effect of surface recombination velocity. The presence of the built-in field increases the photocurrent and reduces the dark current compared with homogeneous base diodes. For *p-n* silicon photodiodes with 5-micron base widths and acceptor concentrations of, say, 2×10^{18} atoms/cm^3 at the surface, photosensitivities of approaching 0.01 ampere per lumen may be achieved.

The transient-response analysis considers the extrinsic delay imposed by the time constant of the junction capacitance and the load resistance, and also the inherent delay caused by the transit time of the minority carriers. With moderate or high load resistances, the extrinsic delay is much larger than the transit-time delay. However, for comparable graded- and homogeneous-base photodiodes, the capacitances of graded junctions are lower, and therefore the transient response is improved on this account. The graded junctions also are shown to have greatly reduced transit-time delays because of the built-in field effect.

A Silicon Medium-Power Transistor for High-Current High-Speed Switching Applications—J. F. Aschner, C. A. Bittmann, W. F. J. Hare, and J. J. Kleimack (p. 251)

A diffused base, diffused emitter, *n-p-n* silicon switching transistor has been developed for high-current applications such as switching magnetic memories.

The transistor is designed to operate as a switch at the 0.75-ampere level. For a collector current of 0.75 ampere, the large signal current gain is 20 and the saturation voltage drop 4 volts. The breakdown voltages are 75 volts collector-to-base, and 6 volts emitter-to-base.

The unit shows fast switching characteristics. The rise, storage, and fall times are each of the order of 0.1 μsec. It has a common emitter unity gain frequency greater than 50 Mc.

The transistor employs a localized emitter produced by photo-resist techniques and oxide masked diffusion. Lead attachment is accomplished by compression bonding. The silicon wafer is bonded through a molybdenum intermediary to a massive copper stud.

The design theory of the device, and the variation of device characteristics with temperature are given. The applicability of this device to RF amplifier service is also discussed.

Avalanche Breakdown Voltages of Diffused Silicon and Germanium Diodes—C. D. Root, D. P. Lieb, and B. Jackson (p. 257)

Avalanche breakdown is defined in terms of the diode's electrical characteristics as well as the internal physical processes. Using the latter definition, and the basic diffusion equation, breakdown voltage is rigorously computed for various diffused junctions. The calculation process is described and similarities to both linear and abrupt junctions are pointed out.

Graphs are presented showing breakdown voltage as a function of diffusion parameters for both germanium and silicon. Intermediate charts used in the calculations are shown also. These give maximum electric field in junction and normalized breakdown voltage for families of diffusions.

A Crossed-Field Multisegment Depressed Collector for Beam-Type Tubes—D. A. Dunn, R. P. Borghi, and G. Wada (p. 262)

The velocity-dependent action of crossed electric and magnetic fields can be used to sort the electrons of the spent beam of a klystron or traveling-wave tube into two or more velocity classes, and to collect each class of electrons at a potential appropriate to its velocity. Secondary electrons produced at the collection surfaces do not return to the interaction space of the tube because of the nonreciprocal action of the crossed fields. An experimental tube permitted two-segment operation with 20 per cent of the beam power dissipated in the collector at small signals. Velocity sorting that permitted operation near this percentage of beam power over a considerable range of RF signal levels was observed. The major limitation was found to be migration of secondary electrons within the collector region from low-potential to high-potential electrodes, and this process prevented the depression of the collector power below 20 per cent of the beam power.

Relaxation Instabilities in High-Perveance Electron Beams—A. D. Sutherland (p. 268)

Spurious low-frequency oscillations similar to those reported by Cutler have been observed in the form of large amplitude modulation of both the RF output and the collector current of klystrons utilizing high-perveance electron beams. This modulation, which appears typically as sawtooth-like relaxation oscillations with frequencies ranging from 10 cps to 100 kc, was observed to occur spontaneously only if the operating pressure was lower than a critical threshold ($\approx 1 \times 10^{-7}$ mm Hg).

Experiments are described which were designed to investigate this phenomenon using dc beam testers. Evidence is presented which indicates that the oscillations are due to a relaxation back and forth between two possible solutions to Poisson's equation which both satisfy all boundary conditions. One of the possible solutions corresponds to the existence of a virtual cathode within the beam, resulting in a partial reflection of electrons. Experimental evidence indicates that positive ions play an important role in triggering the relaxation.

Both theoretical and experimental evidence are presented which point to relaxation instabilities of the type described as a potential hazard in utilizing depressed collector operation for improved efficiency of high-perveance microwave tubes. The same potential hazard exists in the use of periodic electrostatic focusing.

A Periodically Focused Backward-Wave Oscillator—C. C. Johnson (p. 274)

The application of some new periodic focusing techniques to the backward-wave oscillator is discussed. An oscillator utilizing a hollow electron beam was constructed at *S*-band. A periodic electrostatic focusing method was used with considerable success, and an octave tuning range was obtained. A periodic magnetostatic focusing method was attempted with less encouraging results. Some interesting effects were observed when the periodic focusing techniques were applied and some of these are described.

Voltages and Electric Fields of Diffused Semiconductor Junctions—C. D. Root (p. 279)

Equations are developed and graphs are presented, showing a functional relationship between electric field and space-charge widening and also between reverse bias voltage and space-charge widening, for diffused diodes. Planar geometry, with the diffusion made from a constant surface concentration (C_0) into material of constant impurity density, is assumed.

The voltage vs space-charge width relation is presented graphically for the case of silicon. Graphs for other materials are not included, since they would differ from the silicon graphs only by a constant in the voltage axis. These graphs illustrate that for small reverse voltages the junction may be considered as being essentially linear, while for large reverse voltages it may be considered as an abrupt junction.

A graph is provided which may be used to gain a qualitative indication of the electric field distribution within the space-charge region. This graph also gives an indication of the transition of junction behavior from nearly linear to nearly abrupt as the reverse voltage is increased.

Improvement of Beam-Tube Performance by Collector-Potential Depression, and a Novel Design—J. W. Hansen and C. Süsskind (p. 282)

The present state of the art of depressed-collector design is reviewed, and the characteristics of various designs are discussed. A novel design is proposed with desirable characteristics of most present designs but without the common disadvantage of adding to the size and weight of the tube. In the new configuration, the collector is placed off the axis of a magnetic pole piece, whose fringing fields serve to provide, in a simple way, the asymmetry needed to prevent the return of secondary and reflected electrons from the collector region. The performance of the new collector is evaluated experimentally, both in CW and pulsed operation, and is found to be promising. The device is readily adaptable to a rugged, simple collector for a commercial tube. The possibility of modification of the design to include velocity sorting is discussed.

A Nonreciprocal-Loss Traveling-Wave-Tube Circuit—R. N. Carlile and S. Sensiper (p. 289)

After a review of other related work and our own initial work, a nonreciprocal-loss, traveling-wave-tube circuit consisting of a series of re-entrant cavities coupled through ferrite-loaded apertures is described. For the configuration discussed, which is of the periodic magnetically focused variety, some measured characteristics of a possibly typical circuit (one with a cold pass band for the desired interaction mode of 20 per cent width centered near 9 kMc) are as follows: 1) a loss ratio of nearly 10 to 1 in db at band center dropping to about 4 to 1 in db near the band edges, 2) a ratio of backward loss to usual forward gain per unit length adequately larger than unity, and 3) small and practically negligible changes in the phase velocity, interaction impedance, circuit-matching characteristics, and peak gap axial periodic-focusing magnetic field from these same properties in a non-ferrite-loaded circuit. Mention is made of the use of ferrite for obtaining reciprocal loss. Finally, the possible application of the ferrite-loaded-aperture nonreciprocal circuit in modified form to other kinds of traveling-wave interaction devices (parametric amplifiers and masers) is noted.

Recovery Technique for Saturated Masers—Gunter K. Wessel (p. 297)

The practical application of masers is often hindered by saturation effects of the maser material. Any strong signal frequency (*e.g.*, the transmitter pulse of a radar system leaking into the maser) will cause saturation. This paralyzes the maser for a period of time, too long for most intended applications. In this report, a method is described which has been used in desaturating a four-level ruby maser.

The saturation is caused by equalizing the populations of levels one, two, and three of the maser crystal by the leakage of the transmitter pulse, in addition to the action of the maser pump. The application of a desaturation pulse between levels one and four will then desaturate the maser, restoring the excess population density between levels two and one. In a cavity maser, full recovery of the amplification capabilities has been achieved within less than 1 msec after the maser has been saturated. Repetition rates up to 120 cps are tolerable.

An Analysis of Inertial Inductance in a Junction Diode—I. Ladany (p. 303)

When the injected carrier concentration is not small compared to the equilibrium concentration, drift and diffusion processes interact, and an inductance is developed in the base region. This phenomenon has been analyzed by calculation of the small-signal impedance of a PIR diode. Simple expressions have been obtained for the case $\Omega \ll 1$ as well as $\Omega \gg 1$, where Ω^2 is the product of the radian frequency and the diffusion transit time. For intermediate values of Ω, the impedance contains an integral which has been expressed in terms of a tabulated function. Curves are presented showing the diode impedance as a function of frequency and forward bias. For low bias the reactance is capacitive, but with increasing bias, provided the frequency is not too high, the reactance becomes inductive. The resistive component also shows a frequency dependence. The maximum inductive reactance occurs when $\Omega \approx 1$ and the Q does not exceed 1.

Contributors (p. 311)

Annual Index (follows p. 315)

Electronic Computers

Vol. EC-9, No. 4, December, 1960

Statistical Recognition Functions and the Design Pattern Recognizers—T. Marill and D. M. Green (p. 472)

According to the model discussed in this paper, a pattern recognizer is said to consist of two parts: a receptor, which generates a set of measurements of the physical sample to be recognized, and a categorizer, which assigns each set of measurements to one of a finite number of categories. The rule of operation of the categorizer is called the "recognition function." The optimization of the recognition function is discussed, and the form of the optimal function is derived. In practice, a prohibitively large sample is required to provide a basis for estimating the optimal recognition function. If, however, certain assumptions about the probability distributions of the measurements are warranted, recognition functions that are asymptotically optimal may be obtained readily.

A small numerical example, involving the recognition of the hand-printed characters *A*, *B*, and *C* is solved by means of the techniques described. The recognition accuracy is found to be 95 per cent.

The Simplification of Multiple-Output Switching Networks Composed of Unilateral Devices—G. C. Vandling (p. 477)

The purpose of this paper is to show that two-level (no more than two gates in cascade) multiple-output switching networks composed of unilateral switching devices such as diodes can be simplified or minimized in much the same manner as single-output networks.

This is accomplished by extending the notation and techniques used in the simplification of two-level single-output switching networks to multiple-output switching networks. A simple procedure for identifying multiple-output prime implicants is devised and, as a final result, an algorithm is presented which can be used to minimize the switching network corresponding to a number (q) of given Boolean expressions of the same variables. This algorithm is based on the Quine rules but has been modified to take advantage of the so-called "don't care" conditions which occur because some inputs are forbidden or because some outputs are of no concern. This algorithm can readily be programmed on a digital computer if desired.

Uniqueness of Weighted Code Representations—G. P. Weeg (p. 487)

Decimal computers ordinarily use a binary-coded decimal representation. One class of binary-coded decimal digits is the so-called four-bit weighted code representation with weights $w_1\ w_2\ w_3\ w_4$. Each w_i is a nonzero integer in the range $-9 \leq w_i \leq 9$, and the set of weights must have the property that every decimal digit can be represented by the sum

$$\sum_{i=1}^{4} b_i w_i,$$

with the b_i being 0 or 1.

For some weighted codes the weights are such that some digits can be represented by more than one sum of the specified form. For example, the 7421 weighted code has the property that 7 may be represented either as 1000 or as 0111.

This paper produces a necessary and sufficient condition on the weights of a weighted code for the unique representation of each digit by a sum of the specified form. Further, all possible sets of weights are displayed.

Analog Representation of Poisson's Equation in Two Dimensions—R. J. Martin, N. A. Masnari, and J. E. Rowe (p. 490)

A new analog device, called a Poisson cell, has been developed which aids in obtaining solutions to either Laplace's equation or Poisson's equation. The cell may be used to simulate such potentials as electric potential, magnetic potential, gravitational potential, and the velocity potential of irrotational flow; it has applications in the fields of hydrodynamics, heat conduction, and aerodynamics.

The cell is a solid volume-conducting medium made from a homogeneous mixture of hydrostone and graphite. Electrode configurations may be painted on the surface with conducting paint or imbedded directly in the structure. In the case of Poisson's equation, where $\nabla^{-2}\phi(x, y) = f(x, y)$, the function $f(x, y)$ is simulated by injecting currents into the underside of the cell.

The application of the Poisson cell to numerous problems and in particular to problems in electron flow is discussed in detail, along with the incorporation of the cell into either an analog computer system or a combined analog-digital computer system.

A New, Solid-State, Nonlinear Analog Component—L. D. Kovach and W. Comley (p. 496)

Since the inception of the electronic analog computer as a useful engineering tool, the need for practical methods of solving nonlinear problems has steadily increased. This paper describes a passive, nonlinear device which, when used with operational amplifiers, provides the means for obtaining a large class of functions. These are obtained to a degree of accuracy and reliability not previously possible with a simple, economical device. A basic varistor squaring unit is described. The unit has been compensated for the various types of error inherent in the varistor itself, and is capable of providing approximately fifteen of the most basic and commonly used nonlinear functions.

Solving Integral Equations on a Repetitive Differential Analyzer—R. Tomović and N. Parezanović (p. 503)

Methods of practical solution of integral equations on electronic differential analyzers are not well developed. In those cases where such methods have been outlined, special and costly additional equipment is required.

Results presented in this work show that practical solution of integral equations is possible using a repetitive differential analyzer of convenient design.

A New Method for Analog Integration and Differentiation—M. A. Thomae (p. 507)

A technique is described which enables an approach to ideal analog integration or differentiation by means of passive elements only. A series of RC circuits in a cascade arrangement, uncoupled to each other, provides the first, second, third, etc., integrals or derivatives (according to the connection of the RC circuits) of the input function. The theory establishes that if the outputs of each are fed to an analog summing amplifier, its output becomes arbitrarily close to the ideal integral or derivative of the input function as the number of RC stages is raised indefinitely. A device has been developed according to this idea to perform one of these operations (integration) and to check the results obtained in theory. Mathematical proofs of the theory are given in the paper.

Correction to "Conditional-Sum Addition Logic"—J. Slansky (p. 509)

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Military Electronics

Vol. MIL-4, No. 4, October, 1960

In Memoriam, James Q. Brantley, Jr. (p. 392)

Frontispiece, Harold McD. Brown (p. 394)

One Hundred Years of Service (p. 395)

Editorial Committee (p. 396)

One Hundred Years of Research—Harold A. Zahl (p. 397)

On June 21, 1960, the U. S. Army Signal Corps celebrated its 100th birthday—a century of service to the Army and the Nation.

In this narrative, particular attention is directed toward a few selected items of research which characterize the scientific past and present of the Corps; it is concluded by a look toward the future.

Stressed are the facts that, over the years, signal research has had a profound effect on the nation's military posture, that this is of importance in periods of wartime stress, and, finally, that the peacetime economy has also gained, both directly and indirectly, from much of this research. The article also includes a short summary of the Greely mission to the Arctic covering the period of 1881–1884—the tragedies of the expedition, and its scientific achievements.

Looking ahead, the narrative concludes with brief description of signal research in several areas which today show great promise as the nation moves forward into the unknown of tomorrow.

Army Signal Corps Organization for Research and Development—I. R. Obenchain, Jr. (p. 402)

The Signal Corps' research and development and operational mission responsibilities have created capabilities which readily lend themselves to applications in the space electronics field. The past accomplishments in satellite electronics and in satellite ground support are highlighted and an outline of future work is given.

The Signal Corps at the Space Frontier—Hans K. Ziegler (p. 404)

The Army Signal Corps organization for research and development is described. Research and development and combat development are closely related in the Signal Corps. The Office of the Chief Signal Officer staff in Washington and five Signal Corps field agencies have responsibilities in these areas. The relations among these five agencies as well as coordination with other key Signal Corps activities are summarized.

Courier Satellite Communication System—G. F. Senn and P. W. Siglin (p. 407)

This paper on the Courier satellite presents some of the background against which this experiment on an operational satellite communication system is being conducted. Details on the development, construction and testing of both the ground and orbiting segments of the system are outlined. Information on the results expected from the operational aspects of the experiment are also presented.

Exploratory Research in the Signal Corps—N. J. Field and I. A. Balton (p. 414)

This paper reviews the USASRDL exploratory-research program from its inception in 1958 to its current state of development. The treatment centers about three main parts: background and philosophy of exploratory research in the Signal Corps, implementation of the program at USASRDL, and major achievements.

The introductory portion deals with a discussion of the motivation that led to a delineation of exploratory research as a modified basic-research concept, and the activation of an exploratory-research program. This discussion is followed by an exposition of the program and of the scientific areas that are currently being investigated. The article then gives a summary of the important accomplishments resulting from the program. A brief report is also given of plans for the initiation of new research tasks in the future.

Signal Corps Studies of Nuclear Radiation Effects on Electronic Components and Materials—E. T. Hunter, L. L. Kaplan, and A. L. Long (p. 419)

The U. S. Army Signal Research and Development Laboratory has long been interested in the effects of nuclear radiation on electronic components and materials. An historical account of USASRDL participation in earlier weapons tests is presented along with more recent efforts toward the evaluation of component operation during a burst of nuclear radiation using various pulsed reactor facilities. The radiation effects on some present-day components are briefly described, and possible directions of future efforts to develop radiation-resistant components and equipment are proposed.

Precision Frequency Control for Military Applications—E. A. Gerber and J. M. Havel (p. 424)

The paper discusses progress made in the field of frequency control over the past years, with emphasis on the contributions of the U. S. Army Signal Corps. The entire field is reviewed, including piezoelectric resonators and devices using atomic and molecular spectral lines. Several accomplishments which are thought to be of special importance are discussed in greater detail. These subjects include among others: Properties of synthetic quartz, aging prediction of crystal units, new methods of temperature compensation for crystal units, piezoelectric crystals for the VHF range, measurement methods for piezoelectric resonators, cesium-beam frequency standards, ammonia maser, and gas cell devices.

The Signal Corps Synthetic Quartz Program—J. M. Stanley (p. 438)

The role of the Signal Corps in developing the various techniques now being used to synthesize quartz crystals is reviewed from the standpoint of the quality of the material produced, growth rate, growth conditions, seed orientation, feed materials, cost, and suitability for production.

Differences in the structure of synthetic quartz are indicated by variations in the optical absorption coefficients of X-irradiated quartz plates taken from different growth directions. The optical absorption measurements on the different types of synthetic quartz are compared with one another.

Differences in the structure of natural and synthetic quartz are also indicated from measurements of Q made on 5-Mc high-precision resonators fabricated from samples of synthetic quartz grown under different conditions and from the occurrence of large absorption peaks at low temperatures (50°K) noted when measuring such resonators. Data is presented to show that synthetic quartz grown in the Z-direction possesses higher Q's than other quartz growth directions. None of the 5-Mc units made from synthetic quartz approach the high Q typical of natural quartz 5-Mc units between −60°C and 100°C.

Comparison of the frequency temperature characteristics of overtone high frequency resonators made from synthetic quartz with those made from natural quartz show them to be about the same. The growth direction and impurity content, however, change the optimum orientation angle and inflection temperature.

Resonators processed from synthetic quartz for operation in the 60–202-Mc range are shown to perform as satisfactorily as similar resonators made from natural quartz. No differences were noted in the processing characteristics, frequency temperature coefficients, and resonance resistances of the two materials. In both natural and synthetic quartz VHF resonators the Q is shown to decrease at about the same rate with increasing frequency.

The processing potentials and current material costs of natural and synthetic quartz are compared with each other. Areas where additional quartz synthesis information is needed are discussed briefly.

The Signal Corps Program on Magnetic Ferrites—E. Both, I. Bady, and W. W. Malinofsky (p. 448)

The ferrite research and development program pursued by the Signal Corps over the past twelve years is reviewed and its most significant results are described. The discussion includes all major phases and sub-tasks under this program; such as the work on low-loss high-permeability materials for communication uses at frequencies from 0.1 to 500 Mc, square hysteresis loop ferrites for computer applications, work on permanent magnet materials, and materials for microwave applications up to 100 kMc. In addition, studies dealing with the effect of impurities, preparation techniques, grain size and other parameters on certain material characteristics are mentioned briefly. A list of publications originating from this work is included.

New Trends in Signal Corps Transistor Development—B. Riech and J. Shwop (p. 455)

Signal Corps transistors are discussed with special emphasis on device developments during the last five-year period. A generic development is presented starting with germanium and silicon alloy-junction transistors and leading to germanium and silicon diffused junction-type devices that reflect the extended frequency and power capability of the transistor. Devices developed under Signal Corps contracts are described and the importance of these new transistors for military and industrial use is illustrated. In addition, a graphic picture of the growth of the Signal Corps developments is included.

Silicon Integrated Circuits—William B. Glendinning (p. 459)

The use of semiconductor integrated circuitry to reduce size and weight of military electronic systems is discussed. The pure semiconductor integrated circuit approach used to obtain micromicrominiaturization levels is reviewed and characterized. Described in this paper are the design and construction of a silicon integrated microcircuit. Results of examination of the electrical performance of the microcircuit are presented, including results of examination of the thermal behavior of the microcircuit. The silicon structure is analyzed in terms of internal and external geometry, material properties, and electrical performance.

Functional Circuits Through Acoustic Devices—E. Gikow, A. Rand and J. Giannotto (p. 469)

In essence, the communication specialist in the field with his complex electronic gear and the native who beats a hollow log both perform the same function—communication over a distance. However, there is no gainsaying that the hollow log is undoubtedly the more reliable of the two communication equipments.

The implication of this analogy is not as far fetched as it may appear. It has been found that considerable circuit simplification with a consequent increased reliability is attainable through the application to electronic circuits of the mechanical and acoustic properties of certain materials. Miniaturized, rugged, low cost, highly selective band-pass filters ranging from 100 cps to approximately 10 Mc are feasible. Power transformers without core or coil are definite possibilities. Stable delay media provide large delays in small simple packages. Hybrid structures of conventional parts and nonconventional resonators provide techniques for broadbanding and impedance transformation. A new element, it is expected, will soon be generated for use at low frequencies—a passive four-terminal nonreciprocal network.

Electron Tubes and Devices in the 4.3-MM Frequency Range—H. J. Hersh, E. J. Kaiser, F. E. Kavanagh, and I. Reingold (p. 481)

This paper discusses the development programs for a klystron, a magnetron, and various duplexing and switching devices for the 70-kMc region of the radio frequency spectrum during the last decade. An evaluation of the electrical performance and of the behavior under environmental conditions for these devices is also discussed. It is concluded that the quality and performance of the devices make system application feasible. Sections I and II describe the klystron and magnetron program. Section III describes the switching devices program.

Harmonic Generation by Means of Traveling-Wave Tubes—F. E. Kavanagh and K. Westermann (p. 493)

This paper describes the results of an experimental investigation conducted to determine the effectiveness of a traveling-wave tube used as a harmonic generator. It is shown that harmonics of a low frequency applied to one of the gun electrodes will be amplified by the helix and will appear at the output of the traveling-wave tube as a "picket fence" of RF signals. The paper describes the experimental techniques used and discusses the test results obtained. Among the parameters considered are carrier-to-noise ratio, uniformity of power output, and stability of the harmonics produced. Possible applications of a traveling-wave-tube harmonic generator are also discussed.

Advanced Power Sources for Communication Electronics—David Linden (p. 497)

Portable electrical power sources are being used in increasing numbers for a variety of applications on the ground and in outer space. The trend to miniaturization and transistorization has accelerated this practice.

Chemical, nuclear, and solar energy are the three prime sources that are being used in power sources which will fulfill the new requirements. Each of these sources, in combination with new electrical conversion devices, has advantageous and unique characteristics which make it desirable for this application.

The characteristics of various types of recently developed electrical power sources are described and compared, and data are presented which illustrate the best operating conditions for each system.

Two Decades of U. S. Army Mobile High-Frequency Communications—H. C. Hawkins (p. 502)

This paper is an historical account of the trend in development of U. S. Army high-frequency mobile radio communications of medium power in the period 1940 to 1960. Starting with one of the first really mobile sets of this power level, in a wide variety of installation configurations to meet World War II requirements, it relates the progressive fulfillment of higher traffic handling capabilities in both single and multivehicle installations. The increasing demand for smaller but higher capacity equipments indicates the course which future development efforts should follow.

Multichannel Radio Communication Within the Army—Lawrence G. Fobes (p. 505)

Combat communications equipment is a highly specialized weapon in its own right. USASRDL has been an important factor in the development of combat communications and in the establishment of an industrial design capability for such equipment.

An example is given by tracing the growth of Army multichannel communication systems from the radio-link equipment of the Tunisian campaign, through "backbone" radio relay employment exemplified by radio sets AN/TRC-1, AN/TRC-24, and AN/TRC-29, into tactical-mobile systems giving area coverage using AN/GRC-50, 53, 59 and 66 equipment.

Important milestones of concept and technique changes illustrate the continued progress towards reliable tactical radio communications.

Automatic Data Processing Equipment Program for the Field Army—Milton A. Lipton (p. 514)

In 1956, the U. S. Army Signal Corps was directed to undertake the development of automatic data processing (ADP) systems for the Army in the field. The development of a mobile, general purpose, digital computer, named MOBIDIC, was initiated concurrently with an Army-wide analysis of the specific tactical applications in which it could suitably be used. As the application studies progressed, it became evident that a family of computers was required for the different tactical echelons. The FIELDATA Equipment Program evolved and is now current for this purpose.

This program encompasses major activities in computer design, programming techniques, data transmission equipment, input-output equipment, advanced techniques, and human factors studies, the last named being considered most critical for successful tactical usage of ADP. A set of standards has been formulated and adopted for all FIELDATA equipment in the factors of alphanumeric code, control function code, computer word structure, computer order code and equipment interconnection methods.

The first MOBIDIC has been delivered to the U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J., and is now undergoing acceptance and evaluation testing.

A crash program was recently assigned to the Signal Corps for the construction and delivery in this calendar year of a MOBIDIC especially arranged for a large-scale logistics function of the Seventh U. S. Army in Europe. The total Signal Corps role in this program includes equipment design and fabrication, system analysis and programming, troop training, and initial system operation.

The success of ADP in the Field Army will now rest with the yet-to-be-determined validity of system concepts, merits of the equipment designs and degree of user acceptance.

A Four-Wire Electronic Automatic Communication Switching System for a Field Army—G. W. Bartele, R. A. Pfeiffer, and S. B. Weiner (p. 519)

Plans for a new communication network, for a military theater of operations, have been under consideration since the end of World War II. It is axiomatic that a common user communication system, in which channels of communication are shared among a large number of users, must provide means whereby these channels are used with a maximum efficiency. In such a system, efficiency of operation depends to a large extent on the rapidity with which a large volume of communication traffic can be switched throughout the network, with a minimum of time being expended in setting up and taking down numerous connections.

This article recounts the history of the development of a communication channel switching system as dictated by changing military requirements, accenting wide dispersal and mobility and the effect of this system concept on the design of military communication switching equipment. Early efforts to improve manually operated switchboards are described, with the aim of making clear the logical sequence of ideas leading to the development of the military four-wire automatic electronic switching system, now nearing completion. This automatic system is described in some detail as to its operating characteristics, and the reasoning behind the equipment design is discussed. The organized program to insure reliability of operation carried on concurrently with the equipment development is outlined and results described. Finally, the shape of things to come in military electronics switching equipment, as a logical next step from the point where it now stands, is prognosticated.

Electronically Tunable Circuit Elements—Samuel Stiber (p. 527)

During the past sixty years, attempts have been made to develop electronically-controllable circuit elements to be applied to filter networks in tunable receivers and transmitters. Both the polarization of the dielectric material used in the construction of capacitors and the permeability of the ferrite core material used in inductor design, lend themselves to electronic variation; in fact, the change in the incremental permeability of a magnetizable material in the presence of a superimposed varying magnetic field was patented as early as 1901. Usable circuit elements, however, were not realizable for application in the frequency ranges above 100 kc, nor did the available materials provide the desired characteristics for modern filter design of receivers and transmitters.

USASRDL initiated a program in 1948 to investigate the problems associated with electronically variable inductances. New ferrite materials were developed and a better understanding of the over-all problem obtained. This program grew in magnitude, calling upon the efforts of both universities and commercial organizations, in addition to the internal effort within this laboratory. The outcome is a family of controllable inductances, electronically tunable in the frequency range dc to 500 Mc with reasonable Q's, power requirements, and small in size and weight. Large frequency variations are obtainable and they are usable under environmental conditions of temperature, humidity and vibration encountered in field conditions. With these electronically tunable inductors the military now have available receivers which can be remotely tuned, rapidly scanned over frequency ranges of better than two-to-one and stopped quickly on receipt of a signal. They can sweep large frequency ranges and present a full panoramic display of the signal environment present in each range, together with a simultaneous expanded sector display of the signals of interest.

This paper discusses fundamental considerations and definitions of ferromagnetic, ferroelectric and back-biased devices. It describes the problems involved in the development of the electronically controllable inductance, provides a discussion of requirements this device must meet, and gives operating characteristics of inductors developed. Applications of the device in military equipment are discussed, some examples of their use in receivers described and some typical circuits provided. A comparative evaluation of ferromagnetic, ferroelectric and back-biased diode tuning devices completes this paper.

The Signal Corps' Contribution to the Microwave Antenna Art—Leonard Hatkin (p. 532)

This paper contains a brief summary of the Signal Corps' contribution to the antenna art. Emphasis is on the post World War II period and on microwave antennas. Only developments which have had reasonably wide application are discussed.

Considerations of a Tropospheric Scatter System for Field Use—J. C. Domingue (p. 536)

The modern Army will place greater demands on communication requirements for capacity, range and mobility. To meet these demands with conventional equipment results in a serious increase of manpower. The encouraging possibility of employing tropospheric scatter equipment to augment and possibly replace some conventional equipments has distinct advantages. This paper surveys the basic requirements of a tropospheric scatter system and expands them to fulfill the requirements of tactical employment. Subjects such as range and frequency dependence, effect of horizon angle, modulation techniques, and antenna gain degradation are reviewed to develop the technical foundation of a tactical system. The additional requirements of mobility, simplicity, and quick response are added to present a system designed for tactical utilization. Actual results obtained over a number of paths with digital data and facsimile transmissions are presented. The paper concludes with a discussion of inherent problems in tropospheric systems and the developments required to alleviate some of the difficulties.

Electromagnetic Environmental Testing—R. E. Frese and J. H. Homsy (p. 540)

A number of factors which contributed to the establishment of the Army's Electromagnetic Environmental Test Facility (EETF) at the U. S. Army Electronic Proving Ground are reviewed. Technical planning to date covering the layout, orgnization and operation of the EETF is summarized. Finally, predictions as to the possible future scientific role and activity of the EETF are made.

Signal Corps Field Telephones—Raoul A. Faralla (p. 543)

Signal Corps pioneering in the design and development of U. S. Army field telephones is discussed. Explanation is given of the military need for the new features and performance improvements desired in field telephones, and of the methods whereby these new requirements were resolved by Signal Corps engineers. In each instance, the Signal-Corps-developed field-telephone sets performed better than contemporary commercial telephone sets. A few of the Signal Corps inventions and innovations have been adopted by the U. S. telephone industry and incorporated into the design of the latest subscribers' telephone sets, whose improved performance is now comparable to that of present standard Signal-Corps-developed field-telephone sets.

The new features of the recently developed, fully transistorized, Signal Corps field-telephone set for use with the completely automatic, electronic-switching, military-communication system are also discussed. The novel circuit features of this telephone make it possible, for the first time, to have a subscriber's line carry only voice-frequency currents at voice-frequency communications level.

From Eye to Electron—Management Problems of the Combat Surveillance Research and Development Field—William M. Thames (p. 548)

Following an introduction which establishes the fact that increased fire power of modern armies has increased the requirement of those armies for long range combat surveillance and target acquisition capabilities, this article highlights the four basic management problems involved in the combat surveillance R&D field:

1) The initial establishment of relationships with industry, other Signal Corps establishments, other branches of the Army, other Services, and with U. S. CONARC, principal user of the equipment.

2) The development of a logical program.

3) The development of the Combat Surveillance Agency itself as an organization.

4) The development of a technique for monitoring each program and for supervising the agencies, corporations, etc., involved in the various steps of the program.

The solutions that management has applied in order to resolve these problems are discussed in sequence. These solutions include the liaison and information exchange function, a three-pronged program objective, the "systems manager" type of organization and its merits, and a discussion of the "line of balance" chart monitoring system.

The Problem of Combat Surveillance—Herbert A. Nye (p. 551)

Some of the characteristics of modern warfare which contribute to the difficulty of combat surveillance are summarized. Functions required of modern combat surveillance and techniques of performing these functions are outlined. The role of the Signal Corps in combat surveillance development is discussed, and current developments in combat surveillance are discussed, with particular emphasis on problem areas.

Radar in the Signal Corps—Arthur L. Vieweger (p. 555)

The accomplishments of the Signal Corps Laboratories in the development of radar are described with emphasis on the pre-World War II period. The development and evolution of early warning radar and fire control radar in the Army are treated in some detail. The ground radars used in combat during the war by the Army (including the Army Air Corps) are discussed.

Application of Signal Corps Radar to Combat Surveillance—Albert S. White, Jr. (p. 561)

The U. S. Army Signal Research and Development Laboratory has developed several types of radar equipment to fulfill the combat requirements of the Field Army. To aid in obtaining an all-weather continuous surveillance capability, the AN/PPS-4 and AN/TPS-25 ground radar sets were developed. To extend and supplement their coverage of the combat zone, the AN/APS-94 airborne radar was developed. Locating weapons of interest is another important function of combat surveillance. The AN/MPQ-4 mortar locator was developed for this purpose.

Lightweight Integrated Doppler Navigation System for Army Aircraft—Kenneth K. Kelly (p. 565)

This paper presents some of the problems associated with the Signal Corps search for a suitable Doppler navigator for both lightweight fixed wing and rotary wing aircraft. The unique approaches discussed in detail are: The choice of a lightweight 3-axis stabilized antenna, position computation and display in cartesian coordinates for compatibility with Universal Transverse Mercator charts, a moving bug map display, a tape display of absolute and barometric altitude, and an integrated instrument system displaying navigation and flight information for all weather operation. These various components are combined to furnish the Army with a truly automatic self-contained navigation and flight instrument system.

Development of Meteorological Instrumentation at USASRDL—D. A. Deisinger (p. 572)

This article describes the development of electronic meteorological instrumentation at

USASRDL from the early days of 1930, when practically no electronic equipment was employed for meteorological purposes, down to the present time. The discussion covers the development of radiosondes at high, VHF, and microwave frequencies, radio wind-finding equipment both manual and automatic, dropsondes to measure the temperature and humidity when dropped from an airplane and carried to earth on a parachute, weather radar equipment, sferics equipment for locating thunderstorms, rocketsondes, and other meteorological equipment. The history of various developments is presented as well as technical considerations which led to the engineering solutions to the various problems as they arose. The entire development program is related to progress in the field of meteorology and the usages of weather information which produced the urgency for developing the new types of electronic equipment.

Radar Ballistic Instrumentation at White Sands Missile Range—Lyle D. Bonney (p. 583)

Early in the history of White Sands Missile Range a runaway V-2 missile indicated the need for better missile flight surveillance. Modified SCR-584 radars were soon after established as the basic source of surveillance data. Further modifications to the system made reliable all-weather trajectory data, readily achieved by radar. The development and implementation of the AN/FPS-16 radar at White Sands Missile Range represents the product of years of engineering to create a precision radar instrumentation system.

The Development of a Dynamic Target and Countermeasures Simulator—Richard L. Norton (p. 587)

The Dynamic Target and Countermeasures Simulator described herein was developed at the U. S. Army Signal Missile Support Agency Laboratories, White Sands Missile Range, Las Cruces, N. M., by a group of Signal Corps engineers and technicians, utilizing components from scrapped radars and standard RF test equipment.

This versatile simulator can be used to test existing radars for signal saturation and electronic counter-countermeasures techniques, such as regulating RF and video gain, contrast, etc. Provision is made to simulate antenna patterns and to vary the size of the target from 0.1 to 100 square meters.

The expense involved in training radar operators to cope with raids involving hundreds of aircraft has led to the development of various target simulators.

As the state-of-the-art in electronic countermeasures (ECM) improved, it became apparent that ECM simulation was also necessary to train radar operators under the conditions that would prevail when aircraft would be radiating thousands of watts of jamming power. Most of the simulators developed used IF or video injection of target signals in the simulator and did not provide the realism desired or allow the radar operator to exercise electronic counter-countermeasures (ECCM) techniques effectively.

The problem then became one of realistic simulation of both target and jamming signals. Field experience at White Sands Missile Range (WSMR) had demonstrated that simulation could be accomplished most effectively at the RF level. This finding, by WSMR engineers, led to the design and development of the Dynamic Target and Countermeasures Simulator.

The Signal Corps Role in the Development of Continuous-Tone Electrophotography—E. K. Kaprelian and K. Leistner (p. 591)

This paper discusses the reasons which led the Signal Corps to investigate different nonsilver halide photographic processes, and to select electrophotography as the subject of a major research and development effort.

The problems inherent in the process are discussed and several methods with which it was attempted to evolve a workable procedure are shown. In particular, the research directed toward increasing the spectral response and over-all sensitivity of selenium layers, as well as of other semiconductors, is outlined. Other problems discussed are spontaneous decay of the charge in the dark, plate fatigue, and afterimages. The different approaches to the conversion of the electrostatic latent image into a visible, permanent image are discussed, including liquid spray and immersion methods, dry "cascade" and powder cloud development. Emphasis in the discussion is on the powder cloud method and its related problems: selection of the most suitable powder material, uniformity of particle size, polarity and amount of charge of the particles, design of suitable cloud generators, particle agglomeration. It is shown that certain problems inherent in the powder development on selenium plates can be simplified by first transferring the electrostatic latent image from the reusable selenium to the permanent carrier, and developing on this carrier.

The paper finally discussses the possibilities for future developments and applications, and calls attention to a comprehensive list of pertinent patents available to readers on request.

Trends in Army Signal Corps Research and Development Procurement—R. F. Wilson and W. H. Blatti (p. 601)

Signal Corps and general military electronic R&D procurement is undergoing continuing evolution in technology, techniques and controls, which have direct influence on industrial planning, management and operations. The close partnership of the military and industry requires cognizance of and response to these pressures. Systems procurements generate the industrial "team effort" approach. Small business is gaining a strong foothold in the R&D effort through diversity of electronic research.

Certain aspects of R&D are stimulating a limited move towards greater use of fixed price contracts. New trends have effect on industry approach to preparation of bid proposals and the military techniques of evaluation. Engineering efforts require management comptroller surveillance to avoid cost problems. Over-runs in cost contracts are being subjected to increasing military controls and limitations with serious effects on contractor management techniques. Contract management in industry is evaluated in terms of effective response to procurement trends.

Contributors (p. 608)

Annual Index (follows p. 615)

Nuclear Science

VOL. NS-7, NO. 4, DECEMBER, 1960

A High Sensitivity Semiconductor Diode Modulator for DC Current Measurement—Harold E. De Bolt (p. 1)

A semiconductor diode modulator will be described capable of measuring dc currents down to 10^{-10} to 10^{-11} amperes. This modulator utilizes one diode although a second is normally installed for balance purposes in high sensitivity measurements. The modulator utilizes pulse techniques instead of sine waves which makes possible the high sensitivity. The circuit details of the modulator and design criteria will be discussed.

The Beam-Programming System of the Bevatron—Harry G. Heard (p. 4)

A brief description is given of the beam-programming equipment developed for the Bevatron. The system described provides analog-computer-controlled as well as beam-controlled frequency tracking. Equipment is described that permits arbitrary adjustment and control of the radial position of the circulating beam at any energy. Beam-intensity control and the automatic reduction of phase errors in the acceleration system are provided. Included also are units that produce long and short beam pulses. Equipment that permits highly multiple operation is also discussed. The integration of these equipments for extended injection is treated.

A Reactor Safety System Using Transistors and Silicon Controlled Rectifiers—William M. Trenholme (p. 14)

An all-semiconductor reactor safety system is described which consists of a dc amplifier, an adjustable trip level bistable circuit, a diode logic circuit, and a power switching circuit. The dc amplifier is a simple emitter follower transistor stage. Three transistors are used in the bistable circuit, two in a Schmitt-trigger arrangement and one as an output drive. The diode logic circuit uses a combination of OR and AND gates to give 2-out-of-3 coincidence of the bistable circuit outputs. A unijunction transistor and two silicon controlled rectifiers are the active elements in the power switching circuit. These components are particularly well suited for the on-off control needed in reactors and the circuitry is reasonably fail-safe.

The Oak-Ridge Thermonuclear Program—Recent Advances on the DCX—J. L. Dunlap (p. 19)

In the Oak Ridge thermonuclear experiment, 600-kev molecular ions of hydrogen or deuterium are injected into the median plane of a direct current magnetic mirror machine (DCX). A fraction of the molecular ions is dissociated ($H_2^+ \rightarrow H^+ + H^0$) as the ion beam passes through a high-energy carbon arc whose electrodes are located at each end of DCX. The magnetic field configuration is such that the 300-kev atomic ions thus created are trapped by virtue of the change in charge-to-mass ratio and enter a circular orbit concentric with the magnetic axis. The neutral atoms and the remaining undissociated molecular beam leave the machine.

Loss processes, principally charge exchange collisions suffered by the circulating ions, have thus far prevented achieving the appreciable plasma density required for observable fusion reactions. Plasma densities in the order of 10^{+10} ions/cm^3 and mean containment times of 10 to 12 msec have been observed.

The experimental program at the present time is one of diagnostic measurements on the trapped ions and their environment, and of building up the plasma density by increasing the trapping rate and containment time.

Trends in Nuclear Reactor Control—W. C. Lipinski and J. M. Harrer (p. 25)

Dual Phosphor Detectors—R. Monaghan and B. F. Wilson (p. 32)

A technique is described using two scintillation phosphors with different decay times mounted on a single photomultiplier tube. The scintillations from the two phosphors are then separated in a pulse length discrimination circuit. The dual phosphor detector is compact and is suitable for use in survey meters and well logging instruments to record different types of radiation simultaneously.

Energy Resolution Correction for Scintillation Spectrometers—Leonard B. Gardner (p. 36)

A computational scheme is presented for correcting the energy resolution of scintillation spectrometers. This scheme, unlike others previously published, appertains only to experimental data and does not rely on theoretical considerations of energy absorption within the scintillator. Applications to beta spectroscopy and analysis are included and a derivation of the ideal scintillation line and the various factors which in practice statistically modify the ideal shape is given in the Appendix.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*), London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Index to the Abstracts and References published in the PROC. IRE from February, 1959 through January, 1960 is published by the PROC. IRE, June, 1960, Part II. It is also published by *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) and included in the April, 1960 issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extensions and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1, England.

ACOUSTICS AND AUDIO FREQUENCIES

534.133-8 1
Sources of Sound in Piezoelectric Crystals—E. H. Jacobsen. (*J. Acoust. Soc. Am.*, vol. 32, pp. 949–953; August, 1960.) The production of sound in piezoelectric crystals is examined from the viewpoint of the inhomogeneous wave equation. The sound is generated by a gradient in the piezoelectric stress and, consequently, the free surfaces of a crystal are usually the most effective wave sources. The excitation of sound at microwave frequencies is also discussed.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears at the end of the Abstracts and References section.

534.2 2
The Behaviour of a Longitudinal Wave at the Boundary between Two Media—M. Peschel. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 221–226; January, 1960.) The transition from a solid elastic medium to a) a different solid medium, b) a nonviscous fluid, and c) a vacuum, is considered, and results are interpreted with the aid of electroacoustic analogues treated by quadripole theory.

534.2-14 3
Acoustic Measurements in Deep Water using the Bathyscaph—M. Lomask and R. Frassetto. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1028–1033; August, 1960.) A report of simple acoustic measurements made in the frequency range 10–500 cps to depths of 3200 meters is given. The capabilities and limitations of the bathyscaph as a vehicle for acoustic experiments are discussed.

534.2-14:534.88 4
Shallow-Water Under-Ice Acoustics in Barrow Strait—A. R. Milne. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1007–1016; August, 1960.) Recordings of ambient noise have been made in the frequency range 3 cps–10 kc. The results show several harmonically related spectrum lines and, in particular, a prominent line near 7 cps which appeared to have a wind-induced excitation.

534.23-814 5
The Transmission of Sound through a Wake—E. G. Richardson. (*Proc. Phys. Soc., London*, vol. 76, pp. 25–32; plate; July 1, 1960.) A description of measurements of the diffraction of an ultrasonic beam produced in passing through wakes of cylinders in water and air is given. The ultrasonic wavelength was small compared with the vortex size. The decay of the vortices with distance behind the cylinder is discussed.

534.232:538.652 6
Theoretical Investigations of the Complex Impedance of Magnetostrictive Extensional Ring Oscillators—A. Fraude. (*Arch. Elektrotech.*, vol. 44, pp. 399–418; December 19, 1959.)

534.373 7
Effect of Transport Processes on Attenuation and Dispersion in Aerosols—S. L. Soo. (*J. Acoust. Soc. Am.*, vol. 32, pp. 943–946; August, 1960.) The dispersion and attenuation of sonic and ultrasonic sound waves in aerosols have been studied, and the effects of momentum and heat transfer between the solid particles and the gas have been considered.

534.6 8
The Improvement of Objective and Subjective Acoustic Measurement Techniques by the Use of F.M. Band Noise—H. Niese. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 193–202; January, 1960.) A comparison of the results of measurements obtained with bands of white noise and with constant-amplitude FM noise of narrow bandwidth shows the advantage of the latter method in revealing greater detail in the measured characteristics.

534.614-14 9
Sound Velocity Measurement in Liquids—W. D. Wilson and D. D. Taylor. (*Electronics*, vol. 33, pp. 69–71; September 9, 1960.) An application of the method described by Greenspan and Tschiegy (1296 of 1958) is given. Two 5-Mc crystals are used with pulse duration 0.05 μsec. Sound velocity is determined from the pulse repetition frequency which is adjusted so that a new pulse is initiated each time a reflected pulse returns to the transmitter. Dissolved impurities down to 1 part in 10^6 can be detected.

534.76 10
Location Tests for Intensity-Type Stereophony—K. Wendt. (*Frequenz.*, vol. 14, pp. 11–14; January, 1960.) The results are analyzed of subjective tests to determine the apparent direction of the sound as a function of the sound-level difference of the radiator in a two-channel system. The dependence of this relation on frequency is also investigated.

534.79 11
The Inertia of Loudness Growth as a Function of Sound Level—H. Niese. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 143–152; December, 1959.) Subjective tests were made to investigate the dependence of the "inertia" effect in hearing (1426 of 1959) on sound level. The results are considered in relation to the design of objective loudness meters.

534.79 12
The Subjective Loudness Measurement of Band-Pass Noise with Uniformly and Differently Applied Bandwidths for Each Ear to Obtain Laws Governing Hearing and Their Simulation in Objective Loudness Measurements—H. Niese. (*Hochfrequenztech. und Elektroak.*, vol. 68, pp. 202–217; January, 1960.)

621.395.614 13
High-Frequency Calibration of an ADP Crystal Microphone—P. A. Macpherson and D. B. Thrasher. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1061–1064; August, 1960.) The free-field sensitivity of a microphone over the frequency range 10–100 kc is determined by the reciprocity technique, in which the transducers are oriented perpendicular to each other in order to minimize diffraction effects.

ANTENNAS AND TRANSMISSION LINES

621.315.212 14
Broad-Band Coaxial Choked Coupling Design—H. E. King. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 132–135; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1194; June, 1960.)

621.315.212:621.372.51 15
Analysis of Split Coaxial-Line-Type Balun—H. Kogō. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 245–246; March, 1960.)

621.315.212.018.44 16
The Skin Effect in Tubes and Coaxial Lines—W. Held and K. Wenzel. (*Arch. Elektrotech.*, vol. 44, pp. 306–317; September 1, 1959.) The skin effect is calculated for any frequency, using Maxwell's equations.

621.372.2 17
A Study of Multielement Transmission Lines—H. Kogō. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 136–142; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1194; June, 1960.)

621.372.2 18
Complementarity in the Study of Transmission Lines—G. H. Owyang and R. King. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 172–181; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.372.2:621.317.332.1 19
Investigation of Balanced High-Frequency Transmission Lines—K. Lauterjung. (*Arch. Elekt. Übertragung*, vol. 14, pp. 26–36; January, 1960.) The transfer characteristics of a balanced shielded line are investigated relating to its application as a measuring line. The importance of balancing input and output terminations is considered, and two types of measuring-line structure are examined. The design of a balanced measuring line with an impedance of 240 Ω for the frequency range 150–650 Mc is illustrated.

621.372.2:621.372.51 20
A General Theorem on an Optimum Stepped Impedance Transformer—H. J. Riblet. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 169–170; March, 1960. Abstract PROC. IRE, vol. 48, pt. 1, June, 1960.) See also 285 below, and 3594 of 1959 (Solymar).

621.372.2:621.372.51 21
Stepped Transformers for Partially Filled Transmission Lines—D. J. Sullivan and D. A. Parkes. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 212–217; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.372.2:621.396.679.4 22
The Goubau Line for Transmitter Applications: Properties and Operational Experience—F. R. Huber and H. Rudat. (*Rundfunktech. Mitt.*, vol. 3, pp. 277–283; December, 1959.) The uses of the Goubau line as a feeder for a VHF transmitter antenna is described.

621.372.821:621.372.832.6 23
Wide-Band Strip-Line Magic-T—E. M. T. Jones. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 160–168; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.372.823:621.372.852.22 24
Features of Cylindrical Waveguides containing Gyromagnetic Media—R. A. Waldron. (*J. Brit. IRE*, vol. 20, pp. 695–706; September, 1960.) The variations of electromotive field components and power flow are presented as functions of position in the transverse plane for the H_{11} mode in a waveguide of radius a containing a concentric ferrite rod of radius b. Values of b/a from 0 to 1 are taken, and results are discussed in relation to phase-constant curves previously obtained (341 of 1959).

621.372.825 25
A Contribution to the Theory of Corrugated Guides—G. Piefke. (*J. Res. NBS*, vol. 64D, pp. 533–555; September/October, 1960.) The results of three previous papers (3035 of 1957, 2625 of 1958 and 1075 of 1959) are summarized, and the mathematical method for calculating the transmission characteristics of corrugated guides is briefly recapitulated.

621.372.825:621.372.852.323 26
Nonreciprocal Attenuation of Ferrite in Single-Ridge Waveguide—T. S. Chen. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, no. 2, pp. 247–248; March, 1960.) Three types of isolators are investigated; resonance-absorption and field-displacement types, and isolators operating in low-bias magnetic fields.

621.372.829:621.374.5 27
The Behaviour of the Helical Line as a Delay Line with Particular Regard to Attenuation and Characteristics in Rotating Fields—G. Piefke. (*Arch. elekt. Übertragung*, vol. 14, pp. 15–25; January, 1960.) The theory of the helix waveguide developed in 2459 of 1959 is applied to the case of the delay line, taking account of the finite conductor thickness and assuming the wavelength in the guide to be much greater than the spacing between turns.

621.372.82:621.385.632.12 28
Experimental Investigation of the Efficiency of a Helix Waveguide—G. A. Silenck, A. K. Berezin, P. M. Zeidlits and A. M. Nekrashevich. (*Zh. Tekh. Fiz.*, vol. 29, pp. 946–963; August, 1959.) The frequency dependence of the efficiency of a helix waveguide as a slow-wave structure in the range of 150–800 Mc has been determined. Results of measurements are in agreement with theoretical calculations made using a sheath model.

621.372.832.4 29
Resonant Modes in Waveguide Windows—M. P. Forrer and E. T. Jaynes. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 147–150; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.372.832.6 30
The Magic T—W. Stösser. (*Frequenz*, vol. 14, pp. 17–19; January, 1960.) The properties of waveguide hybrid junctions and their matching conditions are obtained with the aid of the scattering matrix.

621.372.852.12:621.317.34 31
Half-Round Inductive Obstacles in Rectangular Waveguide—D. M. Kerns. (*J. Res. NBS*, vol. 64B, pp. 113–130; April–June, 1960.) Formulas for the energy reflected by one or two opposed semicircular cylindrical indentations in the narrow walls appear accurate enough for the configuration to be used as a calculable standard of reflection or impedance.

621.372.852.323 32
On the Theory of the Ferrite Resonance Isolator—E. Schlömann. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 199–206; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.372.853:621.385.632.19 33
Nonperiodic Slow-Wave and Backward-Wave Structures—P. J. B. Clarricoats and R. A. Waldron. (*J. Electronics and Control*, vol. 8, pp. 455–458; June, 1960.) "The behaviour of the phase constant as a function of frequency is examined in the neighbourhood of cut-off for a waveguide of circular cross-section containing a concentric rod of dielectric or ferrite material. It is predicted from this behaviour that with suitable choices of geometry and properties of the rod, slow waves or backward waves will propagate."

621.396.67 34
Cylindrical Antenna Theory—R. H. Duncan and F. A. Hinchey. (*J. Res. NBS*, vol. 64E, pp. 569–584; September/October, 1960.) A survey of cylindrical-antenna theory relating to a tubular model with a narrow gap is given. Numerical results are presented for two typical examples.

621.396.67.095 35
Uniqueness of the Solution to Integral Equations of the First Kind in the Theory of Aerials—N. N. Govorun. (*Dokl. Akad. Nauk SSSR*, vol. 132, pp. 91–94; May 1, 1960.) Fredholm's integral solutions of the first kind are obtained for the current density of an antenna represented as a body of revolution with a surface impedance. The uniqueness of solution is analyzed in the case of perfectly-conducting antennas.

621.396.67.095(204) 36
Field Strength Measurements in Fresh Water—G. S. Saran and G. Held. (*J. Res. NBS*, vol. 64D, pp. 435–437; September/October, 1960.) Depth/attenuation figures derived from experimental observations to 1000 feet at 18.6 kc are compared with theoretical values.

621.396.674.3 37
Ways towards the Optimum Wide-Band Omnidirectional Radiator—W. Stöhr and O. Zinke. (*Frequenz.*, vol. 14, pp. 26–35; January, 1960.) Locus curves are given and reflection coefficients are plotted for ellipsoidal, spherical and conical radiators of various length/diameter ratios, with a conical transition to the coaxial feeder. The effects of the shape of the transition section and the proportions of the radiator on matching and radiation characteristics are discussed. Optimum results are obtained with a modified hemispherical radiator with specially shaped conical transition.

621.396.677 38
Experimental and Theoretical Investigations of Plane Surface Aerials—S. Blume. (*Z. angew. Phys.*, vol. 12, pp. 39–47 and 72–87; January and February, 1960.) The field theory of surface antennas without rotational symmetry is developed, and experimentally obtained radiation diagrams are given and discussed.

621.396.677.3 39
Digital Instrumentation of Antenna Measurements—E. K. Damon. (*Electronics*, vol. 33, pp. 90–93; October 14, 1960.) A test system, incorporating a digital computer, is described for measurements of amplitude and differential phase shift of the signal at a number of points along an end-fire array.

621.396.677.4 40
The Triangular "V" Aerial—E. O. Willoughby. (*Proc. IRE, Australia*, vol. 21, pp. 517–523; August, 1960.) In the antenna described, a smooth transition from transmission line to antenna aperture is achieved using triangular plates. Results obtained with experimental UHF and HF antennas show a 3 or 4:1 bandwidth with good impedance matching, no beam splitting and gains at HF of 18 db.

621.396.677.45 41
Equivalence of 0 and −1 Space Harmonics in Helical Antenna Operation—T. S. Maclean and D. A. Watkins. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, p. 251; March, 1960.) Identical results are obtained whether the fundamental or the −1 space harmonic is considered, the first relating to propagation along the conductor and the second to propagation axially.

621.396.677.6:621.396.663 42
Theory of the General Adcock—K. Baur. (*Arch. elekt. Übertragung*, vol. 14, pp. 1–14 and 57–70; January and February, 1960.) A comprehensive theoretical treatment of the stationary Adcock direction finder is given. Various practical problems of Adcock design and performance are solved with the aid of the basic concept discussed.

621.396.677.71:537.56 43
Radiation and Admittance of an Insulated Slotted-Sphere Antenna Surrounded by a Strongly Ionized Plasma Sheath—J. W. Marini. (*J. Res. NBS*, vol. 64D, pp. 525–532; September/October, 1960.) The performance of a slotted sphere is computed from the voltage distribution along the slot, and the optimum operational frequency is obtained.

621.396.677.73 44
Wide-Band Matching of Horn-Parabola Aerials—H. Laub. (*Frequenz*, vol. 13, pp. 390–397; December, 1959.) The design is given of a junction between circular waveguide and a horn-parabola antenna [2628 of 1956 (Laub and Stöhr)]. This design permits antenna operation over the range 3.6–11.7 Gc with a reflection factor <1 per cent.

621.396.677.83:621.396.65 45
Optimum Design of Microwave Periscopes for Radio Link Installations—P. Münzer. (*Telefunken Ztg.*, vol. 32, pp. 269–278; December, 1959. English summary, pp. 293–294.) The design is detailed of a radio-link antenna system, in which the curved transmitter antenna is pointed upwards so that its radiation impinges on a plane reflector mounted on a tower, which deflects the radiation into the horizontal plane. See, *e.g.*, 38 of 1960 (Possenti).

621.396.677.831 46
Investigation of the Reflection Characteristics of Plane Reflectors in the Decimetre Range—E. Düniss and K. E. Müller. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 185–190; January, 1960.) A report is given of experimental investigations at 23.8 cm λ with an array of four independently-adjustable square metal mirrors illuminated by a transmitter at a distance of 5.75 km. The radiation diagrams obtained are compared with calculated characteristics.

621.396.677.832 47
High-Gain, Very-Low-Side-Lobe Antenna with Capability for Beam Slewing—A. C. Wilson. (*J. Res. NBS*, vol. 64D, pp. 557–561; September/October, 1960.) Theoretical results for the radiation patterns of a large corner-reflector antenna are compared with practical measurements with the main beam 10° off the forward direction. Close agreement is obtained.

621.396.677.85:621.396.965 48
Properties of Phased Arrays—W. H. Von Aulock. (Proc. IRE, vol. 48, pp. 1715–1727; October, 1960.) Multi-element planar and linear arrays of equally spaced radiators are considered. The effect of scanning, by the introduction of phase delay, on the width, shape and direction of the main beam and on the sidelobe pattern are calculated, making use of a projection of the unit sphere on the plane of the array rather than two orthogonal planes.

AUTOMATIC COMPUTERS

681.142 49
Large-Storage Binary-to-Decimal Converter—G. Dearnaley, L. G. Lawrence and H. C. Tresise. (*J. Sci. Instr.*, vol. 37, pp. 240–242; July, 1960.) A digital converter is described which will handle 24-binary-digit numbers with a maximum input speed of 4 bits per second.

681.142 50
"Isabel" (Iso Status Accumulating Binaries using Extraordinary Logic)—J. A. Goss. (*Electronic Engrg.*, vol. 32, pp. 630–634; October, 1960.) "An improved logical arrangement of a decimal counter based on bistable elements is described. The advantages over usual methods are described, and the performance of various circuits using this method is given."

681.142 51
Input and Output Devices for Computers—(*J. Brit. IRE*, vol. 20, pp. 657–683; September, 1960.) The following papers were presented at a symposium held in London on November 4, 1959.

a) The Transport of Paper Tape in Digital Computation—A. D. Booth (pp. 657–660).

b) A High-Speed Tape Reader—R. D. Lacy (pp. 661–668).

c) A New 600-Cards-per-Minute Card-Reader—H. H. G. Groom (pp. 669–674).

d) High-Speed Printers—W. A. J. Davie (pp. 675–683).

681.142:621–526 52
Analyzing Magnetically Detented Stepper Servo Motors—Weber and Weiss. (See 335.)

681.142:621.318.57:539.23 53
A Vacuum-Evaporated Random-Access Memory—K. D. Broadbent. (Proc. IRE, vol. 48, pp. 1728–1731; October, 1960.) The basic multiple-layer magnetic-film structure described is shown to have valuable properties in coincident-signal switching applications, including low magnetic-turnover time and extremely small cell volume.

681.142:621.398 54
A Special-Purpose Analogue-Digital Converter—G. C. Henderson. (*Electronic Engrg.*, vol. 32, pp. 602–608; October, 1960.) This converter is for use in analyzing tape recordings of time-multiplex telemetry signals. One channel is analyzed at each playback, and known transducer nonlinearities and drifts of gain and level are compensated.

CIRCUITS AND CIRCUIT ELEMENTS

621.316.825.4 55
Therm stors, their Theory, Manufacture and Application—R. W. A. Scarr and R. A. Setterington. (*Proc. IEE*, vol. 107, pt. B, pp. 395–405; September, 1960. Discussion, pp. 405–409.) 80 references.

621.318/.319:621.372.22 56
Line Equations for an Inhomogeneous Line System of Cylindrical Symmetry and their Application to Circuit Elements—H. Heywang. (*Frequenz*, vol. 13, pp. 397–400; December, 1959.) Examples considered include the calculation of the skin effect of cylindrical wires, the frequency response of disk-type and wound capacitors, and the impedance of a sintered-Ta electrolytic capacitor.

621.318.57:621.382.3 57
The Realization of Switches for Both Current Directions using Junction Transistors—W. Hilberg. (*Elektron. Rundschau*, vol. 13, pp. 438–440; December, 1959.) Some practical transistor arrangements with and without diodes are described.

621.318.57:621.382.3 58
Static Switching—G. T. Ohlsen. (*Electronic Engrg.*, vol. 32, pp. 609–613; October, 1960.) A transistor NOR unit (output only when all inputs absent) is shown to be suitable as a basic element for building switching systems.

621.372.4/.4 59
The Practical Significance for Communication Technique of the Relation between Hilbert Transformation and Harmonic Analysis—G. Wunsch. (*Arch. elekt. Übertragung*, vol. 13, pp. 503–508; December, 1959.) Conformal mapping is used to show that the Hilbert transformation can be reduced to a harmonic analysis which would simplify attenuation and phase calculations. Practical methods of carrying out the transformation are outlined.

621.372.412 60
Vibration of Quartz Crystal Plates—R. P. Jerrard. (*Quart. Appl. Math.*, vol. 18, pp. 173–181; July, 1960.) The theory of Mindlin (1861 of 1951) is extended to the case of crystal plates of nonuniform thickness.

621.372.412 61
Design and Performance of Ultraprecise 2.5-Mc/s Quartz Crystal Units—A. W. Warner. (*Bell Sys. Tech. J.*, vol. 39, pp. 1195–1217; September, 1960.) Techniques are described for producing crystal resonators with a frequency stability within about 1 part in 10^{10}, and a Q of $5-6\times10^6$. For periods up to 1 month, the frequency stability compares favorably with that of atomic frequency standards.

621.372.412.002 62
On the Electrodes of 100 kc/s GT-Cut Crystal Units used for the Frequency Standards—Y. Hiruta. (*J. Radio Res. Labs. Japan*, vol. 7, pp. 245–261; May, 1960.) The physical properties and method of fabrication of electrodes for crystals used in primary frequency standards are described. A maximum Q is obtained when the gold-sputtered electrode has a thickness of 350–450Å; advantages of annealing are discussed.

621.372.413 63
Temperature Compensation of Coaxial Cavities—J. R. Cogdell, A. P. Deam and A. W. Straiton. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 151–155; March, 1960.) Abstract, Proc. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.372.413:621.317.763 64
Increasing the Frequency Stability of Homogeneous Cylindrical Cavity Resonators—H. J. Oberg. (*Telefunken Ztg.*, vol. 32, pp. 265–268; December, 1959. English summary, p. 293.) The effect of temperature fluctuations and atmospheric conditions on the resonator dimensions and the characteristics of the dielectric investigated to determine means of minimizing frequency variations.

621.372.413:621.372.8 65
How to Design Broadband Microwave Cavities—A. Kiriloff. (*Electronics*, vol. 33, No. 42, pp. 96–100; October 14, 1960.) Design criteria for rhumbatrons of cylindrical and prismatic form are given with a numerical example for each type.

621.372.5 66
Analysis of All-Pass Networks: Part 1—General Remarks and Classification—G. Wunsch. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 169–176; January, 1960.)

621.372.5 67
The Theory of the Realization of Linear Quadripole Networks with Prescribed Effective Transfer Function in the Presence of Lossy Circuit Elements—Nai-Ta Ming. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 190–193; January, 1960.) A summary by the author of the results contained in a more extensive paper published in Chinese is given.

621.372.5 68
The Differences between Single and Periodic Transient Processes in Low-Pass Transmission Systems—R. Elsner and K. H. Steiner. (*Nachrichtentech. Z.*, vol. 12, pp. 618–624; December, 1959.) The differences are calculated for square-pulse waveforms and ideal transmission systems; corrections can be made to obtain approximations applicable to practical low-pass networks.

621.372.51 69
Synthesis Techniques for Gain-Bandwidth Optimization in Passive Transducers—H. J. Carlin. (Proc. IRE, vol. 48, pp. 1705–1714; October, 1960.) The technique produces from the idealized gain-bandwidth constraints, which ensure matching to a given load a set of functions specifying the equalizer alone; the fixed transducer is synthesized from these functions. Examples illustrating the technique are given.

621.372.54 70
An Analogue Device for Problems of Filter Development—W. Poschenrieder and H. Sontheim. (*Frequenz*, vol. 13, pp. 379–385; December, 1959.) The equipment described is used for the design of filters from given operating characteristics. Other applications, such as filter correction and filter design for pulse waveforms, are outlined.

621.372.54 71
Ladder Networks with Tchebycheff Characteristic—M. Trinchieri. (*Alta Frequenza*, vol. 28, pp. 541–578; October–December, 1959.) Ladder networks without poles are considered with Tchebycheff transfer function for inputs from a generator with finite internal impedance, or from ideal current or voltage generators. Design formulas are tabulated.

621.372.54 72
An Alternative Approach to the Realization of Network Transfer Functions: the *n*-Path Filter—L. E. Franks and I. W. Sandberg. (*Bell Sys. Tech. J.*, vol. 39, pp. 1321–1350; September, 1960.) A time-varying network consisting of *n* parallel transmission paths, each containing input and output modulators, is described and analyzed. Some practical applications are discussed in detail, and experimental verification is presented.

621.372.54:621.316.825 73
Low-Frequency Wave Filters employing Thermistors—R. A. Rasmussen. (*Rev. Sci. Instr.*, vol. 31, pp. 747–751; July, 1960.) A general review of thermistor properties and small-signal performance is presented, together with results of measurements at very-low frequencies and response curves of simple filter circuits.

621.372.54:621.396.96 74
Realization of a Comb Filter for the Detection of Pulse Signals in Noise—A. Garino. (*Note Recensioni Notiz.*, vol. 9, pp. 43–60; January/February, 1960.) The use of comb filters in radar systems is discussed, and the results obtained on an experimental-filter system with up to seven pass bands are given.

621.372.542 75
The Correction of Phase Distortion in *RC* Networks near Cut-Off Frequency—D. Gossel. (*Arch. elekt. Übertragung*, vol. 13, pp. 525–529; December, 1959.) Examples of phase-correcting networks and optimum design conditions are given.

621.372.543.2:538.652 76
Mechanical Filters—H. D. Pieper. (*Telefunken Ztg.*, vol. 32, pp. 279–283; December, 1959. English summary, p. 294.) See also 2155 of 1959 (Borner, *et al.*).

621.372.543.3:621.372.55 77
Band-Stop Filters—R. O. Rowlands. (*Electronic Tech.*, vol. 37, pp. 431–433; November, 1960.) "A method of achieving resistance compensation in band-stop filters is described. Performance curves and normalized component values are calculated for a 3-pole filter."

621.372.55 78
High Selectivity with Constant Phase over the Pass Band—A. W. Rihaczek. (Proc. IRE, vol. 48, pp. 1756–1760; October, 1960.) A relatively simple frequency-inversion process is described for compensation of phase shifts in filters, resulting in a constant output signal phase.

621.372.6 79
Equivalent Networks of Generic Multipoles—G. Biorci and L. Piglione. (*Alta Frequenza*, vol. 28, pp. 528–540; October–December, 1959.) A nonreciprocal four-terminal network can be realized by an equivalent circuit which contains only one nonreciprocal three-pole element.

621.373.42.072.6:621.385.62 80
A New Automatic Frequency Regulation System—J. R. Singer. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, p. 249, March, 1960.) A brief description is given of a system for stabilizing the output of a K-band klystron. The error signal is derived from a phase comparison of a 100-kc driving signal with the rectified microwave signal reflected from a dimension-modulated reference cavity. The resonance frequency of the cavity is controlled by a driven Rochelle-salt crystal.

621.373.42.072.7 81
Automatic Phase Control: Theory and Design—T. J. Rey. (Proc. IRE, vol. 48, pp. 1760–1771; October, 1960.) Systems with a sinusoidal reference signal are analyzed and the results verified experimentally.

621.373.44 82
Variable-Program Triggering Source—B. E. Bourne. (*Electronics*, vol. 33, pp. 76–77; September 9, 1960.) A transistorized unit for use with instrumentation cameras is described which generates pulses at an adjustable constant rate or at a rate which is variable for a selected period between predetermined initial and final rates.

621.373.444:621.382.3 83
A Special Trigger Circuit—C. Mira and Y. Sevely. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 488–490; January 18, 1960.) The bistable transistor circuit with common-emitter resistance analyzed earlier [1139 of 1960 (Mira)] has characteristics similar to those of the Schmitt multivibrator circuit. Two variations of the circuit component values are considered, corresponding to "saturating" and "nonsaturating" conditions.

621.374.3:621.382.23 84
High-Speed Scalers using Tunnel Diodes—P. Spiegel. (*Rev. Sci. Instr.*, vol. 31, pp. 754–755; July, 1960.) A simple reliable scaler with resolving time <50 nsec is described.

621.374.4:621.382.3 85
A Cascode Trigger Circuit using a *p-n-p* and *n-p-n* Transistor—P. Arnoldt. (*Electronic Engrg.*, vol. 32, pp. 620–623; October, 1960.) A monostable circuit preceded by a diode-pump integrator acts as a frequency dividing stage.

621.374.5:534.2-8-14 86
Variable Ultrasonic Water Delay Line—G. W. Williams. (*Brit. J. Appl. Phys.*, vol. 11, pp. 358–363; August, 1960.) The unit described produces a delay variable from 200–2600 μsec, and stable to within 0.05 per cent; the bandwidth is 800 kc at 7 Mc.

621.375.1 87
The Noise Figure of Several Amplifiers in Cascade—A. Ascione. (*Note Recensioni Notiz.*, vol. 8, pp. 620–630; November/December, 1959.)

621.375.9:[538.569.4+621.372.44 88
Parametric Devices and Masers: an Annotated Bibliography—E. Mount and B. Begg. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 222–243; March, 1960.)

621.375.9:538.569.4:621.374.42 89
Parametric Diodes in Maser Phase-Locked Frequency Divider—M. L. Stitch, N. O. Robinson and W. Silvey. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 218–221; March, 1960. Abstract, Proc. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.375.9:621.372.44 90
Contribution to the Study of Parametric Amplifiers—G. Marie and Y. Angel. (*C. R. Acad. Sci., Paris*, vol. 250, pp. 311–313; January 11, 1960.) An analysis of parametric amplifier operation in terms of feedback loops is given. Expressions are derived for the insertion gain, noise temperature and bandwidth of different 2- and 3-frequency arrangements, which can be generalized for a 4-frequency system.

621.375.9:621.372.44 91
The Transmission of Signals of Any Form by means of a Degenerate Parametric Amplifier—F. Bertein and A. Jelenski. (*CR. Acad. Sci., Paris*, vol. 250, pp. 88–90; January 4, 1960.) Formulas are derived for the response of a degenerate parametric amplifier comprising an antiresonant circuit with frequency-dependent admittance in parallel with a nonlinear-capacitance diode.

621.375.9:621.372.44:621.382.2 92
Theory of Diode Parametric Amplification—F. Dachert. (*Ann. Radioélect.*, vol. 15, pp. 109–119; April, 1960. English abridgment, pp. 200–201.)

621.375.9:621.372.44:621.382.23 93
The Reactance-Type Straight Amplifier as a Low-Noise Preamplifier Stage in the U.H.F. Range—R. Maurer, K. H. Löcherer and K. Bomhardt. (*Arch. elekt. Übertragung*, vol. 13, pp. 509–524; December, 1959.) The signal and noise response of a cascade-connected amplifier is investigated which consists of a di-

ode-reactance preamplifier [see 3426 of 1960 (Dahlke, *et al.*)] followed by a conventional triode amplifier. Results are given of measurements on an experimental circuit using a signal frequency of 510 Mc with pumping at 1800 Mc.

621.375.9:621.372.44:621.385.63 94
The D.C.-Pumped Quadrupole Amplifier—a Wave Analysis—Siegman. (See 377.)

621.376.2+621.376.3 95
A Contribution to the Theory of the Transmission of Amplitude-Modulated, Frequency-Modulated or Amplitude- and Frequency-Modulated Carriers in Linear Systems: Part 1—E. Augustin. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 177–185; January, 1960.) The standard method of calculating the distortion of FM signals in linear systems, which is based on an asymptotic series, is critically reviewed and its limitations are considered.

GENERAL PHYSICS

535.215 96
Photovoltaic Effect Derived from the Carnot Cycle—A. Rose. (*J. Appl. Phys.*, vol. 31, pp. 1640–1641; September, 1960.) An expression for the maximum power derivable from any photovoltaic device is developed in terms of the incident light intensity measured in units of black-body radiation.

537.29:537.56 97
Velocity Distribution of Electrons in a Strong Electric Field—L. M. Kovrizhnykh. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 1394–1400; November, 1959.) A method is developed for determination of a nonstationary solution of the Boltzmann equation in the case of strong electric fields.

537.29:621.085.032.212 98
Field Emission through Dielectric Layers—R. Gomer. (*Aust. J. Phys.*, vol. 13, pp. 391–401; July, 1960.) A summary is given of the results of field emission experiments on the adsorption of inert gases on tungsten, and a mechanism is described accounting for some of the phenomena encountered.

537.311.33 99
The Spatial- and Time-Fluctuation of Coulomb Energy in the Low-Concentration Impurity Conduction—Y. Toyozawa. (*Prog. Theoret. Phys.*, vol. 23, pp. 380–382; February 1960.) See 223 below.

537.312.8 100
Electron Scattering in High Magnetic Field—A. H. Kahn. (*Phys. Rev.*, vol. 119, pp. 1189–1192; August 15, 1960.) Electrical conductivity in a strong magnetic field is calculated for the case of scattering by delta-function impurities. Collision broadening is neglected, and the scattering by an individual center is solved exactly rather than by perturbation theory.

537.52 101
Microwave Breakdown of Air in Nonuniform Electric Fields—P. M. Platzman and E. H. Solt. (*Phys. Rev.*, vol. 119, pp. 1143–1149; August 15, 1960.) Two cases are treated theoretically: a rectangular microwave cavity and a similar cavity with a hemispherical boss on one of its walls. The theoretical predictions are compared with breakdown measurement in air.

537.52:551.510.52 102
High-Frequency Breakdown of Air—D. Kelly and H. Margenau. (*J. Appl. Phys.*, vol. 31, pp. 1617–1620; September, 1960.) Kinetic theory is applied to the ionization breakdown of air surrounding an antenna on a high-speed vehicle. Breakdown-voltage/altitude curves are given for 225 Mc, 1 Gc, and 10 Gc.

537.56 103
Electrical and Thermal Currents in a Slightly Ionized Gas—M. S. Sodha. (*Phys. Rev.*, vol. 119, pp. 882–886; August 1, 1960.) A general solution is given of the Boltzmann transfer equation for electrons under the influence of ac and dc electric fields, a magnetic field and a temperature gradient.

537.56:538.56 104
Nonlinear Effects on Electron-Plasma Oscillations—M. Sumi. (*J. Phys. Soc. Japan*, vol. 15, pp. 1086–1093; June, 1960.) A nonlinear term in the Boltzmann equation is considered as a perturbation, and the results are examined to investigate the validity of the linear approximation.

537.56:538.56 105
Forces on Charged Particles of a Plasma in a Cavity Resonator—J. W. Gallop, T. L. Dutt and H. Gibson. (*Nature*, vol. 188, pp. 297–298; October 29, 1960.) Two types of confining force are considered: a) the "Mathieu" force due to the electric field acting directly on electrons [see Boot, *et al.* (3411 of 1958)], and b) the electromotive force resulting from the interaction of the electron current with the magnetic field component of the electromotive wave.

537.56:538.56 106
Radio-Frequency Forced Oscillations in Nonuniform Plasmas—R. B R. Shersby-Harvie. (*J. Electronics and Control*, vol. 8, pp. 421–430; June, 1960.) A theory of periodic RF oscillations in a plasma is given, neglecting electron-ion collisions and temperature variations. Small-signal approximations are derived which are applicable to plasma confinement problems.

537.56:538.569.4 107
Plasma Resonance in a Radio-Frequency Probe—K. Takayama, H. Ikegami and S. Miyazaki. (*Phys. Rev. Lett.*, vol. 5, pp. 238–240; September 15, 1960.) Measurements of the dc component of the electron current to the probe show a resonance increase at the electron plasma frequency.

537.56:538.63 108
Negative Electrical Conductivities—J. Schneider. (*Z. Naturforsch.*, vol. 15a, pp. 484–489; May/June, 1960.) The complex electrical HF conductivity σ of a plasma in a magnetic field is calculated on the basis of Boltzmann's equation, and the occurrence of negative values of the real part of σ is discussed.

537.56:538.63 109
Oscillations in a Plasma in a Weak Magnetic Field—K. Kato. (*J. Phys. Soc. Japan*, vol. 15, pp. 1093–1101; June. 1960.) Oscillations of two types were observed in a discharge tube. These are interpreted as the plasma-type oscillation and the cyclotron-resonance oscillation.

537.56:538.63 110
Oscillations of an Inhomogeneous Plasma in a Magnetic Field—L. I. Rudakov and R. Z. Sagdeev. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 1337–1341; November, 1959.) A mathematical analysis of small oscillations of a hot plasma confined by a magnetic field shows that there are two types of wave: a) a slow drift wave, characteristic of an inhomogeneous plasma, and b) a magnetoacoustic wave. If certain relations obtain, the drift current can cause amplification of the oscillations; criteria for an instability of this kind are derived.

537.56:538.63 111
Transport Properties of Plasmas in a Strong Magnetic Field—T. Kihara, Y. Midzuno and S. Kaneko. (*J. Phys. Soc. Japan*, vol. 15, pp. 1101–1107; June, 1960.)

537.56:538.63 112
Compression Waves in a Plasma in a Static Magnetic Field—N. Anderson. (*Proc. Phys. Soc., London*, vol. 75, pp. 905–912; June 1, 1960.)

537.56:551.510.535 113
Ion Charge Exchange Reactions in Oxygen Afterglows—P. H. G. Dickinson and J. Sayers. (*Proc. Phys. Soc., London*, vol. 76, pp. 137–148; July 1, 1960.) Measurements made by a linear-accelerator mass spectrometer of the decay of O^+ ion concentration after a pulsed discharge are discussed with reference to electron loss processes in the upper ionosphere.

538.221:061.3 114
Summarized Proceedings of a Conference on Some Aspects of Magnetism, Sheffield, September 1959—R. S. Tebble and D. E. G. Williams. (*Brit. J. Appl. Phys.*, vol. 11, pp. 307–313; August, 1960.)

538.221:538.569.4 115
Generation of Phonons in High-Power Ferromagnetic-Resonance Experiments—E. Schlömann. (*J. Appl. Phys.*, vol. 31, pp. 1647–1656; September, 1960.) Magneto-elastic waves are produced by coupling between spin waves and elastic vibrations.

538.221:621.318.134:535.37 116
Combinational Scattering of Electromagnetic Waves in Ferromagnetic Dielectrics—F. G. Bass and M. I. Kaganov. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 1390–1393; November, 1959.) Raman-type scattering by oscillations of the magnetic moment is predicted. The extinction coefficient for the scattered radiation is calculated.

538.5 117
The Influence of the Conductivity of the Surrounding Medium on the Input Impedance of a Current Loop—V. G. Zernyatko and D. N. Chetaev. (*Zh. Tekh. Fiz.*, vol. 29, pp. 1009–1013; August, 1959.) Using Fock's solution for the field of a low-frequency current element lying on the surface of a conducting medium, the input impedance of a current loop is calculated by the method of induced EMF's.

538.521:621.3.013.78 118
Shielding of Transient Electromagnetic Signals by a Thin Conducting Sheet—N. R. Zitron. (*J. Res. NBS*, vol. 64D, pp. 563–567; September/October, 1960.) The screening effects of an imperfectly conducting sheet against the transient field of a vertical magnetic dipole with a ramp-function excitation are investigated.

538.561:537.533 119
Megavolt Electronics Cherenkov Coupler for the Production of Millimetre and Submillimetre Waves—P. D. Coleman and C. Enderby. (*J. Appl. Phys.*, vol. 31, pp. 1695–1696; September, 1960.) Cherenkov radiation at the watt level can be produced and beamed by passing a high-harmonic 1-Mev electron beam through a 4-mm hole in a large specially shaped teflon cone. The measured radiated power shows good agreement with theory.

538.541:539.112:537.56 120
Field of a Charged Particle in a Moving Medium—B. M. Bolotovskiĭ and A. A. Rukhadze. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 1346–1351; November, 1959.) Energy losses of the charge due to Cherenkov radiation and excitation of plasma waves are determined.

538.566 121
Wide-Band Absorbers for Electromagnetic Waves—J. Deutsch and P. Thust. (*Z. angew. Phys.*, vol. 11, pp. 453–455; December, 1959.) The construction of shaft-type absorbers suitable for meter waves is described. Measurements of reflection coefficient as a function of frequency and angle of incidence are discussed, and the possibilities of reducing the depth of the absorber and of extending the absorption range to lower frequencies are mentioned.

538.566:535.42 122
The Diffraction of Electromagnetic Waves by a Circular Cylinder in a Homogeneous Half-Space—B. P. D'yakonov. (*Izv. Akad. Nauk SSSR, Ser. geofiz.*, pp. 1332–1343; September, 1959.) A solution is obtained for the distortion by a cylindrical inhomogeneity below the surface of the earth of an electromotive field produced by a point source.

538.566:535.43:537.56 123
Scattering of Electromagnetic Waves from an Infinitely Long Magnetized Cylindrical Plasma—P. M. Platzman and H. T. Ozaki. (*J. Appl. Phys.*, vol. 31, pp. 1597–1601; September, 1960.) The magnetically contained plasma is characterized in terms of its macroscopic dielectric tensor, and the scattering problem is solved analytically. Numerical results are given.

538.566:535.43:537.56 124
Scattering of Radiation by a Plasma—F. I. Boley. (*J. Appl. Phys.*, vol. 31, p. 1692; September, 1960.) A more complete analysis than that given by Herschberger (1953 of 1960) is necessary.

538.566:537.56 125
On Electromagnetic Radiation in Magneto-ionic Media—H. Kogelnik. (*J. Res. NBS*, vol. 64D, pp. 515–523; September/October, 1960.) A "wave matrix" is defined, the zeros of whose determinant are the propagation constants of the ordinary and the extraordinary plane waves.

538.566:537.56 126
Doppler Effect in an Electron Plasma in a Magnetic Field—K. A. Barsukov and A. A. Kolomenskiĭ. (*Zh. Tekh. Fiz.*, vol. 29, pp. 954–957; August, 1959.) A note on the complex Doppler effect occurring in an anisotropic or gyrotropic medium is given.

538.566.029.6:537.56 127
Propagation of Microwaves through a Magnetoplasma, and a possible Method for Determining the Electron Velocity Distributions—A. L. Cullen. (*J. Res. NBS*, vol. 64D, pp. 509–413; September/October, 1960.) A study is made of propagation of circularly polarized waves parallel to the magnetic field, giving special attention to the extraordinary ray.

538.569.2 128
Boundary Conditions and Ohmic Losses in Conducting Wedges—R. M. Chisholm. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 189–198; March, 1960. Abstract, PROC. IRE, vol. 48, pt 1, p. 1195; June, 1960.)

538.569.4 129
Modulation-Effect Corrections for Moments of Magnetic-Resonance Line Shapes—K. Halbach. (*Phys. Rev.*, vol. 119, pp. 1230–1233; August, 15, 1960.)

538.691 130
Electromagnetic Pressure on a Charge Moving in a Magnetic Field—Ya. B. Fainberg and V. I. Kurilko. (*Zh. Tekh. Fiz.*, vol. 29, pp. 939–945; August, 1959.) The radiation pressure on an oscillating particle is considerably greater than that on a free charge. The problem is considered in a nonlinear approximation and the equations of motion are discussed.

539.2:538.1 131
Spin Configurations—F. Bertaut. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 85–87; January 4, 1960.) A direct and a Fourier method are outlined for determining the configurations of ordered spin arrangements.

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523.16 132
Net Electric Charges on Stars, Galaxies and "Neutral" Elementary Particles—V. A. Bailey. (*J. Roy. Soc. N.S.W.*, vol. 94, pp. 77–86; September 9, 1960.) See 2324 of 1960.

523.164 133
The Spectrum of the Cygnus (19N4A) and Cassiopeia (23N5A) Radio Sources below 30 Mc/s—A. C. B. Lovell and H. W. Wells. (*Mon. Not. Roy. Astron. Soc.*, vol. 121, no. 1, pp. 111–114; 1960.) The 250-foot steerable radio telescope and the 220-foot transit radio telescope at Jodrell Bank have been used as an interferometer to measure the ratio of the intensities of the Cygnus and Cassiopeia radio sources from 16 to 26 Mc. The ratio remains constant at 0.53. Results are compared with those obtained by other workers.

523.164:621.396.677 134
The New Radio Astronomy Centre of the University of Sidney—H. Messel. (*Nature*, vol. 188, pp. 528–529; November 12, 1960.) General information is given about the Mills Cross radio telescope which is to be erected at the new center.

523.164.3 135
Radio Detection of the Planet Saturn—J. J. Cook, L. G. Cross M. E. Bair and C. B. Arnold. (*Nature*, vol. 188, pp. 393–394; October 29, 1960.) A brief report is given of the detection of 3.45-cm radiation from Saturn, using an X-band ruby maser in a modified Dicke-type radiometer system installed on an 85-ft paraboloid. A peak antenna temperature of $0.095 \pm 0.02°$K was obtained indicating an equivalent black-body disk temperature of $106 \pm 21°$K.

523.164.3 136
Observed Variations in the Amplitude Scintillations of the Cassiopeia (23N5A) Radio Source—H. J. A. Chivers. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 54–64; September. 1960.) Observations over a period of four years are analyzed. Daily variations are presented as functions of solar and sidereal time, and the effect of the solar cycle is studied.

523.164.3:621.391.812.6 137
Anomalous Winter Absorption of Radio Waves—R. W. Morriss. (*Proc. Phys. Soc.*, vol. 75, pp. 937–939; June 1, 1960.) Measurements made at Cambridge, England, of the absorption of cosmic radio noise on a frequency of 24.3 Mc show that the average midday absorption in winter is about twice as great as in summer. An attempt is made to relate this abnormality to the winter anomaly in absorption of reflected waves. If certain assumptions are made, it can be shown that on the majority of days in winter the absorbing electrons are largely concentrated at a height of about 60 km.

523.165 138
Ionized Gas and High-Energy Electrons in the Vicinity of the Earth and in the Interplanetary Space—K. I. Gringauz, V. G. Kurt, V. I. Moroz and I. S. Shklovskiĭ. (*Dokl. Akad. Nauk SSSR*, vol. 132, pp. 1062–1068; June 11, 1960.) Examination of results obtained by the three Soviet cosmic rockets relating to the Van Allen belts is given. The variation of ion concentration with distance from the earth and the dependence of the trap collector current and the potential of the container on the electron current at given ion concentrations are shown graphically.

523.165 139
Calculations of Cosmic-Ray Trajectories near the Equator—P. J. Kellogg. (*J. Geophys. Res.*, vol. 65, pp. 2701–2703; September, 1960.)

523.165 140
Electrons of the Van Allen Radiation—P. J. Kellogg. (*J. Geophys. Res.*, vol. 65, pp. 2705–2713; September, 1960.) Flux calculations show that electrons in the radiation belts cannot be produced by neutron decay from the cosmic-ray albedo. The acceleration of trapped electrons by electromotive fields is briefly discussed. See also 2217 of 1959.

523.165 141
Rapid Reduction of Cosmic-Radiation Intensity Measured in Interplanetary Space—C. Y. Fan. P. Meyer and J. A. Simpson. (*Phys. Rev. Lett.*, vol. 5, pp. 269–271; September 15, 1960.) Simultaneous observations in Pioneer V and on the ground demonstrate that the Forbush decrease is of solar origin, takes place in interplanetary space and is not related to the presence of the earth or its magnetic field.

523.165:523.745 142
Solar Cosmic Rays and Soft Radiation Observed at 5.000,000 Kilometres from Earth—R. L. Arnoldy, R. A. Hoffman and J. R. Winckler. (*J. Geophys. Res.*, vol. 65, pp. 3004–3007; September, 1960.) A summary is given of the cosmic-ray and soft-radiation effects observed during March/April, 1960, using a space probe and ground observations.

523.165:523.75 143
Observations of Solar Cosmic Rays near the North Magnetic Pole—K. A. Anderson and D. C. Enemark. (*J. Geophys. Res.*, vol. 65, pp. 2657–2671; September, 1960.)

523.165:523.75 144
Comparison of Solar Cosmic Rays Injection including July 17, 1959, and May 4, 1960—K. G. McCracken and R. A. R. Palmeira. (*J. Geophys. Res.*, vol. 65, pp. 2673–2683; September, 1960.)

523.165:523.75 145
Observations of Solar Flare Radiation and Modulation Effects at Balloon Altitudes, July 1959—A. Ehmert, H. Erbe, G. Pfotzer, C. D. Anger and R. R. Brown. (*J. Geophys. Res.* vol. 65, pp. 2685–2694; September, 1960.)

523.165:523.75 146
Propagation of Low Energy Cosmic Ray Particles associated with Solar Flares—K. Sakurai. (*J. Geomag. Geoelec.*, vol. 11, pp. 152–164; 1960.) The propagation mechanism is examined using data obtained during the period June, 1957–May, 1959.

523.165:538.12 147
Cosmic-Ray Intensity Variations and the Interplanetary Magnetic Field—H. Elliot. (*Phil. Mag.*, vol. 5, pp. 601–619; June, 1960.) The characteristics of the variations can be explained in terms of a dipole type of field generated by current systems in the solar corona.

523.165:550.385.4 148
Low-Energy Solar Cosmic Rays and the Geomagnetic Storm of May 12, 1959—J. R. Winckler and P. D. Bhavsar. (*J. Geophys. Res.*, vol. 65, pp. 2637–2655; September, 1960.)

523.165:550.385.4 149
The Effect of the Initial Phase of a Magnetic Storm upon the Outer Van Allen Belt—J. M. Malville. (*J. Geophys. Res.*, vol. 65, pp. 3008–3010; September, 1960.) A theoretical study is made of the motion of a charged particle trapped in a magnetic dipole field and in a compressed dipole field. The latitude shift of the trapped radiation during magnetic storms is found to be small.

523.165:551.507.362.1 150
Experiments on the Eleven-Year Change of Cosmic-Ray Intensity using a Space Probe—C. Y. Fan, P. Meyer and J. A. Simpson. (*Phys. Rev. Lett.*, vol. 5, pp. 272–274; September 15, 1960.) It is shown that the mechanism responsible for the changes is heliocentric, and the region in which the changes occur is located principally outside the earth's orbit.

523.165:551.507.362.1 151
Investigation of Interplanetary Ionized Gas, High-Energy Electrons and Solar Corpuscular Radiation by means of Three-Electrode Traps for Charge-Carrying Particles Installed on the Second Soviet Space Rocket—K. I. Gringauz, V. V. Bezrukikh, V. D. Ozerov and R. E. Rybchinskiĭ. (*Dokl. Akad. Nauk SSSR*, vol. 131, pp. 1301–1304; April 21, 1960.) Between 5500 and 75,000 km from the earth, electron streams of 10^8 electrons/cm² sec and energies >200 ev were recorded. Protons with energies of 25 ev at heights of 125,000 km were also noted.

523.165:551.507.362.2 152
Development of Multiple Radiation Zones on October 18, 1959—B. J. O'Brien and G. H. Ludwig. (*J. Geophys. Res.*, vol. 65, pp. 2695–2699; September, 1960.) Observations made at about 1000-km altitude during a magnetic storm using counters carried on Explorer VII are presented.

523.165:551.507.362.2 153
Outer Radiation Belt and Solar Proton Observations with Explorer VII during March–April 1960—J. A. Van Allen and W. C. Lin. (*J. Geophys. Res.*, vol. 65, pp. 2998–3003; September, 1960.) A preliminary summary is given of observed counting rates which includes rate/time plots for typical passes. Some of the fluctuations in the outer belt during the observation period are referred to briefly.

523.165:551.507.362.2 154
Detection of 10-keV Electrons in the Upper Atmosphere by the Third Earth Satellite—V. I. Krasovskiĭ, I. S. Shklovskiĭ, Yu. I. Gal'perin and E. M. Svetlitskiĭ. (*Izv. Akad. Nauk SSSR, Ser. geofiz.*, pp. 1157–1163; August, 1959.) Electrons of energies 10–40 kev were detected by means of two fluorescent screens covered with thin Al foils of different thickness used in conjunction with photomultipliers. The flux increases sharply with decreasing energy. At high latitudes, the night-time energy flux reaches a few tens of ergs/cm² sec sterad.

523.165:551.510.535 155
Penetration of Fast Charged Particles from the Outer Atmosphere into the Ionosphere—V. D. Pletnev. (*Izv. Akad. Nauk SSSR, Ser. geofiz.*, pp. 1164–1166; August, 1959.) In streams of oscillating particles trapped in the earth's magnetic field, the directions of motion will show pronounced spatial anisotropy due to absorption effects in the lower layers of the atmosphere.

523.165:551.594.5 156
The Outer Radiation Belt and Aurorae—H. Liemohn. (*Nature*, vol. 188, pp. 394–395; October 29, 1960.) Space probe observations are found to be in good agreement with a proposed unsymmetric model for the outer radiation belt. See also 4063 of 1959 (Rees and Reid).

523.165:551.594.5 157
Correlation of an Auroral Arc and a Subvisible Monochromatic 6300-Å Arc with Outer-Zone Radiation on November 28, 1959—B. J. O'Brien, J. A. Van Allen, F. E. Roach and C. W. Gartlein. (*J. Geophys. Res.*, vol. 65, pp. 2759–2766; September, 1960.)

523.53:621.396.96 158
A Southern Hemisphere Survey of the Radiants of Sporadic Meteors—A. A. Weiss and J. W. Smith. (*Mon. Not. Roy. Astron. Soc.*, vol. 121, no. 1, pp. 5–16; 1960.) Results are given of a new survey with higher resolution equipment [see 4208 of 1960 (Weiss)] than that used for the earlier survey [109 of 1958 (Weiss)]. Good agreement is found with the results obtained by Hawkins in the northern hemisphere (2119 of 1957).

525.35:523.745:529.786 159
Effect of the Solar Activity on the Earth's Rotation—H. Uyeda, Y. Sakuri and H. Iwasaki. (*J. Radio Res. Labs.. Japan*, vol. 7, pp. 131–136; May, 1960.) A linear relation is shown to exist between sunspot activity and d/dt (CsT—U.T.2) where CsT is atomic time. This result is discussed in relation to the influence of solar activity on variations in the speed of rotation of the earth on its axis.

550.37 160
The Structure of the Electrostatic Field in the Free Atmosphere from the Evidence Obtained during International Geophysical Year Investigations—I. M. Imyanitov and E. V. Chubarina. (*Dokl. Akad. Nauk SSSR*, vol. 132, pp. 104–107; May 1, 1960.) An examination of the spherical-capacitor hypothesis with reference to the variation of the electric potential with height and time of day is given.

550.37 161
Measurement of the Atmospheric Electric Field in the Free Atmosphere—R. Mühleisen and H. J. Fischer. (*Naturwiss.*, vol. 47, pp. 36–37; January, 1960.) The results of radiosonde measurements up to about 30-km height are compared with measured and calculated field-strength data of other authors.

550.385.4 162
Geomagnetic-Storm Sudden-Commencement Rise Times—A. J. Dessler, W. E. Francis and E. N. Parker. (*J. Geophys. Res.*, vol. 65, pp. 2715–2719; September, 1960.) If the "sudden commencement" of a magnetic storm is interpreted as the impact of a solar plasma "front" on the geomagnetic field, its rise time, usually a few minutes, depends on the different transit times of hydromagnetic waves propagated to the earth's surface from different points on the boundary of the field.

550.385.4 163
A Note on Harmonic Analysis of Geophysical Data with Special Reference to the Analysis of Geomagnetic Storms—M. Sugiura. (*J. Geophys. Res.*, vol. 65, pp. 2721–2725; September, 1960.) The effect of the earth's rotation on the determination of magnetic disturbance variations DS and Dst is discussed.

551.507.362.1 164
Radio-Astronomy Observations of the Second Soviet Space Rocket—V. V. Vitkevich, A. D. Kuz'min, R. L. Sorochenko and V. A. Udal'tsov. (*Dokl. Akad. Nauk SSSR*, vol. 132, pp. 85–88; May 1, 1960.) A report is made of observations of radio signals from the rocket which reached the moon at 0h 02 min 21±1 sec on September 14, 1959. Signal intensity measurements were made using an interferometer technique.

551.507.362.2 165
On Mr. King-Hele's Theory of the Effect of the Earth's Oblateness on the Orbit of a Close Satellite—P. J. Message. (*Mon. Not. Roy. Astron. Soc.*, vol. 121, no. 1, pp. 1–4; 1960.) An examination of the relations between parameters used by King-Hele (2233 of 1959) and the more usual osculating orbital elements is given.

551.507.362.2 166
On the Motion of a Satellite in an Asymmetrical Gravitational Field—P. Musen. (*J. Geophys. Res.*, vol. 65, pp. 2783–2792; September, 1960.) A theory for the inclusion of the effects of tesseral harmonics in the motion is given.

551.507.362.2 167
Angular Motion of the Spin Axis of the Tiros I Meterological Satellite due to Magnetic and Gravitational Torques—W. R. Bandeen and W. P. Manger. (*J. Geophy. Res.*, vol. 65, pp. 2992–2995; September, 1960.) The equations of motion including the two torques have been solved and the calculated and observed directions of the spin axis are in close agreement.

551.507.362.2:531.3 168
Perturbations in Classical Mechanics—L. M. Garrido. (*Proc. Phys. Soc., London*, vol. 76, pp. 33–35; July 1, 1960.) A method similar to quantum-mechanics time-dependent perturbation theory is described, which might be useful in the calculation of satellite orbits and plasma stability.

551.507.362.2:551.576 169
U. S. Meteorological Satellite Cameras photograph Cloud Cover—(*Science*, vol. 131, pp. 1031–1033; April 8, 1960.) A description is given of the Tiros I (1960 $\beta2$) satellite and associated tracking and interrogation equipment (see also 629 of 1960), together with a report on its launching and initial orbits.

551.507.362.2:621.396.96 170
Weather Radar Observations from an Earth Satellite—J. E. Keigler and L. Krawitz. (*J. Geophys. Res.*, vol. 65, pp. 2793–2808; September, 1960.) A survey of the problems involved in designing a weather satellite is given. The use of radar in the satellite is discussed and also the problem of handling the data on the ground and in the satellite.

551.510.53:621.384.8 171
A Method for Measuring Temperature Directly in the Upper Atmosphere with a Rocket-Borne Magnetic Mass Spectrometer—C. Y. Johnson, J. H. Hoffman, J. M. Young and J. C. Holmes. (*J. Geophys. Res.*, vol. 65, pp. 2996–2997; September, 1960.) Gas temperature is determined from the variation of the flux of neutral particles entering the instrument.

551.510.535 172
On the Problem of an Effective Recombination Coefficient in the Ionosphere—B. A. Bagaryatskiĭ. (*Izv. Akad. Nauk SSSR, Ser. geofiz.*, no. 9, pp. 1359–1363; September, 1959.) A discussion is given of possible chemical exchange reactions operative in establishing ion balance in the ionosphere.

551.510.535 173
Changes f_{min} at Kokubunji, Tokyo—H. Hojo. (*J. Radio Res. Labs., Japan*, vol. 7, pp. 213–243; May, 1960.) A statistical analysis is made of f_{min} data recorded during the

period 1946–1958, and a study is given of their relation to ionospheric absorption, atmospheric noise and other geophysical effects.

551.510.535 174

A Statistical Theory of Ionospheric Drifts—J. P. Dougherty. (*Phil. Mag.*, vol. 5, pp. 553–570; June, 1960.) A statistical study of the fluctuations in plasma density is used in an attempt to interpret drift velocities. It is assumed that the ionization is transported by a turbulent flow in the neutral air under the action of the magnetic field and the dynamo electric field. The application of the theory to different heights in the ionosphere is examined.

551.510.535 175

Observations of Sudden Ionospheric Disturbances—R. W. Morriss. (*Proc. Phys. Soc.*, vol. 76, pp. 79–92; July 1, 1960.) Observed phase changes for reflected waves of frequency 2–6 Mc are possible due to an increase in electron density, extending to above the E-layer maximum. Measurements of the absorption of cosmic noise transmitted through the ionosphere show that an increase occurs at heights as low as 65 km or even 50 km.

551.510.535 176

Use of the Incoherent Scatter Technique to Obtain Ionospheric Temperatures—T. E. VanZandt and K. L. Bowles. (*J. Geophys. Res.*, vol. 65, pp. 2627–2628; September, 1960.) If the ionization above the F_2 peak is in diffusive equilibrium, its density decreases exponentially with height, with a scale height proportional to the temperature of the neutral gas. Electron-density profiles obtained by the incoherent scatter method at 41 Mc are consistent with a temperature of about 1000°K between 370 km and 520 km.

551.510.535 177

F-Region Travelling Disturbances and Sporadic-E Ionization—L. H. Heisler and J. D. Whitehead. (*J. Geophys. Res.*, vol. 65, pp. 2767–2773; September, 1960.) A statistical study shows an association between F-region traveling disturbances and increases in E_s ionization at Sydney, Australia. For about one disturbance in 15, f_0E_s increases by approximately 1 Mc in some 13 minutes after the appearance of the disturbance and returns to normal 12 minutes later.

551.510.535 178

Damping Coefficient of Vibrating Electrons—V. Marasigan. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 65–67; September, 1960.) "A Druyvesteyn distribution is substituted for the assumed Maxwellian distribution in the derivation of the damping coefficient of vibrating electrons. The consequences are discussed."

551.510.535 179

Formation of the Sporadic E Layer in the Temperate Zones—J. D. Whitehead. (*Nature*, vol. 188, p. 567; November 12, 1960.) An extension is made of work described earlier [3501 of 1960 (Heisler and Whitehead)] to include regions where the gyro-frequency of the ions is less than their collision frequency.

551.510.535:523.164 180

Further Observations of Radio Stellar Scintillation—I. L. Jones. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 26–36; September, 1960.) Results are described which support those of Spencer (121 of 1956) concerning the elongation of ionospheric irregularities. The size of the irregularities is estimated, and a change in dimensions with elevation of the radio source is observed.

551.510.535:551.507.362.1 181

Electron-Density Distribution in the Upper Ionosphere from Rocket Measurements—J. S. Nisbet. (*J. Geophys. Res.*, vol. 65, pp. 2597–2599; September, 1960.) Seven flights between 1956 and 1959 having maximum heights of 400–800 km are discussed. The electron distribution above the F_2-layer maximum agrees well with that given by Chapman theory.

551.510.535:551.507.362.2 182

A Variable Atmospheric-Density Model from Satellite Accelerations—L. G. Jacchia. (*J. Geophys. Res.*, vol. 65, pp. 2775–2782; September, 1960.) An empirical expression is derived which relates atmospheric density ρ and scale height H to satellite orbital elements and 20-cm solar RF noise flux F. From this, ρ can be obtained as a function of F, height h and solar zenith angle ψ. Tables of H, ρ and $\rho H^{1/2}$ are given for $h=200$–700 km and for two values of F and ψ.

551.510.535:551.507.362.2 183

The Determination of Ionospheric Electron Content from Satellite Doppler Measurements: Part 1—Method of Analysis—W. J. Ross. (*J. Geophys. Res.*, vol. 65, pp. 2601–2606; September, 1960.) Theory is given which is based on measurements of the slopes of the differential Doppler and the Doppler curves obtained from observations of the transmission at 20 and 40 Mc from Sputnik III. A semi-empirical formula is derived for the total electron content.

551.510.535:551.507.362.2 184

The Determination of Ionospheric Electron Content from Satellite Doppler Measurements: Part 2—Experimental Results—W. J. Ross. (*J. Geophys. Res.*, vol. 65, pp. 2607–2615; September, 1960.) The results of observations made from September, 1958, to December, 1959, are presented. A winter to summer variation of 2.5:1 in total midday electron content was observed. The diurnal variation showed an increasing content towards evening, although the content below h_{max} was decreasing. Day-to-day variations correlated well with $1/K_p$ during the summer.

551.510.535:551.507.362.2 185

The Electron Content and Distribution in the Ionosphere—T. G. Hame and W. D. Stuart. (PROC. IRE, vol. 48, pp. 1786–1787; October, 1960.) Addendum to 2009 of 1960.

551.510.535:621.391.812.63 186

Use of Logarithmic Frequency Spacing in Ionogram Analysis—G. A. M. King. (*J. Res. NBS*, vol. 64D, pp. 501–504; September/October, 1960.) An improved method for the reduction of ionograms to electron-density profiles is given.

551.510.535:621.391.812.63:621.3.087.4 187

An Improved Automatic Method for Measurement of Ionospheric Absorption—L. Thomas. (*Electronic Engrg.*, vol. 32, pp. 635–649; October, 1960.) Pulses are transmitted at vertical incidence at a rate of one per second, and the mean echo amplitude over an interval of about one minute is recorded.

551.510.535:621.391.812.63.029.63 188

Some Characteristics of Ionospheric Backscatter Observed at 440 Mc/s—V. C. Pineo, L. G. Kraft and H. W. Briscoe. (*J. Geophys. Res.*, vol. 65, pp. 2629–2633; September, 1960.) An investigation of the E and F layers by the incoherent back-scatter technique, using either vertical or oblique incidence is described. The spectra of radiation scattered from different heights, and an electron-density profile from 200 km to 700 km, are shown.

551.510.535:621.396.96:523.3 189

Faraday Rotation Observations of the Electron Content of the Exosphere—R. B. Dyce. (*J. Geophys. Res.*, vol. 65, pp. 2617–2618; September, 1960.) The moon-echo observations of various workers are summarized. The ratios obtained for the ionospheric electron content above h_{max} to that below are tabulated.

551.510.535:621.396.96:523.3 190

Radar-Lunar Investigations at a Low Geomagnetic Latitude—G. H. Millman, A. E. Sanders and R. A. Mather. (*J. Geophys. Res.*, vol. 65, pp. 2619–2626; September, 1960.) The theory and results are presented for polarization rotation observations of pulsed radar transmissions at 400 Mc reflected from the moon. The diurnal variation of total electron content of the ionosphere is shown.

551.510.536:061.3 191

Symposium on the Exosphere and Upper F Region—(*J. Geophys. Res.*, vol. 65, pp. 2563–2636; September, 1960.) The text is given of papers presented at the symposium held in Washington, D. C., on May 4, 1960. Some of the papers are abstracted separately, others are listed below.

a) The Exosphere and Upper F Region—F. S. Johnson (pp. 2571–2575).

b) Structure of the Earth's Exosphere—S. F. Singer (pp. 2577–2580).

c) Whistler Dispersion and Exospheric Hydrogen Ions—R. E. Barrington and T. Nishizaki (pp. 2581–2582).

d) Electron Densities to 5 Earth Radii deduced from Nose Whistlers.—R. L. Smith and R. A. Helliwell (p. 2583).

e) Comment on Models of the Ionosphere above $h_{max}F_2$—J. W. Wright (pp. 2595–2596).

f) Radio-Wave Scattering by an Ionized Gas in Thermal Equilibrium—J. A. Fejer (pp. 2635–2636).

551.510.536:551.507.362.1 192

A Sounding Rocket Measurement of Electron Densities to 1500 Kilometres—W. W. Berning. (*J. Geophys. Res.*, vol. 65, pp. 2589–2594; September, 1960.) Doppler measurements were made at 37 and 148 Mc. Above the F_2-layer peak, the electron-density decrease indicates a scale height of 30 km at 270 km, and a linear gradient in scale height of 0.45 from 270 to 760 km.

551.510.536:551.507.362.2 193

Radio Propagation Measurements using the Explorer VI Satellite—C. D. Graves. (*J. Geophys. Res.*, vol. 65, pp. 2585–2587; September, 1960.) Coherent signals at 108 and 378 Mc were emitted by the satellite. The results gave electron densities of about 10^4/cm³ at 18,000 km above the earth. This is higher than expected, and may have been influenced by an intense geomagnetic storm, or by the need to make large corrections.

551.510.536:[551.594.6+551.507.362.2 194

The Electron Density Distribution derived from Whistler Data and Faraday-Fading Observations—K. H. Schmelovsky. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 68–71; September, 1960.) An interpolation law allows satellite observations to be combined with whistler data.

551.594.221 195

E.L.F. Electric Fields from Thunderstorms—A. D. Watt. (*J. Res. NBS*, vol. 64D, pp. 425–433; September/October, 1960.) From an assumed cloud-to-ground discharge model calculations of wide-band vertical electric field variations indicate a small decrease with distance d up to 4 km, a d^{-3} relation to 20 km, and a d^{-1} relation beyond 30 km. Radiation-field spectrum analysis over the frequency range 1 cps–100 kc indicates that, below 300 cps, long discharges produce more energy than short discharges, and that intercloud discharges may produce as much as cloud-to-ground discharges. Conclusions are supported by experimental evidence.

551.594.221 196
The Thunderstorms of 5 September 1958: Lightning Discharges and Atmospherics—F. Horner. (*Met. Mag., London*, vol. 89, pp. 257–261; October, 1960.) A report is given on visual observations made at Slough, and an analysis is made of records taken by lightning-flash counters at Slough and other stations in the vicinity of the storm.

551.594.5 197
Direct Measurement of Particles producing Visible Auroras—C. E. McIlwain. (*J. Geophys. Res.*, vol. 65, pp. 2727–2747; September, 1960.) A large amount of auroral light is produced by electrons with energies <10 kev.

551.594.5:550.385.4 198
Large-Scale Auroral Motions and Polar Magnetic Disturbances: Part 1—A Polar Disturbance at about 1100 Hours on 23 September 1957—S. I. Akasofu. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 10–25; September, 1960.) A detailed study of a particular event suggests that both the westward auroral electrojet and the eastward auroral motion are produced simultaneously by a southward electric field, the origin of which is discussed. Over 50 references.

551.594.5:621.396.96 199
The Height and Geometry of Auroral Radio Echoes—C. D. Watkins. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 1–9; September, 1960.) At Jodrell Bank, echoes at 4 meters λ have been obtained only from regions where the line of sight is within 1° of perpendicularity to the local magnetic field lines at a height of about 110 km. See also 1429 of 1958 (Bullough, *et al.*).

551.594.6 200
Analysis of Smooth-Type Atmospheric Waveforms—F. Hepburn. (*J. Atmos. Terr. Phys.*, vol. 19, pp. 37–53; September, 1960.) Over 250 traces are analyzed in an attempt to resolve previous conflicting results [*e.g.*, 2888 of 1948 (Hales) and 2809 of 1952 (Caton and Pierce)]. The precise nature of observations necessary for the accurate analysis and reliable detection of deviations from simple theory is indicated.

551.594.6 201
Guiding of Whistlers in a Homogeneous Medium—R. L. Smith. (*J. Res. NBS*, vol. 64D, pp. 505–508; September/October, 1960.) The velocity of energy flow of whistlers as a function of wave-normal angles is computed. Results differ from the longitudinal value except at very-low frequencies or very-small angles.

551.594.6:539.16 202
Electromagnetic Signals from Nuclear Explosions in Outer Space—M. H. Johnson and B. A. Lippmann. (*Phys. Rev.*, vol. 119, pp. 827–828; August 1, 1960.) X-rays produced by a nuclear burst in outer space cause polarization currents in the medium, which, if distributed anisotropically, will emit electromotive radiation. A rough calculation indicates that a detectable 10-Mc signal is produced at a range of 1 km from a burst of X-ray energy equivalent to 1 ton of high explosive. The signal power varies as the square of the electron density, so this effect may provide a sensitive measure of the electron density in outer space

551.594.6:539.16 203
Bomb-Excited "Whistlers"—B. A. Lippmann. (Proc. IRE, vol. 48, pp. 1778–1779; October, 1960.) Reasons are given for the possible excitation of whistler modes by nuclear explosions, and characteristics of the signals to be expected are discussed.

551.594.6:551.594.5 204
Audio-Frequency Electromagnetic Radiation in the Auroral Zone—G. Gustafsson, A. Egeland and J. Aarons. (*J. Geophys. Res.*, vol. 65, pp. 2749–2758; September, 1960.) Spectograms of the energy in the frequency range 10 cps–10 kc were recorded in N. Sweden during the winter and early summer of 1958–1959. The energy is low during the day and high at night, and the ratio is greatest in the range 20–200 cps. No clear-cut correlation with magnetic indexes could be found, except during some geomagnetic disturbances. The main contribution to the energy probably originates in atmospheric noise from thunderstorms, but magnetic fluctuations may be important at lower frequencies.

LOCATION AND AIDS TO NAVIGATION

534.2–14:534.88 205
A Systematic Error in Underwater Acoustic Direction-Finding—J. J. Freeman. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1025–1027; August, 1960.) The effect of the coherence between noise measured at different locations is examined, and an evaluation is made of the systematic error introduced by this coherence in cross-correlator DF systems.

621.396.663:621.396.677.6 206
Theory of the General Adcock—Baur. (See 42.)

621.396.931.2 207
Direction Finding at Low Frequencies—L. E. Orsak and D. W. Martin. (*Electronics*, vol. 33, pp. 74–77; September 16, 1960.) Details are given of a small rotating-loop direction finder for frequencies between 15 and 500 kc. Automatic operation is provided to give aural null indication or a CRO display of the antenna pattern.

621.396.96 208
Polarization and Depression-Angle Dependence of Radar Terrain Return—I. Katz and L. M. Spetner. (*J. Res. NBS*, vol. 64D, pp. 483–486; September/October, 1960.) Recent experimental data indicate that the polarization dependence of sea return cannot be explained by interference pattern analysis.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 209
Photoconductivity of Complex Inorganic Compounds—H. Schindler. (*Z. angew. Phys.*, vol. 12, pp. 33–38; January, 1960.) Measurements made on $K_3Fe(CN)_6$ and $K_3Mn(CN)_5NO$ are discussed.

535.215:545.23 210
Some Electrical Properties of Amorphous Selenium Films—R. A. Fotland. (*J. Appl. Phys.*, vol. 31, pp. 1558–1565; September, 1960.) Results of measurements of dark-current and photocurrent characteristics, including hysteresis and transient effects are given, mainly in graphical form. 27 references.

535.215:547.672 211
Charge-Carrier Production and Mobility in Anthracene Crystals—R. G. Kepler. (*Phys. Rev.*, vol. 119, pp. 1126–1229; August 15, 1960.) The drift mobilities of electrons and holes in anthracene crystals have been measured using a pulsed photoconductivity technique. The experimental results indicate that the charge carriers are produced at the surface of the crystal and not in the interior.

535.376 212
Measurement of Luminescence Decay Times of Inorganic Phosphors with Excitation by Ions—A. Scharmann. (*Z. Phys.*, vol. 157, pp. 301–315; November 16, 1959.) Measurements of decay times down to about 10^{-7} s on $MgWO_4$, $ZnWO_4$, $CdWO_4$, $CaWO_4$, CsI-Tl and NaCl-Ag show the decay to be exponential and independent of wavelength. The luminescence of ZnS-Cu decreases hyperbolically, and the blue emission band of ZnS-Mn shows similar behavior.

535.376 213
The Luminescence Decay of ZnS-Cu, ZnS-S, ZnS-Zn, and ZnO-Zn with Excitation by 60-keV H_2† Ion Pulses—K. H. Härdtl. (*Z. Phys.*, vol. 157, pp. 316–325; November 16, 1959.) Increasing decay times were observed with the lengthening of ion-pulse duration. See also 212 above.

537.226:536.421 214
Influence of Crystallographic Orientation on the Charge Formation during Phase Changes in Solids—S. Mascarenhas and L. G. Freitas. (*J. Appl. Phys.*, vol. 31, pp. 1684–1685; September, 1960.) An experimental investigation is made of the thermodielectric effect, which may be important in zone-refining techniques.

537.226:621.319.2:535.37 215
Some Effects observed in Studying the Luminescence of ZnS Electrets—V. M. Fridkin. (*Dokl. Akad. Nauk SSSR*, vol. 131, pp. 290–292; March 11, 1960.) The intensity of luminescence was greater when the applied field was opposite to the direction of the field during polarization. An effect similar to thermoluminescence was produced by heating.

537.227.228 216
Dielectric Properties of Lead Titanate Zirconate Ceramics at Very Low Frequencies—R. Gerson. (*J. Appl. Phys.*, vol. 31, pp. 1615–1617; September, 1960.)

537.227 217
Crystal Structure of Ferroelectric $LiH_3(SeO_3)_2$—K. Vedam, Y. Okaya and R. Pepinsky. (*Phys. Rev.*, vol. 119, pp. 1252–1255; August 15, 1960.)

537.227:546.431'824–31 218
Temperature Dependence of the Velocity of Sidewise 180° Domain-Wall Motion in $BaTiO_3$—A. Savage and R. C. Miller. (*J. Appl. Phys.*, vol. 31, pp. 1546–1549; September, 1960.) From 25° to 100° C, the dependence is primarily due to the activation field and not in the extrapolated velocity for $E=\infty$. The effect of impurities and crystal thickness is considered.

537.227:546.431'824–31 219
The Electrical Conductivity of Barium Titanate Single Crystals—A. Branwood and R. H. Tredgold. (*Proc. Phys. Soc., London*, vol. 76, pp. 93–98; July 1, 1960.) Conductivity was measured as a function of temperature, applied field, electrode material, thickness and time. Possible explanations of the striking nonohmic nature of the conductivity are briefly discussed.

537.227:546.431'824–31 220
The Influence of the Addition of Fe_2O_3, V_2O_5 and SnO_2 on Dielectric Properties of Barium Titanate Ceramics—G. Lapluye, G. Morinet and P. Palla. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 79–81; January 4, 1960.)

537.227:546.431'824–31 221
The Effect of the Incorporation of Metallic Oxides on the Properties of a Barium Titanate

—G. Lapluye, G. Morinet and P. Palla. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 305–307; January 11, 1960.)

537.228.1 222

Measurement of Coupling Coefficient and Q of Low-Q Piezoelectric Ceramics—R. M. Glaister. (*Brit. J. Appl. Phys.*, vol. 11, pp. 390–391; August, 1960.) Two methods are described for deriving Q and the coupling coefficient for ceramics which do not exhibit antiresonance. An example is given.

537.311.33 223

The Role of Electron-Phonon Interaction in the Impurity Conduction of Semiconductors—Y. Toyozawa. (*Prog. Theoret. Phys., Kyoto*, vol. 23, pp. 378–380; February, 1960.) Factors contributing to impurity conduction in the low-concentration range are examined qualitatively.

537.311.33 224

Statistics of the Occupation of Dislocation Acceptors (One-Dimensional Interaction Statistics)—R. M. Broudy and J. W. McClure. (*J. Appl. Phys.*, vol. 31, pp. 1511–1516; September, 1960.) Occupation statistics are derived, taking account of interactions between nearest-neighbor electrons along a dislocation line in a semiconductor. Numerical results have been computed and are tabulated.

537.311.33 225

Bounds on the Nonlinear Diffusion-Controlled Growth Rate of Spherical Precipitates—J. A. Morrison and H. L. Frisch. (*J. Appl. Phys.*, vol. 31, pp. 1621–1627; September, 1960.)

537.311.33 226

Study of Charge Carriers in Semiconductors subjected to Strong Electric Fields—J. Bok. (*Ann. Radioélect.*, vol. 15, pp. 120–146; April, 1960. English abridgment, pp. 201–203.) A detailed experimental investigation is made of effects associated with "hot" electrons. See also 3781 of 1959 (Bok and Veilex).

537.311.33 227

The Mechanism of Diffusion-Controlled Field Instabilities in Semiconductors at High Field Strengths—H. Rother. (*Ann. Phys., Lpz.*, vol. 5, pp. 203–210; January 15, 1960.) The conditions are derived for the growth of strong space-charge fields in semiconductors. A theoretical basis is established for the space-charge effects discussed by Böer and Rompe (249 below).

537.311.33:534.2 228

Interaction of Conduction Electrons with Acoustic Waves in Simple Semiconductors—N. Mikoshiba. (*J. Phys. Soc. Japan*, vol. 15, pp. 982–989; June, 1960.) A semiclassical theory is given. The extent to which various methods of approach are applicable is discussed, and results are compared with those for monovalent metals.

537.311.33:538.614 229

The Faraday Effect in Semiconductors due to Free Carriers in a Strong Magnetic Field—L. É. Gurevich and I. P. Ipatova. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 1324–1329; November, 1959.) The refractive indexes calculated for various directions of the magnetic field relative to the main crystallographic axes depend strongly on the field direction. The components of the effective mass tensor can therefore be determined from measurement of the angle of rotation of the plane of polarization.

537.311.33:538.615 230

Theory of Line Shapes of Interband Magneto-optical Absorption in Semiconductors—T. Ohta, M. Nagae and T. Miyakawa. (*Prog. Theoret. Phys., Kyoto*, vol. 23, pp. 229–250; February, 1960.) A fuller report of work described earlier by Ohta and Miyakawa (2422 of 1960) is given.

537.311.33:[546.28+546.289+546.682'19 231

The Magnetic Susceptibility of Electrons in Silicon, Germanium and Indium Arsenide—D. Geist. (*Z. Phys.*, vol. 157, pp. 333–361; November, 1959.) Magnetic-balance methods of measuring susceptibility are discussed; the special balance used for measurements at 141° and 297°K analyzed in this paper will be described later. For preliminary results see 3510 of 1958 and 1690 of 1960.

537.311.33:[546.28+546.289 232

The Magnetic Susceptibility of Defect Electrons in Silicon and Germanium—D. Geist. (*Z. Phys.*, vol. 157, pp. 490–498; December 17, 1959.) See also 231 above.

537.311.33:[546.28+546.289]:539.12.04 233

Auger Electron Ejection from Germanium and Silicon by Noble Gas Ions—H. D. Hagstrum. (*Phys. Rev.*, vol. 119, pp. 940–952; August 1, 1960.)

537.311.33:546.28 234

Diffusion Rate of Li in Si at Low Temperatures—E. M. Pell. (*Phys. Rev.*, vol. 119, pp. 1222–1225; August 15, 1960.) The method of ion drift in the electric field of an n-p junction has been used to measure the diffusion constant of Li in Si between 25 and 125°C.

537.311.33:546.28 235

Effect of Li-B Ion Pairing on Li^+ Ion Drift in Si—E. M. Pell. (*J. Appl. Phys.*, vol. 31, pp. 1675–1679; September, 1960.)

537.311.33:546.28 236

Diffusion of Li in Si at High T and the Isotope Effect—E. M. Pell. (*Phys. Rev.*, vol. 119, pp. 1014–1021; August 1, 1960.) Outdiffusion measurements with Li^6 and Li^7 at 800°C and 1350°C are described, and the results are discussed.

537.311.33:546.28:538.63 237

Low-Field Magnetoresistance Effect of Plastically Deformed n-Type Silicon—K. Kamada. (*J. Phys. Soc. Japan*, vol. 15, pp. 998–1005; June, 1960.) Measurements show violation of the symmetry relation between the relaxation-time tensor and the effective-mass tensor. Theoretical consequences are discussed.

537.311.33:546.28:541.135 238

Capacitance Measurements at the Boundary Surface Silicon-Electrolyte—K. Böke. (*Z. Naturforsch.*, vol. 15a, pp. 550–551; May/June, 1960.) A discontinuity in the capacitance/reverse-voltage characteristic of thin Si specimens immersed in an electrolyte is used to prove the existence of a regular barrier layer formed in the specimen.

537.311.33:546.289 239

Exo-electron Emission of Germanium—R. Seidl. (*Z. Phys.*, vol. 157, pp. 568–575; January 8, 1960.) A report is made on experimental investigations in the temperature range 20°–260°C. Results indicate that traps located in the bulk of the material also give rise to the emission. See 3907 of 1960.

537.311.33:546.289 240

Optical Observation of Spin-Orbit Interaction in Germanium—J. Tauc and E. Antonkčík. (*Phys. Rev. Lett.*, vol. 5, pp. 253–254; September 15, 1960.) A double peak in the reflection spectrum is explained by spin-orbit interaction.

537.311.33:546.289:535.215:538.63 241

Spectral Distribution of the Photomagnetoelectric Effect in Ge: Experiment—F. A. Brand, A. N. Baker and H. Mette. (*Phys. Rev.*, vol. 119, pp. 922–925; August 1, 1960.) A study is made of the effect in Ge as a function of wavelength from 0.5–2 μ, with particular reference to the conditions under which a sign reversal is obtained. Results are in good qualitative agreement with theoretical work of Gärtner (2493 of 1957).

537.311.33:546.289:539.12.04 242

Drift Mobility in Neutron-Irradiated n-Type Germanium—W. H. Closser. (*J. Appl. Phys.*, vol. 31, p. 1693; September, 1960.) Measurements of the variation of drift mobility with integrated neutron flux are explained by the Gossick-Crawford model, together with scattering by point defects.

537.311.33:546.289:539.12.04 243

Impurity Conduction in Transmutation-Doped p-Type Germanium—H. Fritzsche and M. Cuevas. (*Phys. Rev.*, vol. 119, pp. 1238–1245; August 15, 1960.) The Hall coefficient and the resistivity of transmutation-doped Ge have been measured between 1.2 and 300°K. The impurity conduction was studied as a function of the separation between majority impurities, and the results are compared with existing theories.

537.311.33:546.289:548.5 244

Germanium Crystals Grown from Hollow Cylindrical Seeds—R. C. Frank and J. E. Thomas, Jr. (*J. Appl. Phys.*, vol. 31, pp. 1689–1690; September, 1960.) Observations of the growth of single-crystal and polycrystalline tubular Ge seeds are reported.

537.311.33:546.289'231 245

Electrical Properties of Germanium Selenide GeSe—S. Asanabe and A. Okazaki. (*J. Phys. Soc. Japan*, vol. 15, pp. 989–997; June, 1960.) Experimental studies were made of electrical resistivity and Hall effect from 100° to 800°K. Results are interpreted qualitatively.

537.311.33:546.47'221 246

The Interpretation of the ΔE Values of ZnS Phases—O. G. Folberth. (*Z. Naturforsch.*, vol. 15a, pp. 432–434; May/June, 1960.) The variation in the width of the forbidden zone in semiconducting compounds of ZnS structure is discussed. See also 2044 of 1960.

537.311.33:546.47'221 247

The Bonding in Crystals with "Normal Valency" with Particular Reference to ZnS and Wurtzite Phases—O. G. Folberth. (*Z. Naturforsch.*, vol. 15a, pp. 425–431; May/June, 1960.) A modification of the earlier view, that the bonding in crystals of ZnS structure is mainly covalent, is justified.

537.311.33:546.47–31 248

The Exo-electron Emission (Kramer Effect) of Zinc Oxide—R. Menold. (*Z. Phys.*, vol. 157, pp. 499–509; December 17, 1959.) Exo-electron-emission and conductivity glow curves were obtained for single-crystal and pure or doped powder material after X-ray bombardment. The maxima observed are attributed to centers in the sorption layer. See also 1686 of 1960.

537.311.33:546.48'221 249

Space-Charge Oscillations in Semiconductors at High Field Strengths—K. W. Böer and R. Rompe. (*Ann. Phys., Lpz.*, vol. 5, pp. 200–202; January 15, 1960.) Periodic and quasi-periodic phenomena observed in CdS single crystals [1637 of 1960 (Böer)] are discussed, and their similarity with phenomena in low-pressure gas discharges is noted.

537.311.33:546.48'221 250
Temperature-Stable Barrier-Layer-Free Contacts of CdS Single Crystals—K. W. Böer and K. Lubitz. (*Z. Naturforsch.*, vol. 15a, pp. 91–92; January, 1960.) Contacts are obtained by vacuum deposition of Al at 200°C, and can withstand heating up to 350°C in vacuum for 15 minutes without the formation of a barrier layer being observed.

537.311.33:546.561–31:539.23 251
Change of Conductivity of Thin Cuprous Oxide Films with Electrostatic Charge—A. Deubner and F. Schulz. (*Ann. Phys., Lpz.*, vol. 5, pp. 113–128; January 15, 1960.) Conductivity and field effect were measured to determine the effect of the conditions under which the film was prepared. Field-effect measurements in vacuum were made in the temperature range −56° to +30°C. The results obtained at room temperature agree with those of Zückler (1452 of 1954); the behavior observed at low temperatures cannot yet be interpreted.

537.311.33:546.681'19 252
Thermal Conversion in *n*-Type GaAs—J. J. Wysocki. (*J. Appl. Phys.*, vol. 31, p. 1686; September, 1960.) Material with a carrier concentration of $5\times10^{16}/cm^2$ or less changed to *p*-type when heated to 800°C for short times, due to Cu impurity in etching agents.

537.311.33:546.681'19 253
Growth of GaAs Crystals in the <111> Polar Direction—P. L. Moody, H. C. Gatos and M. C. Lavine. (*J. Appl. Phys.*, vol. 31, pp. 1696–1697; September, 1960.)

537.311.33:546.824–31:534.614–8 254
Ultrasonic Measurement of the Elastic Moduli of Rutile—G. L. Vick and L. E. Hollander. (*J. Acoust. Soc. Am.*, vol. 32, pp. 947–949; August, 1960.)

537.311.33:546.873'241:537.323 255
The Thermal Conductivity and Thermoelectric Power of Bismuth Telluride at Low Temperatures—P. A. Walker. (*Proc. Phys. Soc., London*, vol. 76, pp. 113–126; July 1, 1960.)

537.311.33:621.387.462 256
Measurement of the Coefficient of Charge Collection in *p-n*-Junction Particle Detectors. New Method of Measuring Lifetimes—L. Koch, J. Meisser and Q. Kerns. (*J. Electronics and Control*, vol. 8, pp. 289–300; April, 1960. In French.) An accurate value for the diffusion length of minority carriers may be obtained from the variation of the coefficient of charge collection as a function of nuclear-particle penetration.

537.312.62 257
Changes in Superconducting Critical Temperature Produced by Electrostatic Charging—R. E. Glover, III, and M. D. Sherrill. (*Phys. Rev. Lett.*, vol. 5, pp. 248–250; September 15, 1960.) Measurements of the superconducting transition temperature for In show a decrease with negative charging, and an equal increase with positive charging. Opposite changes are found for Sn.

538.221 258
Magnetic Interaction Domains—D. J. Craik and E. D. Isaac. (*Proc. Phys. Soc., London*, vol. 76, pp. 160–162, plate; July 1, 1960.) A note is given on domain patterns observed in a specially prepared material comprising elongated single-domain Fe particles embedded in a nonmagnetic matrix.

538.221 259
The Relaxation of Ferromagnetic Domain Walls—B. Rothenstein. (*Z. Naturforsch.*, vol. 15a, pp. 169–171; February, 1960.) Internal friction effects are discussed with reference to Bloch-wall mobility measurements on iron wire at frequencies between 20 and 200 cps [see also 3794 of 1959 (Rothenstein and Hrianca)].

538.221 260
Domain Tube Structures in Silicon Iron—L. F. Bates and R. Carey. (*Proc. Phys. Soc., London*, vol. 75, pp. 880–884; June 1, 1960.) Results of a quantitative study of Néel spike formation are discussed in terms of a formula given by Brenner (485 of 1956) and an attempt is made to assess the importance of the phenomena in the explanation of coercivity.

538.221 261
Some Observations on the Domain Walls in Thin Films—Y. Gondō and Z. Funatogawa. (*J. Phys. Soc. Japan*, vol. 15 pp. 1126–1127; June, 1960.) A report is given of observations of colloidal patterns on 20 per cent Fe-Ni films with uniaxial magnetic anisotropy.

538.221 262
Energy of Bloch Walls in Iron and Nickel—L. Špaček. (*Ann. Phys., Lpz.*, vol. 5, pp. 217–228; January 15, 1960.) Calculations are made for 90° and 180° walls in Fe and for 70°, 110° and 180° walls in Ni.

538.221 263
Thermal and Electrical Properties of Armco Iron at High Temperatures—M. J. Laubitz. (*Canad. J. Phys.*, vol. 38, pp. 887–907; July, 1960.)

538.221:538.13 264
Network-Analogue Method and Graphical Procedure for the Determination of Magnetic Fields with Locally Variable Permeability—P. A. Tschopp and A. H. Frei. (*Arch. Elektrotech.*, vol. 44, pp. 441–454; December 19, 1959.) Two methods for the determination of internal magnetic fields in ferromagnetic solids are described.

538.221:538.569.4 265
The Influence of Inhomogeneities on Ferromagnetic Resonance in Metals—W. Döring and H. Vial. (*Z. Naturforsch.*, vol. 15a, pp. 434–447; May/June, 1960.) The influence of magnetic inhomogeneities and of the stray magnetic field on the width of the permeability resonance curve is investigated for polycrystalline metals.

538.221:539.23 266
Investigation of the Magnetization of Thin Nickel Films—T. Rappeneau. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 674–676; January 25, 1960.) A report is given of measurements on Ni films between 300 and 700 Å thick in a magnetic field from 0 to 100 oersteds. Hysteresis cycles are shown, and results are compared with magnetoresistance data.

538.221:539.23 267
The Uniaxial Magnetic Anisotropy of Thin Films and its Change with Time—Z. Málek, W. Schüppel, O. Stemme and W. Andrä. (*Ann. Phys., Lpz.*, vol. 5, pp. 211–216; January 15, 1960.) Measurements were made on vapor-deposited Ni and permalloy films, and on electrolytically deposited Ni film.

538.221:539.23 268
Observations of the Magnetization Reversal Process in Thin Films of Nickel-Iron, using the Kerr Magneto-optic Effect—M. Prutton. (*Brit. J. Appl. Phys.*, vol. 11, pp. 335–338; August, 1960.) Experiments, using uniaxial films about 1500 Å thick, suggest that magnetization reversal may occur in three stages: a) coherent rotation; b) domain nucleation and growth; c) further coherent rotation.

538.221:539.23 269
Cross-Tie Walls in Thin Permalloy Films—M. Prutton. (*Phil. Mag.*, vol. 5, pp. 625–633; June, 1960.) A cross-tie wall is derived theoretically from a boundary-wall equation halfway between that of a Bloch and a Néel boundary. The energy density of the wall is estimated to be about 10 ergs/cm².

538.221:546.74–31 270
Reproducible Ferromagnetism under Conditions of Stoichiometric Impurity in Nickel Oxide—N. Perakis and G. Parravano. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 677–679; January 25, 1960.) For a range of excess oxygen, ferromagnetic remanence is observed during cooling in a magnetic field; this disappears on reheating to the Néel temperature.

538.221:621.318.124 271
Trigonal Magnetocrystalline Anisotropy in Hexagonal Oxides—L. R. Bickford, Jr. (*Phys. Rev.*, vol. 119, pp. 1000–1009; August 1, 1960.) Experimental results and a theoretical discussion are given of torque measurements on single crystals of Co_2Y and Co_2Z between 77°K and 300°K.

538.221:621.318.134 272
High-Permeability Garnets—E. E. Anderson and J. R. Cunningham, Jr. (*J. Appl. Phys.*, vol. 31, pp. 1687–1688; September, 1960.) Partial substitution of In for Fe in Y-Fe garnet results in a 500 per cent increase in initial permeability.

538.221:621.318.134 273
Some Observations of Bitter Patterns on Polycrystalline "Square Loop" Ferrites and a Theoretical Explanation of the Loop Shape and Pulse Characteristics of the Material—J. E. Knowles. (*Proc. Phys. Soc., London*, vol. 75, pp. 885–897; June 1, 1960.) It is postulated that the magnetization in each grain of a polycrystalline Mg-Mn ferrite lies along the [111] direction, and generally reverses by the motion of 180° domain walls.

538.221:621.318.134:556.569.4 274
The Wall Effect in Ferrimagnetic-Resonance Experiments—W. B. Nash and K. J. Standley. (*Proc. Phys. Soc., London*, vol. 76, pp. 99–103, plate; July 1, 1960.) The magnitude and sign of the wall shift was determined at wavelengths in the range of 0.87 to 3.12 cm and at temperatures between 20 and 300°C, using small spheres of Ni ferrite, Mg-Mn ferrite and Y-Fe garnet. A proposed simple model agrees with results in sign and order of magnitude.

538.221:621.318.134:538.569.4 275
Walker Modes in Large Ferrite Samples—B. A. Auld. (*J. Appl. Phys.*, vol. 31, pp. 1642–1647; September, 1960.) "The modes of resonance for a ferrite strip and a ferrite post resonator between conducting planes are found by taking appropriate combinations of uniform plane-wave solutions. It is shown that magnetostatic resonances (Walker modes) may exist even when the resonator dimensions are arbitrarily large provided the mode indices satisfy certain restrictions."

538.221:621.318.134:538.569.4 276
Magnetostatic Modes in Disks and Rods—J. F. Dillon, Jr. (*J. Appl. Phys.*, vol. 31, pp. 1605–1614; September, 1960.) Ferrimagnetic-resonance experiments on samples of single crystals of Mn ferrite and Y-Fe garnet show many magnetostatic modes as absorption maxima. Varying RF field configurations aid the identification of modes. The narrowest

lines were observed when the steady field was normal to the plane of a thin disk.

538.221:621.318.134:538.569.4 277

Investigation of the Variation of the Magnetic Moment of an Yttrium Garnet Single Crystal in a Ferromagnetic-Resonance Experiment at 9350 Mc/s—J. Hervé and M. Sausade. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 82–84; January 4, 1960.)

538.221:621.318.134:538.632 278

Hall Effect in Ferrites near the Curie Temperature—K. P. Belov and E. P. Svirina. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 1212–1216; November, 1959.) A report is given of measurements of Hall EMF in single-crystal and polycrystalline specimens of Ni-Zn and Mn ferrites.

538.222 279

Spin-Phonon Interaction in Paramagnetic Crystals—R. D. Mattick and M. W. P. Strandberg. (*Phys. Rev.*, vol. 119, pp. 1204–1217; August 15, 1960.)

538.222:538.569.4 280

Electron Spin-Lattice Relaxation in Dilute Potassium Chromicyanide at Helium Temperatures—J. G. Castle, Jr., P. F. Chester and P. E. Wagner. (*Phys. Rev.*, vol. 119, pp. 953–961; August 1, 1960.)

538.222:538.569.4 281

Paramagnetic Resonance of V^{4+} in TiO_2—H. J. Gerritsen and H. R. Lewis. (*Phys. Rev.*, vol. 119, pp. 1010–1012; August 1, 1960.) Observations at 10.14 and 22.68 Gc indicate that the spectrum is due to single *d* electrons of tetravalent V ions located at titania sites in the lattice.

538.222:538.569.4 282

Cross-Relaxation in Ruby—W. B. Mims and J. D. McGee. (*Phys. Rev.*, vol. 119, pp. 1233–1237; August 15, 1960.) A pulsed microwave method has been used to study relaxation in synthetic ruby. Cross-relaxation effects appear in the form of decay traces having two or more periods.

538.222:538.569.4:621.375.9 283

Chromium-Doped Titania as a Maser Material—H. J. Gerritsen, S. E. Harrison and H. R. Lewis. (*J. Appl. Phys.*, vol. 31, pp. 1566–1571; September, 1960.) A description is given of the properties of a new maser material which can be used to extend the frequency range of masers and should lead to improved amplifiers at the lower frequencies. Energy level diagrams are included for magnetic fields in the (110), ($1\bar{1}0$), (001) and (100) planes.

621.315.55:621.3.032.27 284

Hard Gallium Alloys for Use as Low-Contact-Resistance Electrodes and for Bonding Thermocouples into Samples—G. G. Harman. (*Rev. Sci. Instr.*, vol. 31, pp. 717–720; July, 1960.) Ga alloys are described which can be prepared in a similar way to dental amalgams and packed into cavities in semiconductors to bond wires. Operating temperature range extends to 900° C.

MATHEMATICS

512.3 285

Minimal Positive Polynomials—J. E. Eaton. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, p. 171; March, 1960. Abstract, Proc. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

517.512.71 286

Canonical and Hamiltonian Formalism applied to the Sturm-Liouville Equation—M. A. Biot and I. Tolstoy. (*Quart. Appl. Math.*, vol. 18, pp. 163–172; July, 1960.) The Sturm-Liouville equation is expressed in Hamiltonian form. The method is developed with particular reference to the wave equation; it unifies many apparently diverse treatments, and leads to new procedures.

517.923.4 287

Note on the Solution of Riccati's Differential Equation—H. H. Howe. (*J. Res. NBS*, vol. 64B, pp. 95–98; April–June, 1960.) Solutions given in terms of recurrence formulas are suitable for use in radio propagation problems.

517.942.9:517.949 288

The Stability of Computation of the Pierce-Cauchy Problem—B. Meltzer. (*J. Electronics and Control*, vol. 8, pp. 449–453; June, 1960.) "Limits are found analytically for rate of growth of error in computation—by 'marching' methods—of solutions of the finite-difference form of Laplace's equation in two dimensions, with Cauchy boundary conditions. A series of alternative nonmarching methods are proposed, which should be stable."

519 289

The Kron Method of Tearing and the Dual Method of Identification—A. I. Weinzweig. (*Quart. Appl. Math.*, vol. 18, pp. 183–190; July, 1960.) A precise mathematical formulation of the Kron method for solution of network problems (3582 of 1957) is presented. This establishes the validity of the method, and simplifies and extends it.

MEASUREMENTS AND TEST GEAR

529.786:525.35:523.745 290

Effect of the Solar Activity on the Earth's Rotation—Uyeda, Saburi and Iwasaki. (See 159.)

621.3.018.41(083.74):621.372.412.002 291

Adjusting Methods of the Frequency and its Temperature Coefficient of 100 kc/s GT-Cut Crystals used for the Primary Frequency Standards—Y. Hiruta. (*J. Radio Res. Labs., Japan*, vol. 7, pp. 263–276; May, 1960.) See also 62 above.

621.317.088.6:519.281.3 292

The Practice of Linear and Parabolic Correction of Measured Values—G. Ohl. (*Arch. elekt. Übertragung*, vol. 13, pp. 530–532; December, 1959.) Explicit correction formulas are given, based on the method of least squares.

621.317.3/.4:538.632 293

The Hall Generator and its Application to Measurement Techniques—F. Kuhrt. (*Elektron. Rundschau*, vol. 14, pp. 10–13; January, 1960.) The characteristics of Hall generators are summarized, and some Hall-effect devices and their applications are briefly reviewed. 21 references.

621.317.332.1:621.372.2 294

Investigation of Balanced High-Frequency Transmission Lines—Lauterjung. (See 19.)

621.317.335.3:621.372.8 295

A Graphical Method for Measuring Dielectric Constants at Microwave Frequencies—C. B. Sharpe. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 155–159; March, 1960. Abstract, Proc. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.317.34:621.372.852.12 296

Half-Round Inductive Obstacles in Rectangular Waveguide—Kerns. (See 31.)

621.317.373:621.396.663 297

Goniometer Measuring Systems for High Frequencies: Part 2—Phase Angle Measurement—H. Fricke. (*Arch. tech. Messen*, no. 286, pp. 225–226; November, 1959.) Part 1: 3613 of 1960.

621.317.373.029.6:621.372.632 298

Measurement of Relative Phase Shift at Microwave Frequencies—C. A. Finnila, L. A. Roberts and C. Süsskind. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 143–147; March, 1960. Abstract, Proc. IRE, vol. 48, pt. 1, pp. 1194–1195; June, 1960.)

621.317.39:534.232 299

A Capacitive Probe for the Measurement of Very Small Deflections in the Frequency Range 50c/s–2.0 Mc/s—H. Herold and A. Lenk. (*Hochfrequenz. und Elektroak.*, vol. 68, pp. 152–158; December, 1959.) The design is described of equipment for measuring the amplitude of vibrations, in the range 0.1–1 mμ, of piezoelectric resonators.

621.317.39.029.6:537.525 300

Microwave Method of Investigating the Afterglows of Pulsed Gaseous Discharges—M. C. Sexton, J. J. Lennon and M. J. Mulcahy. (*Brit. J. Appl. Phys.*, vol. 10, pp. 356–359; August, 1959.) Apparatus suitable for recording the decrease of electron density in the afterglows of pulsed microwave discharges at 3 and 10 cm λ is described, which measures the frequency shift induced in a resonant cavity by free electrons. An improved method allows the modulation envelope to be removed by means of a hybrid T-junction, so that only the cavity reflection is fed to the display apparatus.

621.317.42:538.569.4 301

Radio Measurement of a Value of Magnetic Induction and of its Stability in Time—C. Fric and H. Hahn. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 680–682; January 25, 1960.) A method is described for measuring directly the magnetic induction in the air gap of an electromagnet by an instantaneous reading of the frequency of a maser-type self-oscillator [798 of 1960 (Fric)].

621.317.42:538.632 302

Hall-Effect Devices of Small Dimensions—V. Andresciani. (*Note Recensioni Notiz.*, vol. 9, pp. 61–72; January/February, 1960.) The method of construction of devices used for magnetic-field and microwave power measurements is described.

621.317.44:539.23 303

Sensitive Flux Measurement of Thin Magnetic Films—H. J. Oguey. (*Rev. Sci. Instr.*, vol. 31, pp. 701–709; July, 1960.) The difficulties of flux calibration and noise reduction are examined.

621.317.61 304

Measurement of Complex Voltage Ratio—E. R. Wigan. (*Electronic Tech.*, vol. 37, pp. 422–428; November, 1960.) A description is given of a technique for the precise measurement of the vector ratio between two alternating voltages by allowing the apparatus to measure its own errors. The technique is applied to measurements using a "Cartesian-coordinate" ac potentiometer.

621.317.733.011.3 305

Improved Square-Wave Inductance Bridge—B. Howland. (*Rev. Sci. Instr.*, vol. 31, pp. 763–768; July, 1960.) The bridge described is used to determine the elements of an inductor with one square-wave null measurement in the range 0.1–350 kc. A design for use at VHF is also given. Screening problems are simplified by using a wide-band transformer with a coaxial secondary winding.

621.317.79.029.6:621.372.012 306
Analysis of Microwave Measurement Techniques by means of Signal Flow Graphs—J. K. Hunton. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 206–212; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

621.317.79.029.65 307
High-Resolution Millimetre-Wave Fabry-Perot Interferometer—W. Culshaw. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 182–189; March, 1960. Abstract, PROC. IRE, vol. 48, pt. 1, p. 1195; June, 1960.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.317.39:612 308
The Problem of the Electronic Measurement and Recording of Physiological Phenomena in Test Subjects under Physical Stress—W. Nicolai. (*Elektronik*, vol. 9, pp. 5–11; January, 1960. Correction, vol. 9, p. 64; February, 1960.) Equipment for recording and transmitting data on muscle and heart activity of athletes is described, and recordings obtained are discussed.

621.383.8:527 309
Star Tracker uses Electronic Scanning—W. D. Atwill. (*Electronics*, vol. 33, pp. 88–91; September 20, 1960.) A star image on a phototube is scanned electronically, and the error signal, which is generated when the optical axis of the tracker deviates from the line of sight to the star, is used to correct the telescope alignment.

621.384.61 310
Natural Functions of Betatron Oscillations in a Circular Phasotron—Zh. Loshak. (*Zh. Tekh. Fiz.*, vol. 29, pp. 995–1008; August, 1959.) An analysis is made of the natural oscillations in a simplified model of an accelerator with an alternating radial field. The application of the expressions derived for studying "resonance tones" and the effects of distortion of the magnetic field are considered.

621.385.833 311
Coherent Splitting of an Electron Beam by Magnetic Fields—K. Krimmel. (*Z. Phys.*, vol. 158, pp. 35–38; January 25, 1960.) A description is given of a magnetostatic biprism for electron-interferometry applications.

621.385.833 312
Defocusing Contrast in Coherent Illumination: Case of a Periodic Object in Electron Microscopy—M. Fagot and C. Fert. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 94–96; January 4, 1960.)

621.385.833 313
Shot and Scintillation Noise in Cold Electron Emission—C. Kleint and H. J. Gasse. (*Z. Naturforsch.*, vol. 15a, pp. 87–88; January, 1960.) A preliminary note is given on noise measurements in field-emission microscopes.

621.387.462 314
The Individual Counting of β Particles by means of Germanium and Silicon Junction Diodes—H. D. Engler. (*Z. Naturforsch.*, vol. 15a, pp. 82–84; January, 1960.) β-spectra obtained with the diodes described are given.

621.387.462 315
Large-Area Germanium Surface-Barrier Counters—F. J. Walter, J. W. T. Dabbs and L. D. Roberts. (*Rev. Sci. Instr.*, vol. 31, pp. 756–762; July, 1960.) A solid-state counter of good resolution and stability with a fast rise time suitable for counting heavy charged particles is described.

PROPAGATION OF WAVES

621.391.812.3.029.63 316
Computation and Measurement of the Fading Rate of Moon-Reflected U.H.F. Signals—S. J. Fricker, R. P. Ingalls, W. C. Mason, M. L. Stone and D. W. Swift. (*J. Res. NBS*, vol. 64D, pp. 455–465; September/October, 1960.) Calculations of the total libration rate lead to an evaluation of the Doppler spread, from which fading rates can be deduced. Good agreement with observational data at a frequency of 412 Mc is obtained.

621.391.812.6.029.62/64 317
Observational Results of V.H.F., U.H.F. and S.H.F. Propagation Beyond the Radio Horizon—A. Takahira, H. Irie and T. Nakamura. (*J. Radio Res. Labs., Japan*, vol. 7, pp. 197–211; May, 1960.) Propagation experiments have been conducted at 150 and 600 Mc and 3 and 10 Gc over a 125-km land path. During the day, a) in the VHF band, reflected and diffracted waves are superposed, b) in the UHF band, reflected waves predominate, and c) at 3 and 10 Gc scattered waves predominate. At nighttime, reflected waves are predominant in all bands.

621.391.812.62.029.62/.63 318
Studies in the U.H.F. Overland Propagation Beyond the Horizon—M. Hirai, K. Nishikori, M. Fukushima, Y. Kurihara, R. Inoue, M. Ikeda, S. Niwa and Y. Kido (*J. Radio Res. Labs., Japan*, vol. 7, pp. 137–176; May, 1960.) A report is given of a series of experiments conducted over paths of 226 and 345 km on frequencies of 160, 600 and 2_{120} Mc. See also 244 of 1959 (Onoe, *et al.*).

621.391.812.621 319
Methods of Predicting the Atmospheric Bending of Radio Rays—B. R. Bean, G. D. Thayer and B. A. Cahoon. (*J. Res. NBS*, vol. 54D, pp. 487–492; September/October, 1960.) Results derived from radiosonde measurements are incorporated in a comparison analysis to assess the relative accuracy of three methods for the prediction of bending when the refractive-index profile is unknown.

621.391.812.63.029.62 320
Experiments of Long Distance Ionospheric Propagation on V.H.F.—K. Tao, K. Sawaji, A. Sakurazawa and M. Yamaoka. (*J. Radio Res. Labs. Japan*, vol. 7, pp. 177–196; May, 1960.) Results are given of tests at 50 Mc over distances from 1100 to 3380 km during a summer-period of one month. The received signals have been statistically analyzed and classified as propagated by sporadic-E, ionospheric scattering and a combination of both types of propagation.

621.391.812.63.029.63:551.510.535 321
Some Characteristics of Ionospheric Back-Scatter Observed at 440 Mc/s—Pineo, Kraft and Briscoe. (See 188.)

RECEPTION

621.391.82 322
Detection of a Signal Specified Exactly with a Noisy Stored Reference Signal—T. G. Birdsall. (*J. Acoust. Soc Am.*, vol. 32, pp. 1038–1045; August, 1960.) The performances of both the optimum receiver for detecting the presence of a signal in white Gaussian noise and the cross-correlation receiver with storage noise are calculated and compared for a special case.

STATIONS AND COMMUNICATION SYSTEMS

621.391 323
Some Further Theory of Group Codes—D. Slepian. (*Bell. Sys. Tech. J.*, vol. 39, pp. 1219–1252; September, 1960.)

621.394.441 324
A Voice-Frequency Telegraphy System with Narrow-Band Frequency Modulation and Transistors—H. Heller. (*Nachrichtentech. Z.*, vol. 12, pp. 595–601; December, 1959.)

621.396:621.391.812.63 325
A Proposed Technique for F-Layer Scatter Propagation—W. C. Vergara and J. L. Levatich. (PROC. IRE, vol. 48, p. 1790; October, 1960.) The technique proposed is to use frequencies above the classical F_2-layer MUF which are calculated assuming a Gaussian distribution of "elemental" MUF's associated with irregularly ionized volumes in the ionosphere.

621.396.4:523.5 326
Loss in Channel Capacity Resulting from Starting Delay in Meteor-Burst Communication—G. R. Sugar. (*J. Res. NBS*, vol. 64D, pp. 493–494; September/October, 1960.) For idealized specular underdense trails, theory confirms the experimentally observed exponential signal burst-width/amplitude distribution for durations up to 0.5 second, and predicts a channel capacity loss which increases with frequency below 100 Mc.

621.396.4:523.5 327
Elementary Considerations of the Effects of Multipath Propagation in Meteor-Burst Communication—G. R. Sugar, R. J. Carpenter and G. R. Ochs. (*J. Res. NBS*, vol. 64D, pp. 495–500; September/October, 1960.) An examination is made of the effects of a) the simultaneous existence of two meteor trails, b) the existence of a Rayleigh-fading background continuum, and c) the existence of two first Fresnel zones along a single meteor trail.

621.396.43:629.19 328
Scattering by a Spherical Satellite—E. M. Kennaugh, S. P. Morgan and H. Weil. (PROC. IRE, vol. 48, p. 1781; October, 1960.) A critical comment is made on 2515 of 1960 (Vea, *et al.*).

621.396.43:629.19 329
Scattering Properties of Large Spheres—N. A. Logan. (PROC. IRE, vol. 48, p. 1782; October, 1960.) A critical comment is made on 2515 of 1960 (Vea, *et al.*).

621.396.65 330
7-Gc/s Miniature Radio Link—R. Schienemann. (*Telefunken Ztg.*, vol. 32, pp. 251–264; December, 1959. English summary, pp. 292–293.) Details are given of the design and performance of radio-link equipment which provides six speech or teletype channels and one service channel. The range is 50 km, but can be extended to 300 km by means of relay stations.

621.396.74.029.62(43) 331
List of V.H.F. Transmitters in the German Federal Republic and the Soviet-Occupied Zone—(*Rundfunktech. Mitt.*, vol. 3, pp. 293–296; December, 1959.) The lists and map of transmitter locations give the position as at November 1, 1959, For similar data on short-wave and long- and medium-wave transmitters, see *ibid.*, pp. 297–301.

621.396.931 332
Communication System for Highway Traffic Control—E. A. Hanyaz, J. E. Stevens and A. Meduvsky. (*Electronics*, vol. 33, pp. 81–83; October 14, 1960.) A VLF loop system at the

roadside transmits messages to passing cars. A tape recorder transmits continuous fixed messages, using a pick-up system which exerts no frictional force on the tape.

621.396.97:534.76 333

Stereophonic Transmission of Broadcasts using F. M. Signals and an A. M. Subcarrier—F. L. H. M. Stumpers and R. Schutte. (*Elektron. Rundschau*, vol. 13, pp. 445–446; December, 1959.) A compatible method is described which involves little expenditure on the receiver side and requires only a narrow bandwidth.

621.396.97:534.76 334

The P.A.M. Method of Stereophonic Broadcasting—G. Janus. (*Elektron. Rundschau*, vol. 13, pp. 447–449; December, 1959.) A single-channel, compatible method is described.

SUBSIDIARY APPARATUS

621–526:681.142 335

Analyzing Magnetically Detented Stepper Servo Motors—H. J. Weber and M. Weiss. (*Electronics*, vol. 33, pp. 71–74; September 23, 1960.) Digital test equipment for evaluating the performance of a motor used for digital-to-analog conversion is described.

621.3.087.4:621.395.625.3 336

Recording Accuracy in Magnetic Tape Recording—H. Völz. (*Elektron. Rundschau*, vol. 14, pp. 23–25; January, 1960.) The accuracy is defined on the basis of information theory, and the capacity of a magnetic-tape communication channel is calculated taking account of signal-to-noise ratio and undesirable AM and FM effects.

621.311.69:621.383.2 337

Photovoltaic Effect Produced in Silicon Solar Cells by X and Gamma Rays—K. Scharf. (*J. Res. NBS*, vol. 64A, pp. 297–307; July/August, 1960.)

621.316.722 338

On the Realization of a Highly Stable Direct-Current Power Supply with a Wide Range of Output Voltage—G. Giralt and J. Lagasse. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 91–93; January 4, 1960.) The principle of operation of a stabilized power supply incorporating two feedback loops is described. The properties of the circuit are deduced from transfer functions.

621.316.722.078:621.382.3 339

Designing Transistorized Voltage Regulators—E. Wilson. (*Electronics*, vol. 33, pp. 62–65; September 23, 1960.) A detailed analysis is made of the design of a dc series regulator with temperature compensation.

TELEVISION AND PHOTOTELEGRAPHY

621.397.132 340

Color Television Standards—R. D. A. Maurice. (*Wireless World*, vol. 66, pp. 536–537; November, 1960.) The ½-line-frequency offset commonly used for the NTSC-type color subcarrier is considered inferior to ⅓-line-frequency "precision best offset" from the point of view of co-channel interference. It is proposed that the color subcarrier frequency for the 625-line system should be changed from 4.4296875 Mc to 4.430800 Mc and, that the number of lines per picture be changed from 625 to 627.

621.397.132:535.62 341

Does Dr. Edwin H. Land's Theory of Colour Vision Form the Basis for a New Colour Television System?—P. Neidhardt. (*Elektron. Rundschau*, vol. 13, pp. 451–457; December, 1959.) The two-component theory of color vision [see, *e.g.*, 2677 of 1960 (Woolfson)] cannot provide a satisfactory basis for a color television system.

621.397.132:621.372.55 342

The Vestigial-Sideband Equalization of the Chrominance Signal in the NTSC System—H. Schönfelder. (*Arch. elekt. Übertragung*, vol. 14, pp. 37–46; January, 1960.) Crosstalk and frequency-response distortion due to vestigial-sideband transmission of the *I*-component can be eliminated by means of equalizer circuits in the color television receiver. Such circuits are described, and their effect on picture quality is considered.

621.397.132:621.397.62 343

Generator of Synchronizing Signals for the Color Subcarrier—L. Accardi. (*Note Recensioni Notiz.*, vol. 8, pp. 653–662; November/December, 1959.) A circuit is described which is capable of producing bursts of oscillation conforming to the waveform requirements of the NTSC system.

621.397.132:621.397.62 344

Demodulation of Chrominance Signals in Color Television Receivers—C. Massari. (*Note Recensioni Notiz.*, vol. 9, pp. 73–112; January/February, 1960.) The design of decoding circuits for the NTSC system is reviewed.

621.397.132.001.4 345

Color Bar Generator for the NTSC Standard—G. Bolle. (*Telefunken Ztg.*, vol. 32, pp. 237–243; December, 1959. English summary, pp. 291–292.) In the equipment described, closely coordinated luminance and chrominance signals are produced, and the sequence of colors can be changed at will.

621.397.332 346

Synchronizing System for Two Sinusoidal Voltages of Nearly Equal Frequency Applied to the Scanning of Camera Tubes—C. Curie and B. Cazeneuve. (*C.R. Acad. Sci., Paris*, vol. 250, pp. 491–493; January 18, 1960.) A scanning system is proposed based on the properties of Lissajous figures. Two sinusoidal signals at frequencies of 16.17 and 16.20 kc, derived by frequency division from a single quartz oscillator, are used to obtain a scanning rate of 30 pictures/second. Details are given of the oscillator and frequency-division circuits. The "definition" obtained on a closed-circuit system was comparable to that of a normal television image.

621.397.61 347

A Vertical-Aperture Equalizer for Television—W. G. Gibson and A. C. Schroeder. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 395–401; June, 1960.) General description of a vertical-aperture equalizer which, by delaying the video signal by one scanning line, increases the subjective sharpness of the television picture. See also 1885 of 1956 (Schroeder and Gibson).

621.397.61 348

The Minimum Distances between Interfering Television Transmitters—H. Eden and K. H. Kaltbeitzer. (*Rundfunktech. Mitt.*, vol. 3, pp. 271–276; December, 1959.) A method of calculating the minimum geographical separation between transmitters is given. For English version, see *E.B.U. Rev.*, no. 58A, pp. 14–18, December, 1959.)

621.397.62 349

Automatic Matching of Contrast and Background Brightness in Television Receivers to Room Lighting—R. Suhrmann. (*Elektron. Rundschau*, vol. 13, pp. 441–444; January, 1960. Correction, vol. 14, p. 33; December, 1959.) Photocell circuits for automatic contrast control are given, and methods are discussed for raising screen brightness in proportion to ambient lighting, with or without a threshold level for the control action.

621.397.62.001.4 350

Equipment for the Subjective Testing of Television Receivers—H. F. Lelgemann. (*Telefunken Ztg.*, vol. 32, pp. 244–250; December, 1959. English summary, p. 292.) The equipment described enables the images produced by two receivers to be displayed in adjacent sections on the same screen; this is done by means of an electronic switch, and facilitates image comparison.

621.397.621 351

Flywheel-Synchronized Horizontal-Deflection Systems in Television Receivers under Conditions of Fluctuating Line Frequency—H. Grosskopf and H. Springer. (*Radio Mentor*, vol. 26, pp. 269–271; April, 1960.) The problem of minimizing jitter on indirectly synchronized receivers is discussed. Establishment of tolerance limits for the sensitivity of receivers to line-frequency fluctuations would be desirable.

621.397.7 352

Television Switching Centre Frankfurt/Main—K. Thöm. (*Rundfunktech. Mitt.*, vol. 3, pp. 257–259; December, 1959.) Planning and design of a regional video-signal switching center are discussed; a sound-signal switching center for the same network has been in operation for some time.

621.397.7 353

Portable Television Outside-Broadcast Equipment—E. Legler. (*Rundfunktech. Mitt.*, vol. 3, pp. 253–256; December, 1959.) The equipment described comprises a vidicon camera, pulse generator and band-IV transmitter with antenna; it weighs 11.4 kg, and can be carried by one person.

621.397.7–182.3 354

The Equipment of a New Television Outside-Broadcast Vehicle for the N.R.W.V. Hamburg—G. Schadwinkel. (*Rundfunktech. Mitt.*, vol. 3, pp. 263–265; December, 1959.) Details additional to those contained in 1037 of March (Schadwinkel and Käding) are given.

621.397.7(43) 355

The Television Network of the German Federal Republic and the Soviet-Occupied Zone—(*Rundfunktech. Mitt.*, vol. 3, pp. 285–292; December, 1959.) Tabulated data on television transmitters as at November 1, 1959, with maps showing their location and that of television links.

TUBES AND THERMIONICS

621.382 356

The Oscillistor—New Type of Semiconductor Oscillator—R. D. Larrabee and M. C. Steele. (*J. Appl. Phys.*, vol. 31, pp. 1519–1523; September, 1960.) A new magneto-oscillatory effect has been observed in the electron-hole plasma within a semiconductor. When the specimen is subjected to an electric field and a magnetic field, current oscillations can be detected across a series load resistance.

621.382(083.74) 357

I.R.E. Standards on Solid-State Devices: Definitions of Semiconductor Terms, 1960—(Proc. IRE, vol. 48, pp. 1772–1775; October, 1960.) Standard 60 IRE 28.S1.

621.382.23 358

Temperature Dependence of Tunnel-Diode Characteristics—Y. Furukawa. (*J. Phys. Soc. Japan*, vol. 15, p. 1130; June, 1960.) The temperature dependence of the maximum current was measured as a function of carrier concentration. Appreciable differences were observed between As- and Sb-doped samples. Results are briefly explained.

621.382.23:621.373.029.64/.65 359

Esaki-Diode Oscillators from 3 to 40 Gc/s—R. Trambarulo and C. A. Burrus. (Proc. IRE, vol. 48, pp. 1776–1777; October, 1960.) The maximum power observed was 50 μw at 16.7 Gc using GaAs diodes.

621.382.3 360

Noise of Germanium and Silicon Transistors at High Current Densities—B. Schneider and M. J. O. Strutt. (*Arch. elekt. Übertragung*, vol. 13, pp. 495–502; December, 1959.) An equivalent circuit is derived for a *p-n* junction under high-current conditions, which comprises an inductance as well as resistances and capacitances. Formulas are obtained for shot noise, which give results in good agreement with experimental values.

621.382.3 361

A Study of the Charge Control Parameters of Transistors—J. J. Sparkes. (Proc. IRE, vol. 48, pp. 1696–1705; October, 1960.) The most significant performance parameters are considered to be the collector time constant, the saturation time constant, the "on demand" current gain, the collector capacitance charge and the dc current gain. Their variation with dc level is considered.

621.382.3:621.374.3 362

The Transistor as a Passive Circuit Element—A. Darré. (*Frequenz*, vol. 14, pp. 6–10; January, 1960.) The use of transistors in alternating-voltage switching circuits and in logic circuits is considered.

621.382.3:621.391.822 363

Shot and Thermal Noise in Germanium and Silicon Transistors at High-Level Current Injections—B. Schneider and M. J. O. Strutt. (Proc. IRE, vol. 48, pp. 1731–1739; October, 1960.) A theoretical treatment is given for the case of shot noise at high current densities; formulas in good agreement with experimental values are obtained. A new equivalent circuit for *p-n* junctions is employed.

621.382.3.012.8 364

Equivalent Circuits of the Transistor used as a Linear Amplifier—W. Benz. (*Elektron. Rundschau*, vol. 14, pp. 5–9 and 59–64; January and February, 1960.) A review is given of equivalent circuits representing the idealized transistor, and of simplified equivalent circuits for practical application.

621.382.3.078:621.316.825 365

The Use of the Silicon Resistor in the D.C. Stabilization of Transistor Circuits—D. H. Mehrtens, and J. T. Zakrewski. (*Electronic Engrg.* vol. 32, pp. 624–629; October, 1960.) The properties are outlined, and applications to both small-signal and power transistors are given.

621.382.33 366

Analytical Studies on Effects of Surface Recombination on the Current Amplification Factor of Alloy-Junction and Surface-Barrier Transistors—T. Sugano and H. Yanai. (Proc. IRE, vol. 48, pp. 1739–1749; October, 1960.) Analytical formulas giving the survival factor of minority carriers are derived for various geometries of alloy-junction and similar transistors. Useful design data are obtained, such as the optimum radius ratio of the emitter electrode and the collector.

621.382.333 367

Matz Theory of the Junction Transistor at High Injection Levels—M. Drăgănescu. (*J. Electronics and Control*, vol. 8, pp. 459–462; June, 1960.) A simplification of the Matz theory (1815 of 1960) is presented.

621.382.333.33 368

The Emitter Diffusion Capacitance of Drift Transistors.—J. Lindmayer and C. Wrigley. (Proc. IRE, vol. 48, pp. 1777–1778; October, 1960.) The relation between the capacitance calculated for the common-base configuration and that measured on the common-emitter configuration is determined.

621.382.333.3:621.372.632 369

Parametric Amplification Properties in Transistors—R. Zuleeg and V. W. Vodicka. (Proc. IRE, vol. 48, pp. 1785–1786; October, 1960.) See 4420 of 1960 (Vodicka and Zuleeg).

626.383.292 370

A Grid-Controlled Electron-Multiplier Tube—D. C. Brown and M. F. Penny. (*J. Electronics and Control*. vol. 8, pp. 431–439; June, 1960.) "The properties of an experimental grid-controlled photomultiplier tube were investigated, and, under pulsed operation, a transfer conductance of 60 mA/V was achieved. The primary electron current was produced by placing a suitable scintillator near the photocathode and energizing it by means of a radioactive source of ^{90}Sr."

621.385.032.213.13 371

Electron Emission Current from Oxide-Coated Cathode subject to High Electric Field—C. Shibata. (*J. Phys. Soc Japan*, vol. 15, p. 1124; June, 1960.) The relation between current and applied field is found to differ experimentally from that of the Schottky effect.

621.385.032.213.13:621.317.332 372

A New Method for the Measurement of Cathode Interface Impedance—J. Tamiya. (*Rev. Sci. Instr.*, vol. 31, pp. 696–700; July, 1960.) The tube containing the cathode interface impedance to be measured, forms a bridge with a reference tube and a variable RC circuit. Measurement is made by balancing simultaneously at 200 kc, 1.1 and 5 Mc.

621.385.032.213.13:621.317.332 373

Nonlinear Effect upon the Equivalent Circuit of Interface Impedance in Pulse Methods—J. Tamiya. (*Rev. Sci. Instr.*, vol. 31, pp. 786–788; July, 1960.) A quantitative analysis is made of the effect of nonlinear tube characteristics and rectification in the cathode layers. These introduce spurious RC elements in the equivalent circuit.

621.385.1.004.6 374

Mathematical and Statistical Methods of Considering Lifetime Data and their Application to Thermionic Valves—A. Deixler and E. Rusch. (*Nachrichtentech. Z.*, vol. 12, pp. 613–618; December, 1959.) The validity of a mathematical law of lifetime is examined on the basis of a statistical model test. A comparison with actual tube life data leads to the proposal of simplified statistical reliability concepts.

621.385.623.5:621.376.23.029.65 375

Use of Reflex Klystrons as Millimetre-Wave Detectors—K. Ishii. (*Electronics*, vol. 33, pp. 82–83; September 9, 1960.) A gain of 12 db over conventional crystal detectors can be obtained when the klystron is used as a regenerative detector.

621.385.624 376

Instability Effects in High-Power C.W. Klystrons—J. C. Vokes and C. P. Lea-Wilson. (*J. Electronics and Control*, vol. 8, pp. 401–419; June, 1960.) Instability effects observed in a four-cavity 2-kw X-band klystron amplifier with an efficiency of 40 per cent are described. An explanation is given accounting for all the major effects observed.

621.385.63:621.375.9:621.372.44 377

The D.C. Pumped Quadrupole Amplifier—a Wave Analysis—A. E. Siegman. (Proc. IRE, vol. 48, pp. 1750–1755; October. 1960.) An analysis is given of the electron-beam amplifier which is pumped by direct voltage via a twisted quadrupole structure. Amplification is found to result from the coupling of the slow and fast cyclotron waves. The device appears promising as a microwave amplifier.

621.385.632.19:621.372.853 378

Nonperiodic Slow-Wave and Backward-Wave Structures—Clarricoats and Waldron. (See 33.)

621.385.633.14 379

Parasitic Modulation Phenomena in Long-Beam Tubes of the Carcinotron O Type—D. Reverdin. (*Ann. Radioélect.*, vol. 15, pp. 147–168; April, 1960. English abridgment, pp. 203–204.) Results of an experimental investigation show that ionic plasma oscillations are the source of spurious modulation effects. These effects may be reduced by a) improving the vacuum, and b) using ion and slow-electron traps.

MISCELLANEOUS

621.37/.38:539.12.04 380

Radiation Pulses and Electronics—J. W. Clark and T. D. Hanscome. (*Nucleonics*, vol. 18, pp. 74–77; September, 1960.) The effects of radiation pulses acting on electronic components are considered. A distinction is made between ionization and atomic displacement effects, and the general design principles for radiation-resistant circuits are outlined.

Translations of Russian Technical Literature

Listed below is information on Russian technical literature in electronics and allied fields which is available in the U. S. in the English language. Further inquiries should be directed to the sources listed. In addition, general information on translation programs in the U. S. may be obtained from the Office of Science Information Service, National Science Foundation, Washington 25, D. C., and from the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C.

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Solid State Physics (Fizika Tverdogo Tela)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
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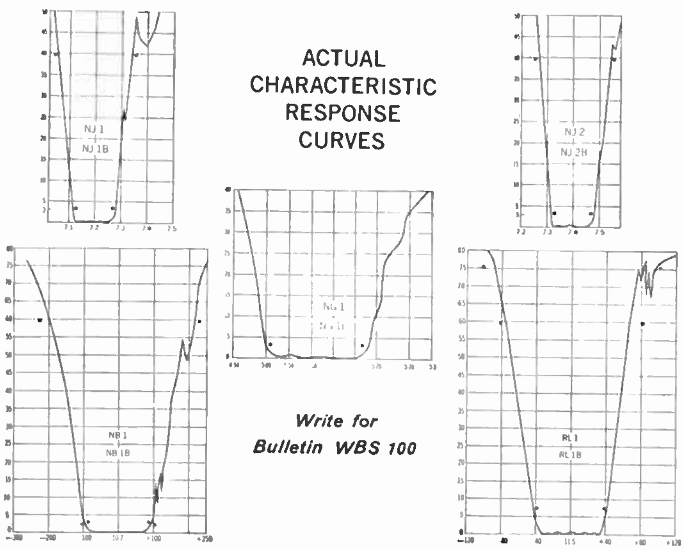
Type	Center Freq.	3db Bandwidth Minimum	40db Bandwidth Max.	60db Bandwidth Max.	75db Bandwidth Max.	Ultimate Discrim. Minimum	Insertion Loss Max.	Impedance ohms	Inband Ripple Max.	Package Type
NJ-1	7.2MC	160KC	300KC			60db	6db	13K	1db	A
NJ-1B	7.2MC	160KC	300KC			60db	6db	13K	.5db	B
NJ-2	7.4MC	160KC	300KC			60db	6db	13K	1db	A
NJ-2B	7.4MC	160KC	300KC			60db	6db	13K	.5db	B
NG-1	5.09MC	160KC	350KC			60db	6db	20K	1db	A
NG-1B	5.09MC	160KC	350KC			60db	6db	20K	1db	B
NB-1	10.7MC	200KC		450KC		75db	12db	50	1db	A
NB-1B	10.7MC	200KC		450KC		85db	8db	50	.5db	B
RL-1	11.5MC	80KC		160KC	200KC	85db	6db	50	.5db	C
RL-1B	11.5MC	80KC		160KC	200KC	90db	5db	50	.5db	B

Operating Temp.: —55°C to +90°C Shock: 100g Vibration: 15g to 2KC Units hermetically sealed



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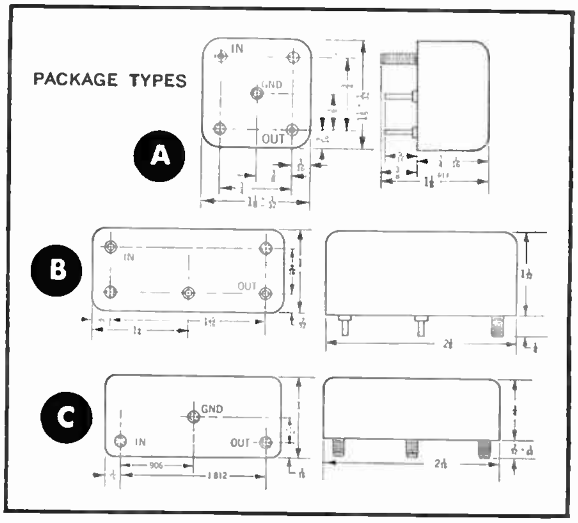
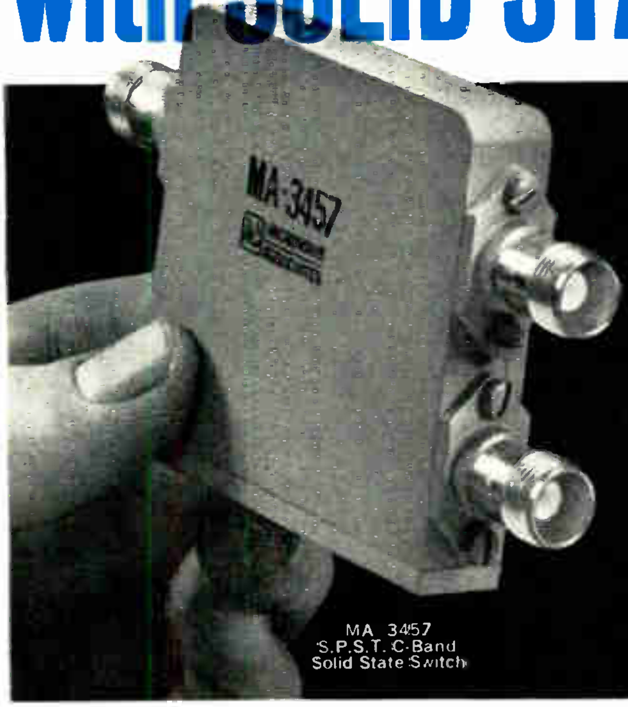


2 NANOSECOND MICROWAVE SWITCHING with SOLID STATE RELIABILITY



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LOW POWER LEVEL COAXIAL SWITCHES

Frequency (Mc)	Insertion Loss (Max)	Isolation (Min)	Switching Power
210-240	0.2 db	20 db	10 mw
260-340	0.2 db	18 db	10 mw
400-500	0.3 db	20 db	10 mw
570-630	0.3 db	20 db	10 mw
900-1000	0.3 db	20 db	10 mw
1250-1350	0.5 db	20 db	10 mw
MEDIUM POWER LEVEL COAXIAL SWITCHES			
200-1000	1.5 db	22 db	70 mw
1000-2000	1.5 db	20 db	70 mw
2000-4000	2.0 db	16 db	70 mw

LOW POWER LEVEL VOLTAGE VARIABLE ATTENUATORS

Frequency (Mc)	Attenuation Range
260-340	0.2 db-18 db
400-450	0.3 db-20 db
570-630	0.3 db-20 db
1250-1350	0.5 db-20 db

Narrow-band higher frequency units are available with lower loss and increased isolation.

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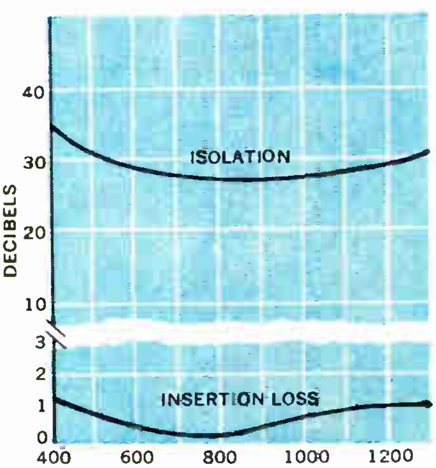
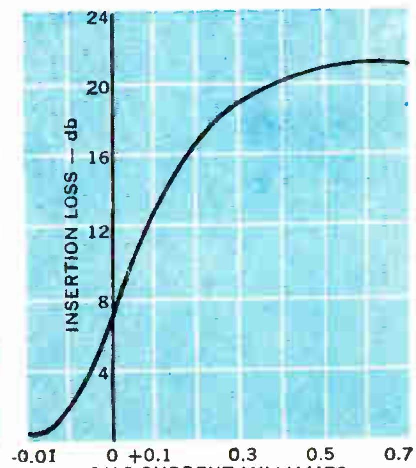
SWITCH SPST



TYPICAL PERFORMANCE
VOLTAGE VARIABLE ATTENUATOR
(425 Mc ± 25)



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Two 50 mm wide channels . . . separate floating input DC amplifiers . . . 4 chart speeds . . . mv or volt inputs

Operate this 1-cubic-foot recorder vertically, horizontally, or tilted at a 20° angle on carrying handle. Inputs are floating and guarded . . . 12 sensitivities from 0.5 mv/mm to 20 v/cm . . . response DC to 125 cps within 3 db, at 10 div peak-to-peak . . . max. non-linearity 0.25 mm . . . common mode rejection 140 db min. DC . . . built-in 10 mv calibration signal and electrical limiting . . . internal 1 sec. timer . . . monitor output connectors for each channel. Galvanometers are rugged, low impedance type with velocity feedback damping; most circuitry for each channel is mounted on a single, easily serviced card.

1 Channel . . . 20 lbs., briefcase size . . . 10 mv/div DC Model . . . 10 uv rms/div AC strain gage Model

Extremely compact, highly versatile recorders for general purpose DC inputs (Model 299) and AC strain gage recording (Model 301). Two chart speeds: 5 and 50 mm/sec . . . inkless, rectangular coordinate recording . . . response from DC to 100 cps within 3 db, at 10 div peak-to-peak . . . gain stability better than 1% to 50°C and for line voltage variation from 103 to 127 volts. Model 299 has balanced to ground input, 10 switch-selected sensitivities, calibrated zero suppression. Model 301 has wide sensitivity ranges, can be used with strain gages and inductive transducers, provides excitation voltage of approximately 4.5 volts rms at 2400 cps, and has uncalibrated zero suppression.

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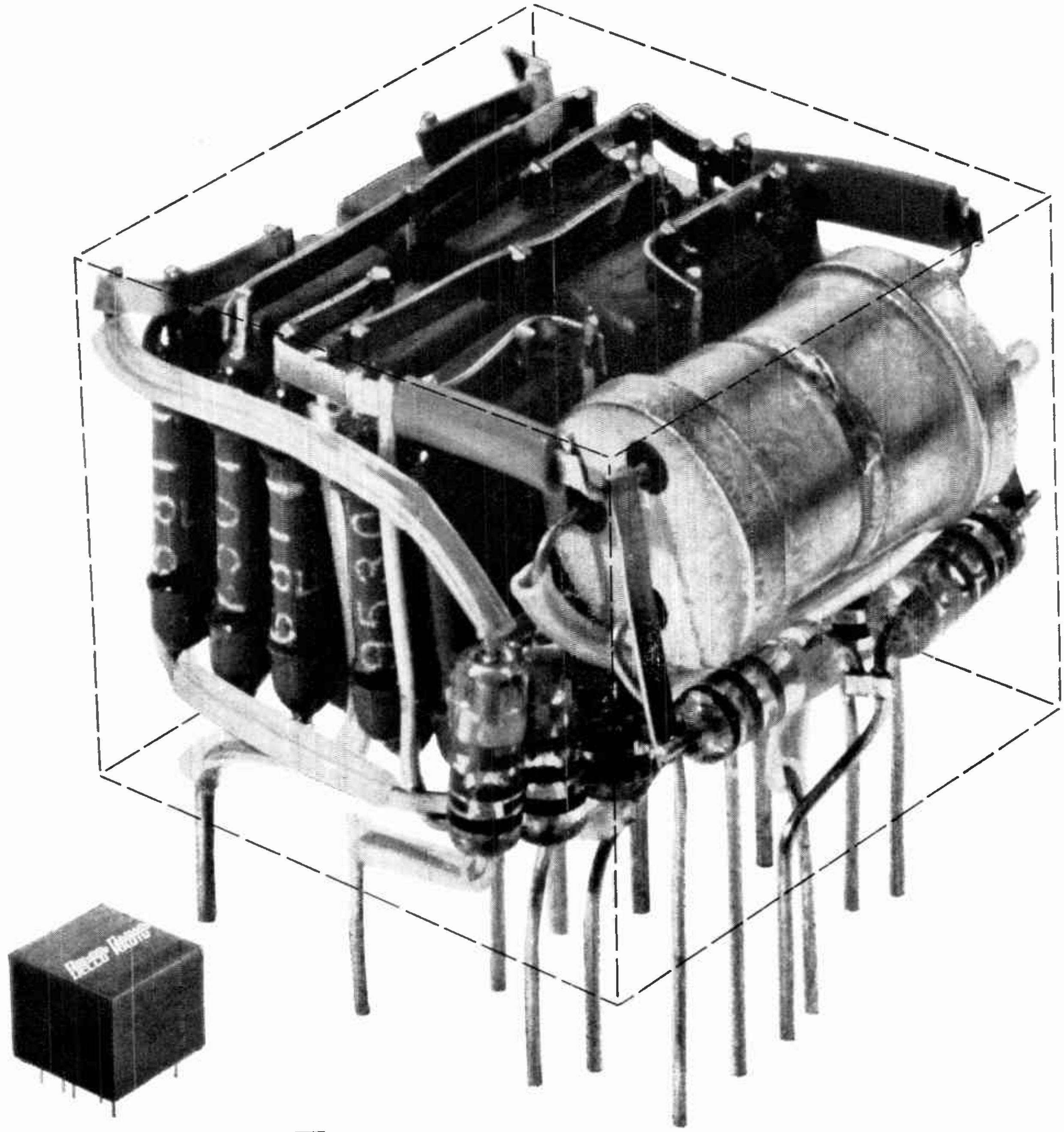

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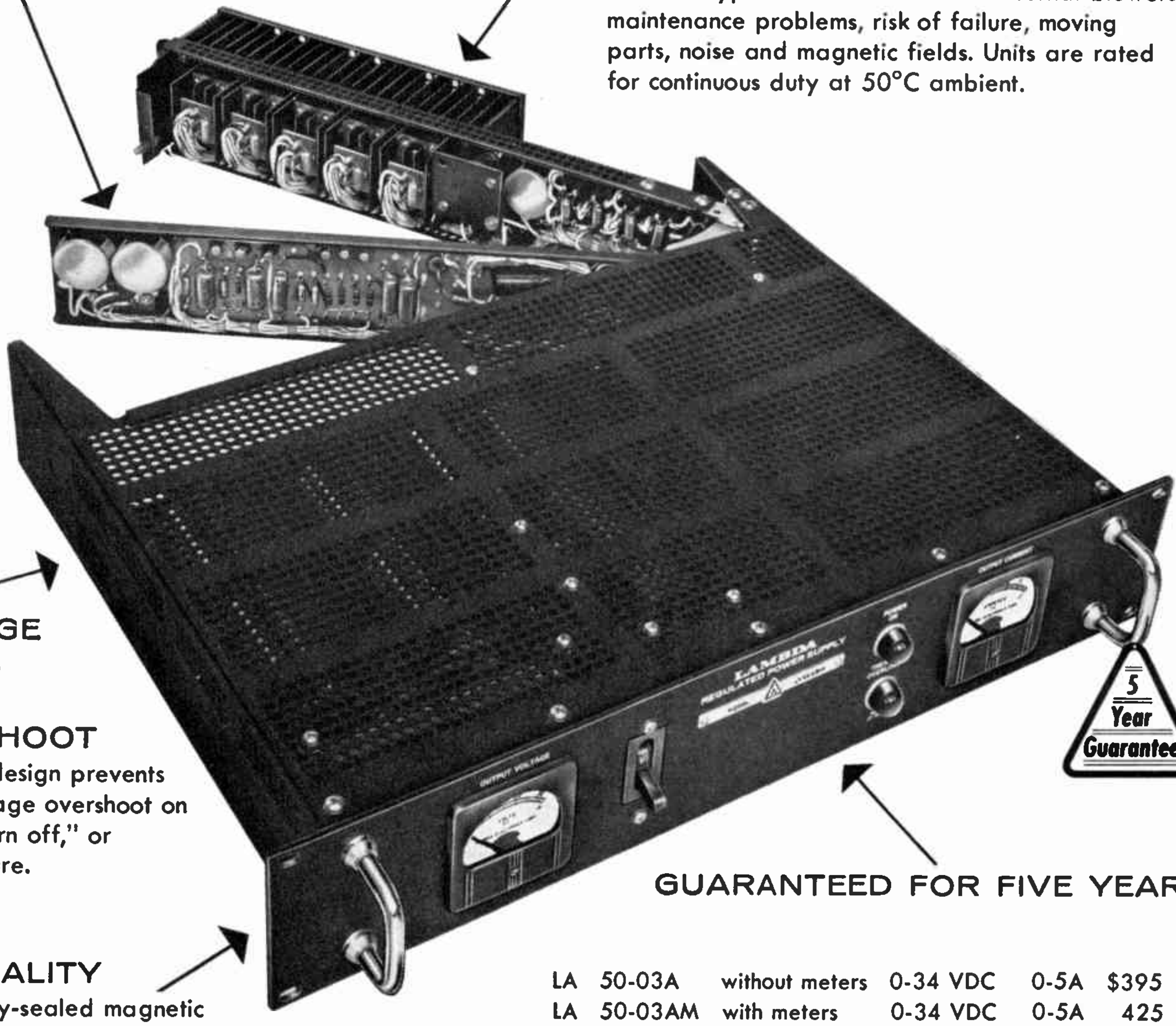

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GUARANTEED FOR FIVE YEARS

MIL QUALITY
Hermetically-sealed magnetic shielded transformer designed to MIL-T-27A quality and performance. Special, high-purity foil, hermetically-sealed long life electrolytic capacitors.

LA 50-03A	without meters	0-34 VDC	0-5A	$395
LA 50-03AM	with meters	0-34 VDC	0-5A	425
LA 100-03A	without meters	0-34 VDC	0-10A	510
LA 100-03AM	with meters	0-34 VDC	0-10A	540
LA 200-03A	without meters	0-34 VDC	0-20A	795
LA 200-03AM	with meters	0-34 VDC	0-20A	825

COMPACT
NO BLOWERS

5 AMP 3½" HIGH
10 AMP 7" HIGH
20 AMP 10½" HIGH

Patent Pending

Lambda LA Series Power Supplies are compact, convection cooled and rated for continuous duty at 50°C ambient temperature.

LAMBDA Transistorized 5 and 10 AMP LA Series

COMPLETE SPECIFICATIONS OF LAMBDA LA SERIES

DC OUTPUT (Regulated for line and load)

Model	Voltage Range[1]	Current Range[2]	Price
LA 50-03A	0-34 VDC	0- 5A	$395
LA 50-03AM	0-34 VDC	0- 5A	$425
LA100-03A	0-34 VDC	0-10A	$510
LA100-03AM	0-34 VDC	0-10A	$540
LA200-03A	0-34 VDC	0-20A	$795
LA200-03AM	0-34 VDC	0-20A	$825

[1]The output voltage for each model is completely covered in four steps by selector switches plus vernier control and is obtained by summation of voltage steps and continuously variable DC vernier as follows:

MODEL	VOLTAGE STEPS
LA 50-03A, LA 50-03AM	2, 4, 8, 16 and 0-4 volt venier
LA100-03A, LA100-03AM	2, 4, 8, 16 and 0-4 volt venier
LA200-03A, LA200-03AM	2, 4, 8, 16 and 0-4 volt venier

[2]Current rating applies over entire output voltage range

Regulation (line) ... Better than 0.05 per cent or 8 millivolts (whichever is greater). For input variations from 100-130 VAC.

Regulation (load) Better than 0.10 per cent or 15 millivolts (whichever is greater). For load variations from 0 to full load.

Transient Response (line) ..... Output voltage is constant within regulation specifications for step function line voltage change from 100-130 VAC or 130-100 VAC.

Transient Response (load) .... Output voltage is constant within regulation specifications for step-function load change from 0 to full load or full load to 0 within 50 microseconds after application.

Internal Impedance ... LA 50-03A less than .008 ohms
LA100-03A less than .004 ohms
LA200-03A less than .002 ohms

Ripple and Noise ... Less than 1 millivolt rms with either terminal grounded.

Polarity .... Either positive or negative terminal may be grounded.

Temperature Coefficient .... 0.025 %/°C

AC INPUT

.... 100-130 VAC, 60 ± 0.3 cycle[3]
LA 50-03A ... 360 watts[4]
LA100-03A 680 watts[4]
LA200-03A . 1225 watts[4]

[3]this frequency band amply covers standard commercial power lines in the United States and Canada.
[4]with output loaded to full rating and input at 130 VAC.

AMBIENT TEMPERATURE AND DUTY CYCLE

... Continuous duty at full load up to 50°C (122°F) ambient.

OVERLOAD PROTECTION:

Electrical .... Magnetic circuit breaker front panel mounted. Special transistor circuitry provides independent protection against transistor complement overload. Fuses provide internal failure protection. Unit cannot be injured by short circuit or overload.

Thermal .... Thermostat, manual reset, rear of chassis. Thermal overload indicator light front panel.

INPUT AND OUTPUT CONNECTIONS

.... Heavy duty barrier terminal block, rear of chassis. 8 foot, 3 wire detachable line cord.

METERS

.... Voltmeter and ammeter on metered models.

CONTROLS:

DC Output Controls Voltage selector switches and adjustable vernier-control rear of chassis.

Power .... Magnetic circuit breaker, front panel.

Remote DC Vernier ... Provision for remote operation of DC Vernier.

Remote Sensing .... Provision is made for remote sensing to minimize effect of power output leads on DC regulation, output impedance and transient response.

PHYSICAL DATA:

Mounting .... Standard 19" Rack Mounting

Size	LA 50-03A	3½" H x 19" W x 14⅜" D	
	LA100-03A	7" H x 19" W x 14⅜" D	
	LA200-03A	10½" H x 19" W x 16½" D	
Weight	LA 50-03A	55 lb Net	85 lb Ship. Wt.
	LA100-03A	100 lb Net	130 lb Ship. Wt.
	LA200-03A	140 lb Net	170 lb Ship. Wt.

Panel Finish .... Black ripple enamel (standard). Special finishes available to customers specifications at moderate surcharge. Quotation upon request.

Send for complete Lambda Catalog.

LAMBDA ELECTRONICS CORP.
515 BROAD HOLLOW ROAD, HUNTINGTON, L. I., NEW YORK 516 MYRTLE 4-4200

LA114

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

MEDICAL ENGINEER

Engineer or physicist to help with clinical applications and research associated with artificial heart-lung apparatus and development of artificial heart to replace the human heart. Advanced degree and/or experience desirable. Salary to $10,000. Large University Hospital. Send resume to Box 2036.

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor of E.E., M.S. or Ph.D. required. Openings in electronics and microwave theory. Beginning salary approximately $9,000-$10,000 per year depending on qualifications. Employment effected according to Civil Service regulations. Write, Head, Dept. of E.E., Institute of Technology (Air University), Wright-Patterson AF Base, Ohio.

INERTIAL SYSTEM TEST ENGINEER

Test and evaluate early prototype inertial systems. Design and develop test methods, equipment, and procedures including precision voltage analog measurements and servo, gyro, and accelerometer performance. Must have degree and applicable experience. Send resume to Mr. D. Krause, Litton Systems, Inc., 336 N. Foothill Rd., Beverly Hills, Calif.

TEST EQUIPMENT DESIGN (SYSTEM CHECKOUT)

System and circuit analysis and circuit design for automatic and manual ground-based and airborne checkout equipment. Must have degree and applicable experience. Send resume to Mr. D. Krause, Litton Systems, Inc., 336 N. Foothill Rd., Beverly Hills, Calif.

TEACHERS

Expanding graduate work in Electrical Engineering at the University of Tennessee will require the services of 2 new staff members with Ph.D. in E.E. or allied fields. A combination of teaching and research is available so that the income should be comparable with industrial salaries. Write to P. C. Cromwell, Head, E.E. Dept., University of Tennessee, Knoxville, Tenn.

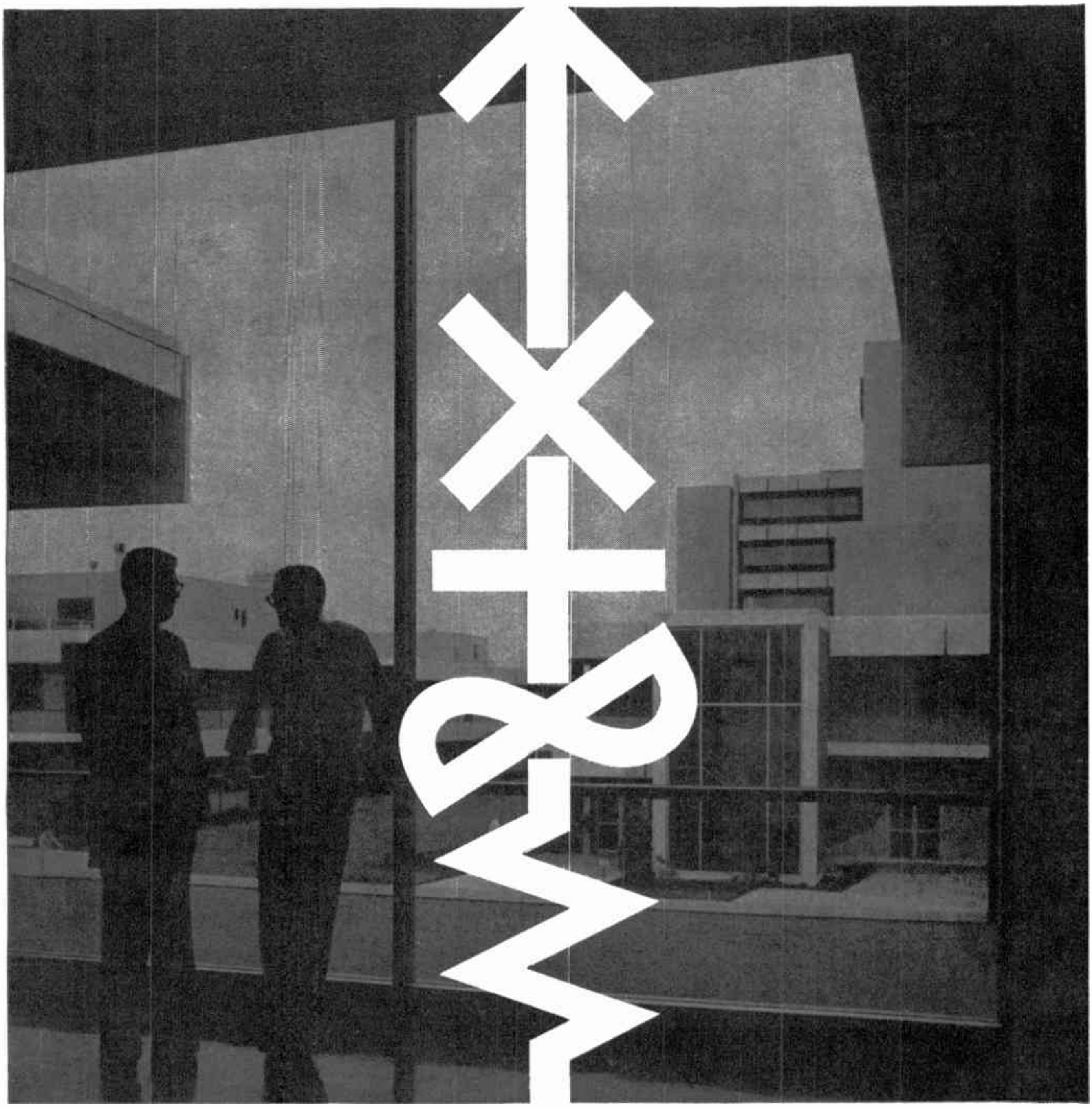
AUDIO ENGINEER

Design and development, circuitry and some mechanical design in tape and disc music recording and allied areas. Excellent opportunity with leading electronic manufacturer. BSEE degree or equivalent. 1-3 years' experience preferred, interest in music. Attractive salary and fringe benefits. Send resume to Box 2037.

OVERSEAS POSITIONS

Robert College, Istanbul, Turkey, presents a challenge in education in a vital part of the world. West and Middle East are contributing side by side to the development of a young and vigorous nation. Opportunities are available in

(Continued on page 102A)



Positions Open

(Continued from page 98A)

engineering, business administration and economics, the sciences, and the humanities. Graduate degrees required. Address inquiries to Dr. Howard P. Hall, Dean of Faculty, Robert College, Bebek Post Box 8, Istanbul, Turkey with copy to Near East College Assoc., 548 Fifth Ave., New York 36, N. Y.

RESEARCH PROFESSOR

Electronics engineer with B.S. or M.S. degree plus radar design experience to conduct research in weather radar. Salary dependent on qualifications and experience. Contact H. W. Hiser, Head, Radar Meteorological Research, Marine Lab., University of Miami, Coral Gables, Florida.

TEACHING—ELECTRICAL ENGINEERING

Teaching in Electrical Engineering. Background in power distribution, rotating machinery, and/or automatic control theory. General undergraduate E.E. program. Prefer M.S. or B.S. with experience. Appointment effective February 1961. Address: Head, Electrical Engineering Dept., College of the Pacific, Stockton, Calif.

ASSISTANT DIRECTORS & SENIOR SCIENTIFIC OFFICERS

A growing Institute, the national laboratory in India for electronics, has 4 senior research positions open, with opportunities to lead research and development groups in the fields of vacuum tubes, electronic instrumentation, communications and solid state devices. A Ph.D. degree or M.S. with experience in industrial research is required. Attractive housing and good educational facilities for children are also available. Apply Amarjit Singh, Assist Director-in-Charge, Central Electronics Engineering Research Institute, Pilani (Rajasthan) India.

ENGINEERS & TEACHERS

The University of Denver's College of Engineering and Research Institute are seeking electronic engineers and physicists with Ph.D. or D.Sc. degrees for research and teaching positions in the areas of electromagnetic wave propagation, antenna research, thin film and solid state electronics, circuit development, servomechanisms and infrared techniques. Also available are overseas field engineering assignments in Japan, Australia, South America and South Africa for periods of 1 to 2 years. Address inquiries to C. A. Hedberg, Head, Electronics Div., Denver Research Institute, University of Denver, Denver 10, Colo.

ELECTRONIC ENGINEERS AND PHYSICISTS

An expanding R&D laboratory has openings for qualified engineers and physicists to participate in and direct research and development programs in the microwave tube area. This is an opportunity for competent individuals who are interested in the direction of R&D programs at the leading laboratories in the country as well as participating in microwave in-house research. Salary from $7560 to $12,210, depending on experience and training. Send complete resume to Box 2038.

ENGINEERS

Expanding opportunities for research and graduate instruction at master's and doctor's level. Steady growth has created a need for highly qualified persons with a doctor's degree and a background of research and teaching in one of

(Continued on page 106A)

Honeywell Aero...
for the best of both

CLIMATE FOR CREATIVITY: Pictured here is the Polaris Missile for which Honeywell Aero is now producing inertial platforms. In addition, Aeronautical Division Engineers have created an Electrically Suspended Gyro for use on Polaris launching submarines. This unique gyro is potentially capable of providing accuracies never before achieved with an inertial navigation system. This project is typical of the advanced systems concepts and ideas which are being evolved and further developed into working hardware and products at Honeywell Aero.

Creativeness is not unique at Honeywell. It is widely encouraged. An engineer at the Aero Division finds himself exposed in almost all of his professional and non-professional contacts to individuals who understand, appreciate and respond to the creative individual.

Don't get us wrong—we are in business and have specifications and schedules to meet. We don't have unlimited freedom for each man to follow his every inclination. But within the framework of our business there is plenty of room for a man to exercise his imagination, drive and talents and in the process grow in professional stature and have his accomplishments recognized and rewarded.

CLIMATE FOR ENJOYMENT: No matter what sports interest you, either as a spectator or participant, you'll find them in the Minneapolis area. For example, this is the heart of America's finest fresh water fishing country. At the end of a busy day or week you can angle for pike, bass, trout, or scrappy panfish in the more than 80 lakes within 25 miles of Minneapolis. A journey northward will take you to the wilderness canoe country of the famed Superior National Forest. Here is one aspect of the recreational, educational, social and cultural living that you and your family enjoy when you work at Honeywell's Aeronautical Division in Minneapolis.

ENGINEERING OPPORTUNITIES: These are some of the areas in which 1200 Honeywell Aero engineers and scientists find a Climate for Creativity and satisfaction by working on a wide diversification of missile, space and advanced aircraft contracts and programs. There are openings for qualified personnel in each of these areas:

- **Satellites and Space Vehicles**
- **Missile Systems**
- **Marine Systems (ESG)**
- **Gyro Design**
- **Flight Control & Reference Systems**
- **Measurement & Display Instrumentation**
- **Ground Support Equipment**
- **Human Factors**
- **Applied Research (in any of the preceding areas)**
- **Evaluation Laboratory**
- **Components, Applications and Standards**
- **Material Engineering & Physical Chemistry**

Select your area and send your resume or request for information on specific openings to: Mr. James H. Burg, Technical Director, Aeronautical Division, 2610 Ridgway Road, Minneapolis 40, Minn.

Honeywell

H HONEYWELL Military Products Group

To explore professional opportunities in other Honeywell operations, coast to coast, send your application in confidence to: Mr. H. T. Eckstrom, Honeywell, Minneapolis 8, Minnesota.

Positions Open

(Continued from page 102A)

the following fields: mechanical dynamics; metallurgy; thermodynamics; applied physics; mathematics (statistics and operations research); electrical engineering (electronics and control). Contact Dean of the Hartford Graduate Center, Rensselaer Polytechnic Institute, East Windsor Hill, Conn.

ENGINEER OR PHYSICIST

Engineer or physicist to head Electronics Dept. of large bio-medical research institute in New York City. Knowledge of radiation physics detectors and circuits essential. Job involves design of new instruments and supervising maintenance. Send complete resume to Box 2040.

MICROWAVE SPECIALISTS

Microwave engineers needed for research and engineering in precision measurement. This work is in connection with the expanding Ordance Calibration Program for which Frankford Arsenal has world-wide responsibility. Engineers are needed to evaluate and apply newest microwave techniques to attenuation, power and impedance measurement both pulse and cw. Development of specialized equipments in a modern, fully equipped microwave lab. is offered. Salaries to $10,635. Federal Civil Service benefits and requirements. Apply Recruitment Office, Frankford Arsenal, Bridge and Tacony Sts., Phila. 37, Penna. Refer to ad by Mr. A. Bruno.

EE'S, PHYSICISTS

Expanding electronics concern has prestige position for qualified patent attorney to join its legal staff. LL.B. preferred but will consider minimum 2 years of law study. Ability to work closely with scientific and management people essential. Salary to $15,000. Northeast location. Please forward complete resume to Box 2041.

PROFESSOR OR ASSOCIATE PROFESSOR

Professor or Associate Professor with salary to $11,000 for session of two semesters. This fully accredited well-equipped E.E. Dept. needs one or two senior-level staff members with high qualifications. Ph.D. essential, to ensure full participation in graduate program and research activities. Prefer experience in electron devices, solid state or computer electronics, but other specialities considered. Medium size engineering college (700 undergraduate students) in brand new building. Exceptionally good retirement plan. Numerous sources of extra income. Pleasant living and working conditions in attractive uncrowded city. Address inquiries to Robert A. Chipman, Chairman, E.E. Dept., University of Toledo, Toledo 6, Ohio.

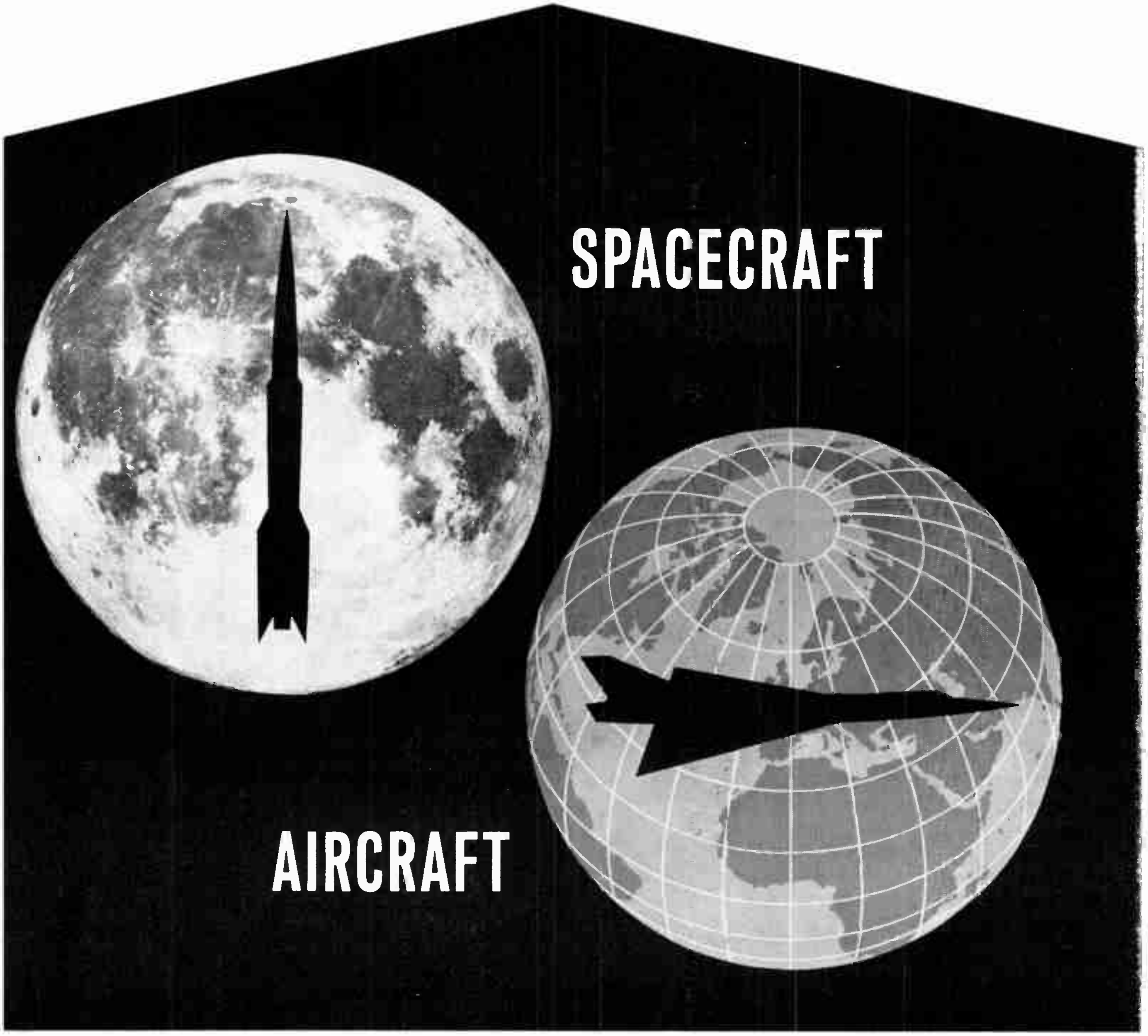
TEACHING AND RESEARCH

Opening for Electrical Engineer with Ph.D. to supervise graduate research and do some teaching. Research experience in microwave or semiconductors desirable. Private research or consultation encouraged. Excellent salary and rank arrangements. Department offers work leading to Ph.D. Position available Sept. 1961 or sooner. Apply Chairman, E.E. Dept., Vanderbilt University, Nashville 5, Tenn.

NOISE RESEARCH ENGINEERS

As a result of recent rapid expansion, Quan-Tech Laboratories—a leading manufacturer of equipment for the measurement and analysis of component noise—has openings for imaginative E.E.'s interested in this important new field. Address replies to Chief Engineer, Noise Research Div., Quan-Tech Labs. Inc., 60 Parsippany Blvd., P.O. Box 187, Boonton, New Jersey. Tel. Deerfield 4-8500.

(Continued on page 108A)



NOW UNDER ONE ROOF AT LOCKHEED / BURBANK

Lockheed California Division has expanded its already great resources at Burbank. For it now encompasses, under one roof, the two vast worlds of Spacecraft and Aircraft.

This typifies Lockheed policy to steadily advance the state of the art through basic research in related and scientific fields.

Result? Now—more than ever—Lockheed offers Scientists and Engineers uncommon opportunities; equips them with every modern facility; gives them freedom to explore and express new ideas; makes possible greater individual recognition.

Now being designed and developed in Spacecraft and Aircraft are: Hypersonic interceptors; V/STOL; supersonic interceptors; limocopters; missiles; manned spacecraft; and satellites.

Scientists and Engineers of outstanding talent are now invited to participate in this new, dual enterprise. Immediate openings are available to: Aerodynamics engineers; thermodynamics engineers; dynamicists; electric research engineers; servosystem engineers; electronic systems engineers; biophysicists; infrared physicists; hydrodynamicists; ocean systems scientists; physio-psychological research specialists; mammalian culture research specialists; and radiation hazards specialists.

Write today to Mr. E. W. Des Lauriers, Manager Professional Placement Staff, Dept. 1802, 2402 N. Hollywood Way, Burbank, California.

LOCKHEED

CALIFORNIA DIVISION

Positions Open

(Continued from page 106A)

STATISTICAL COMMUNICATION THEORIST

Senior Staff opportunity to initiate and direct advanced research programs in such areas as statistical communication and information theory; automatic pattern recognition and processing; adaptive, self-organizing learning systems; bionics; advanced computer theory. Contact J. R. Campbell, Chief Scientist, Chance Vought Electronics Div, P.O. Box 1500, Arlington, Texas.

OPERATIONS RESEARCH SPECIALIST

Establish and direct operations research activity. Analyses and recommendations must be provided for both (a) operational application of present products, and requirements for potential future products, and (b) intra- and inter-departmental operations. Contact J. R. Campbell, Chief Scientist, Chance Vought Electronics Div., P.O. Box 1500, Arlington, Texas.

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor with Ph.D. in Electrical Engineering or Physics. Interest in electronics, microwaves, solid state or circuit theory. Active research and graduate program with opportunities for outside consulting. Send resume to W. P. Smith, Chairman E.E. Dept., University of Kansas, Lawrence, Kansas.

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

ELECTRONICS TECHNICAL WRITER

Desires permanent position with a future. Age 30. 8 years experience in military (radar) and civilian electronics. 2 years experience as Technical Writer. Active security clearance. Member of IRE and STWP. Box 3006 W.

ENGINEERING MANAGER

Washington D.C. situation desired with military electronics firm engaged in R & D and production. Experienced in project and contract technical administration, customer service and liaison, market analysis, and operational evaluation planning. BSEE, plus USN special schools; age 41; married, 2 children. Will relocate from New York area. Resume upon request. Box 3012 W.

(Continued on page 110A)

Positions Wanted

By Armed Forces Veterans

(Continued from page 108A)

ENGINEER

Ph.D., summer 1961. 5 years college teaching experience. 1 year industrial experience. Interested in teaching and research progressive E.E. department. Prefer western U. S. Box 3013 W.

ENGINEER

Expects to receive MEE. degree from the University of Tokyo. January 1961. BEE. Cornell 1953. U. S. citizen. Experience in transistor circuits. electronic test equipment design and fabrication, technical writing, and technical translating. Military experience includes teaching and technical intelligence. Good knowledge of written and spoken Japanese. Desires position in Japan requiring professional competence. Box 3014 W.

MANUFACTURERS' REPRESENTATIVE

California experience. Now in New York area planning April 1961 return to West coast. Interested in interviewing companies anxious for better representation and more sales. If not near New York suggest we meet during IRE Show. but please write now. Box 3015 W.

BIOPHYSICIST—ENGINEER

Ph.D., MSEE., wishes faculty appointment, teaching and research. Publications and author of two books. Can develop biomedical instrumentation program. Equivalent industrial positions considered. Box 3020 W.

TEACHING

Naval officer, aged 38. B.S. in Engineering Electronics plus 35 graduate hours. Retiring in July 1961. 5 years teaching experience, both graduate and undergradaute. Desires teaching position at university in West or Southwest. Textbook author. Resume upon request. Box 3021 W.

R & D MANAGER

Desires assignment in industry or university. 20 years experience in industry, government and universities in R & D teaching and management encompassing broad fields of physics, electronics, earth sciences, education and administration. Box 3022 W.

SENIOR ENGINEER

MSEE. Age 30. Presently employed in practical network synthesis in time and frequency domains from audio to .5 Kmc. Desires position involving more network research challenges and responsibilities. Salary secondary. Will relocate. Box 3023 W.

PATENT ATTORNEY

BEE., B.Aero E., LLB; 4 years corporate and law office experience writing and prosecuting applications, interferences, infringement studies and licensing. 6 years diversified engineering experience including transistor circuitry and aircraft instruments. Interested in position in New York City area. Box 3024 W.

PUBLIC RELATIONS EXECUTIVE

Desires opportunity to help put medium or large company and its management on map with small to medium budget; strong with financial community, press, educators; electronics, wire service background. Now with a "top ten" company. Box 3025 W.

(Continued on page 112A)

Positions Wanted

By Armed Forces Veterans

(Continued from page 110A)

TECHNICAL WRITER.

Electronics Technician Chief would like to ghost or write under a dual byline with electronics engineers. If you have ideas which you feel fit the popular market but do not have the time to develop and write a practical do-it-yourself article we could probably work to our mutual advantage. Would also like technical writing assignments. Box 3026 W.

INTERNATIONAL OPERATIONS

Former Lieutenant-Colonel Marine Corps with strong civilian background for executive or liaison responsibilities, heavy experience South America and Europe. Technically trained to assist manufacturing, sales or field engineering management overseas. Box 3027 W.

Membership

(Continued from page 94A)

Webb, P. L., Vienna, Va.
Weber, R. G., Minneapolis, Minn.
Weimer, J. W., Bellevue, Wash.
Wise, B., Havertown, Pa.
Wright, P. W., Charlottesville, Va.
Young, R. T., Bala Cynwyd, Pa.

Admission to Member

Amsterdam, R. E., Wakefield, Mass.
Armstrong, J. L., London, Ont., Canada
Ashkin, A., Murray Hill, N. J.
Barci, H. C., Glenside, Pa.
Barr, D. D., Beaverton, Ore.
Beals, D. F., Webster, N. Y.
Behrendt, W. E., Verona, N. J.
Bental, L. J., Barner, Herts., England
Berger, H., Dorchester, Mass.
Bowen, H. D., New York, N. Y.
Bowen, J. L., Millington, Tenn.
Bower, K. D., Mountain View, Calif.
Braddock, J. V., El Paso, Tex.
Burnett, B., Flushing, N. Y.
Bruns, A. J., Hempstead, N. Y.
Bush, C. M., Dallas, Tex.
Buster, C. F., Jr., Arlington, Va.
Cage, J. H., Charlotte, N. C.
Camey, G. A., Flushing, N. Y.
Candy, C. J., Murray Hill, N. J.
Cannon, D. S., Sacramento, Calif.
Casterline, B., Des Plaines, Ill.
Cauble, R. A., Orlando, Fla.
Conn, R. L., Dallas, Tex.
Constance, M., Flushing, N. Y.
Crocker, W. S., Lynnfield Center, Mass.
Crosby, J. F., New York, N. Y.
Davidson, L. M., Flushing, N. Y.
Deboo, G. J., North Babylon, N. Y.
Dechter, A., Flushing, N. Y.
Dermatis, S. N., Youngwood, Pa.
Devot, A. J., Culver City, Calif.
Di Pietro, V. J., Bordentown, N. J.
Dooley, J. E., Seattle, Wash.
Edeline, W. E., Quito, Ecuador, S. A.
Emerson, F. P., Fabyan, Conn.
Farrell, R. J., Philadelphia, Pa.
Friedman, J., New York, N. Y.
Fulcher, R. W., Arlington, Va.
Furman, G. G., El Cerrito, Calif.
Gage, R. A., Oxnard, Calif.

(Continued on page 114A)

Membership

(*Continued from page 112A*)

Galle, R. C., Hawthorne, N. J.
Geanes, Z. K., Pacific Palisades, Calif.
Gioia, A. J., Warren, Pa.
Glasby, P. A., Saxonville, Mass.
Gold, J., Kingston, N. Y.
Goldman, R. F., Framingham, Mass.
Graham, C. E., Greensboro, N. C.
Greenwood, E., Flushing, N. Y.
Griffin, J. L., Cincinnati, Ohio
Harris, H. M., Jr., Webster, N. Y.
Harvey, J. C., Jr., Wellesley Hills, Mass.
Hassett, R. E., Farmingdale, L. I., N. Y.
Henry, I. G., Pasadena, Calif.
Hess, N. E., Burbank, Calif.
Hickox, G. R., Adelphi, Md.
Hinnant, H. O., Houston, Tex.
Holey, J. G., Atlanta, Ga.
Ihara, Y., Tokyo, Japan
Jensen, P. A., Baltimore, Md.
Johnson, J. L., Soddy, Tenn.
Johnson, G. W., Greensboro, N. C.
Joyner, W. T., Jr., Hampden-Sydney, Va.
Jurenko, J. A., Jr., Glenside, Pa.
Kaplan, A., St. Paul, Minn.
Kerrigan, P. R., Stamford, Conn.
Kishel, J. J., Bloomfield, N. J.
Klein, S., New York, N. Y.
Kohus, J. P., Los Angeles, Calif.
Kramer, A. T., Bedford, Mass.
Kries, J. T., Chicago, Ill.
Krstansky, J. J., Chicago, Ill.
Lahti, A. W., Cambridge, Mass.
Lana, W. M., Culver City, Calif.
Lee, R. A., Natick, Mass.
Leo, E., Stamford, Conn.

(*Continued on page 116A*)

Membership

(Continued from page 114A)

Levine, J. H., Englewood, N. J.
Lewis, F. R., Spenard, Alaska
Lipp, A. E., Federal Way, Wash.
Lundstrom, J. W., Santa Barbara, Calif.
Marxheimer, R. B., Daly City, Calif.
McDonnell, P. J., Toronto, Ont., Canada
Mellen, W. R., Chelmsford, Mass.
Michaels, J. V., Neptune, N. J.
Michaelson, A., Schenectady, N. Y.
Miller, B., Los Angeles, Calif.
Miller, B. J., Columbus, Ohio
Moore, G. E., Owensboro, Ky.
Moulton, W. S., Dayton, Ohio
Nagelberg, E. R., New York, N. Y.
Nelson, R. E., Duarte, Calif.
Newman, I. M., New York, N. Y.
Niemi, A. N., Fort Huachuca, Ariz.
Noll, G. T., Poughkeepsie, N. Y.
Nooteboom, P., Manhattan Beach, Calif.
Noyes, J. G., Baughurst, Nr. Basingstoke, Hampshire, England
Oppenheim, J. B., Bangor, Me.
Paharia, R. K., Oak Park, Ill.
Palmer, M. L., Jr., Houston, Tex.
Pascola, A. N., Springfield, Va.
Pattenaude, H. E., Downey, Calif.
Pegram, J. B., New York, N. Y.
Pennington, A. J., Ann Arbor, Mich.
Pfeiffer, W. A., Silver Spring, Md.
Planteijdt, F. J., Baarn, Holland
Quill, J. S., Schenectady, N. Y.
Reese, O. T., Houston, Tex.
Renton, C. A., New York, N. Y.
Resker, R. C., Boston, Mass.
Reynolds, A. T., Jr., Baltimore, Md.
Roloff, H. A., Cocoa, Fla.
Rubin, F., Brooklyn, N. Y.
Rupley, D. C., East Syracuse, N. Y.
Rush, R. B., Granada Hills, Calif.
Salt, S. D., Midland, Tex.
Sarles, M. D., Baltimore, Md.
Sauve, W. A., Montreal, Que., Canada
Schuette, K. R., Los Angeles, Calif.
Schuytemaker, J., Amsterdam, Netherlands
Serra, G. F., Caldwell, N. J.
Simmons, R. H., Huntsville, Ala.
Smith, A. P., Costa Mesa, Calif.
Smith, W. F., Jr., Milwaukee, Wis.
Stebbins, J. T., San Diego, Calif.
Steere, E. A., Albany, N. Y.
Straight, P. E., Pittsburgh, Pa.
Thoms, E. R., Dayton, Ohio
Timlin, T. N., Norristown, Pa.
Traub, J. L., Whitewater, Wis.
Vacha, F., Babson Park, Mass.
Valin, R. E., Montreal, Que., Canada
Van Run, J. L. A., Amsterdam, Netherlands
Van Shura, C. F., Washington, D. C.
Vendelin, G. D., San Jose, Calif.
Vermeulen, J. C., Amsterdam, Netherlands
Wallace, P. W., Redwood City, Calif.
Webster, J. G., Melbourne, Fla.
Weibel, D. O., Palo Alto, Calif.
Wen, W. L., San Diego, Calif.
Wheeler, E. G., Culver City, Calif.
White, J. R., Wilmington, Del.
Whitescarver, C. D., Las Cruces, N. Mex.
Williams, O. L., Jr., Wright-Patterson AFB, Ohio
Wyman, N., New York, N. Y.

Admission to Associate

Abbot, H. J., Seaford, N. Y.
Abbott, F. J., Clinton, Ont., Canada
Allard, H. K., Jr., Cambridge, Mass.
Alvarez, E. C., Woodland Hills, Calif.
Arlt, J. E., Flushing, N. Y.
Aubert, J. L., Framingham, Mass.
Auer, L., Hicksville, N. Y.

(Continued on page 118A)

Membership

(*Continued from page 116A*)

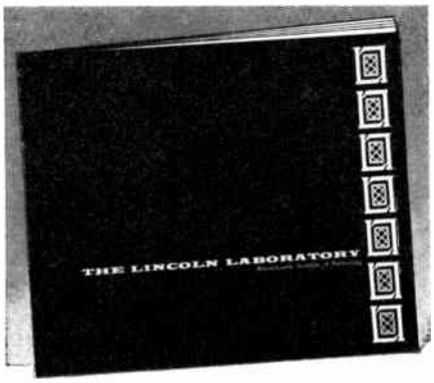
Barry, G., Fresh Meadows, N. Y.
Belice, M. M., Philadelphia, Pa.
Birsten, B. R., Douglaston, L. I., N. Y.
Blasman, J., Los Angeles, Calif.
Bodak, L., Flushing, N. Y.
Bos, R. A., La Grange, Ill.
Braden, R. L., Los Angeles, Calif.
Brener, V., Sao Paulo, Brazil
Broaddus, A. G., Jr., Richmond, Va.
Cabán, J. P., New York, N. Y.
Chalekian, J., Montreal, Que., Canada
Chandler, A., Jamaica, N. Y.
Chen, W. W., Brighton, Mass.
Cline, J. D., Greensboro, N. C.
Collier, C., Fairborn, Ohio
Conley, S. D., Jr., Ipswich, Mass.
Connolly, J., Flushing, N. Y.
Corcoran, B. W., Lynnfield, Mass.
Cortelyou, R. G., Staten Island, N. Y.
Coulson, E. F., Malvern, Pa.
Cox, R. W., Denver, Colo.
Crangle, W. J., Flushing, N. Y.
Cunningham, J. A., Batavia, Ill.
Cusimano, J. P., Rome, N. Y.
Davis, D. B., Evanston, Ill.
Davis, R. D., Neptune, N. J.
Davison, C. A., Asbury Park, N. J.
Dolkas, J. B., Brooklyn, N. Y.
Dowd, J. J., Chicago, Ill.
Doyling, V. G., Clinton, Ont., Canada
Drevet, C., Montreal, Que., Canada
Dub, L. M., Washington, D. C.
Edelberg, S., New York, N. Y.
Ekola, E. C., El Paso, Tex.
Eliason, R. D., Plainview, N. Y.
Elustondo, A. E., Brooklyn, N. Y.
Eskedal, T., La Jolla, Calif.
Eubanks, F. G., Pasadena, Calif.
Farrell, T. C., Dayton, Ohio
Finkelstein, P. N., Brooklyn, N. Y.
Forant, S. A., Culver City, Calif.
Gabbert, V. N., Los Angeles, Calif.
Golubinski, A. P., Brooklyn, N. Y.
Gommellini, G., Rome, Italy
Goodman, R. J., South Euclid, Ohio
Goodwin, H. R., San Jose, Calif.
Gould, R. D., Flushing, N. Y.
Greenberg, S. F., Utica, N. Y.
Groehnert, W. A., Flushing, N. Y.
Hall, S. L., Culver City, Calif.
Halls, E. J., White Bear Lake, Minn.
Harris, R. L., Centredale, R. I.
Hartman, C. T., Flushing, N. Y.
Hawkes, R. M., Ossining, N. Y.
Heilman, P. M., Needham, Mass.
Heldman, J. H., Don Mills, Ont., Canada
Huffsmith, R. R., Denver, Colo.
Jagodzinski, N. S., Buffalo, N. Y.
Johnson, A. J., Allston, Mass.
Jones, M. M., Scotch Plains, N. J.
Jurkovich, G. M., Gardena, Calif.
Katz, D., Flushing, N. Y.
Kawalek, A. M., Baltimore, Md.
Kearney, J. E., New York, N. Y.
Kelley, A. S., Roosevelt Field, N. Y.
Kinney, K. K., Livermore, Calif.
Kristianson, L., Flushing, N. Y.
Lametta, J., Flushing, N. Y.
Laramee, A., Hawthorne, Calif.
Larson, H. E., Plainfield, N. J.
Le Claire, W. L., Middle Village, N. Y.
Leland, R. M., Wellesley Hills, Mass.
Ligotti, R. E., Flushing, N. Y.
Ludtke, J. U., Laval des Rapides, Que., Canada
Marchese, R., Silver Spring, Md.
Marchi, T. J., Oceanside, N. Y.
Martin, R. O., Sudbury, Ont., Canada
Maurer, J. G., Flushing, N. Y.
McFadden, R. M., Long Beach, Calif.

(*Continued on page 120A*)

Membership

(*Continued from page 118A*)

McKinney, F. L., Daphne, Ala.
Merrit, A. H., III, Crestwood, N. Y.
Metzger, J. M., Flushing, N. Y.
Mezzio, G. R., Shreveport, La.
Miller, J. W., San Francisco, Calif.
Miller, J. S., McKeesport, Pa.
Morgan, J. F., Point Lookout, N. Y.
Morgan, M. M., Boston, Mass.
Muzyczyszyn, O., Whitestone, N. Y.
Myers, J. E., Pawtucket, R. I.
Nardo, F. K., East Boston, Mass.
Needham, G. A., Collingswood, N. J.
Nicholas, E. R., Thomasville, Ala.
Nolan, J. F., Royal Oak, Mich.
Noroian, E. H., Baltimore, Md.
Papierski, F., Worcester, Mass.
Payton, T. J., Flushing, N. Y.
Perazzoli, E. H., Philadelphia, Pa.
Perez, J. M., Buenos Aires, Argentina
Perryon, P., St. Laurent, Montreal, Que., Canada
Petros, J. D., Turlock, Calif.
Pfarrer, J., Claifon, N. J.
Poinsot, W. B., Sherman Oaks, Calif.
Polovneff, G. J., San Francisco, Calif.
Porter, D. K., Pomona, Calif.
Post, M. D., Ft. Lauderdale, Fla.
Powers, K. T., Waltham, Mass.
Prokop, A. A., Dorchester, Mass.
Quigley, L. E., Downey, Calif.
Radansky, H., New York, N. Y.
Rader, S., New York, N. Y.
Rasmussen, F. M., Berkeley, Calif.
Reniker, H. T., San Diego, Calif.
Reynolds, J., Bogota, Colombia, S. A.
Rieman, W. E., Arlington, N. J.
Riley, J. F., Bradford, Pa.
Roble, C. W., Dahlgren, Va.
Robarge, R. G., Hamden, Conn.
Roman, S., Flushing, N. Y.
Rosenstein, M., Bergenfield, N. J.
Roth, J. G., Portland, Ore.
Roulston, A. M., Winnipeg, Manit., Canada
Sadofsky, M. B., Brooklyn, N. Y.
St. George, R. C., Manchester, N. H.
Shellkopf, R. E., Baltimore, Md.
Sherwood, C. S., St. Albans, N. Y.
Shingles, R. D., Kitchener, Ont., Canada
Stanfield, J. W., Dallas, Tex.
Sternberg, H. S., Yonkers, N. Y.
Stuart, H. P., Passaic, N. J.
Teeple, W. H., Los Angeles, Calif.
Tippett, W. D., High Point, N. C.
Thom, R. H., Flushing, N. Y.
Tolston, R. B., Baltimore, Md.
Van Velzer, V. C., Palo Alto, Calif.
Victorine, G. B., San Carlos, Calif.
Vilfroy, R. H., El Paso, Tex.
Walls, D. L., Chicago, Ill.
Webster, R. E., Ottawa, Ont., Canada
Weiner, I. L., New York, N. Y.
Weinstein, R. L., Far Rockaway, N. Y.
Whiting, D. F., Jr., Silver Spring, Md.
Willemsen, W. J., Poughkeepsie, N. Y.
Williams, D. D., Anaheim, Calif.
Wilson, L. M., Toronto, Ont., Canada
Wright, C. M., Falls Church, Va.
Zapin, A. J., Brooklyn, N. Y.
Zetterberg, T. W., Cambridge, Mass.
Zimmerman, J. D., Degraff, Ohio



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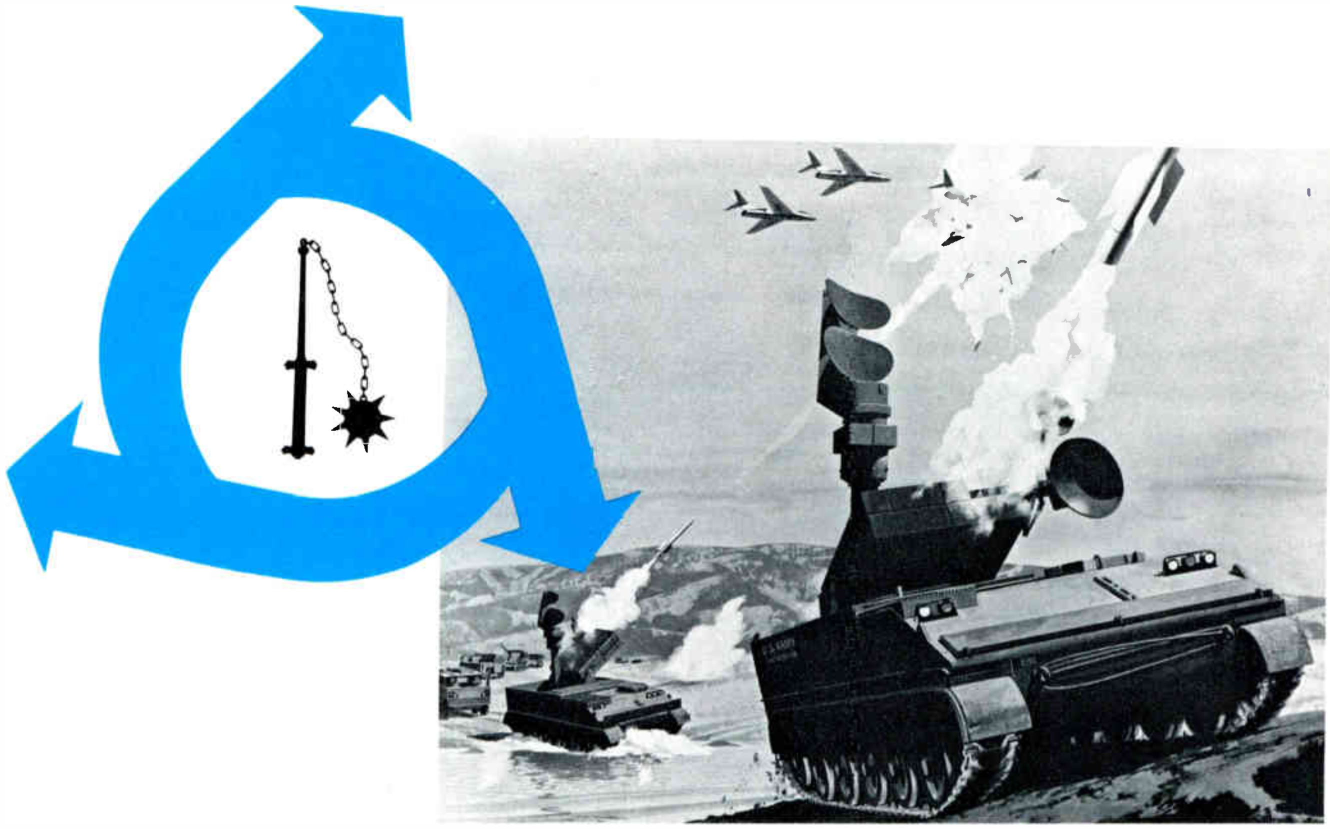
ME	☐	☐	☐	___	Other: ___	☐	☐	☐	___
Mathematics	☐	☐	☐	___	___	☐	☐	☐	___

School (Highest Degree) ______



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CONVAIR/POMONA CONVAIR DIVISION OF **GENERAL DYNAMICS**

Section Meetings

AKRON

"A Three-Dimensional Automatic Pattern Analyzer," Hugo Fulmer, Scientific Atlanta, Inc. 11/15/60.

ANCHORAGE

"New Developments in Computer Technology," R. B. Kier, IBM Corp. 11/7/70. Election of Officers. 12/5/60.

ATLANTA

"The General Electric Thin Strip Microminiaturization Program," J. P. Dietz, GE Co. 12/16/60.

BALTIMORE

"Space Exploration—Status & Implication," Leo Steg, GE Co. 11/14/60.

BAY OF QUINTE

"How to Engineer Breakthroughs," Trevor Clark, Westinghouse Elec. Corp. 11/16/60.

BEAUMONT-PORT ARTHUR

"Technical Talk on Power Plant Operation," Jim Derr, Gulf States Utilities. 11/1/60.

"Magnetic Tape," Edward Schmidt, Reeves Soundcraft Corp. 11/15/60.

"Communications Systems Using Integrated VHF & Microwave for Complete Coverage," Virgil Akerman, Motorola. 12/13/60.

BINGHAMTON

"Impressions of Automatic Control in Russia," Hal Chestnut, GE Co. 10/13/60.

BUFFALO-NIAGARA

"Scientific Shock in Military Electronics," L. S. Sheingold, Sylvania Advanced Res. Labs. 11/16/60.

CHINA LAKE

"Radiation Effects on Semiconductors," H. L. Taylor, Texas Instruments, Inc. 11/14/60.

CINCINNATI

"Microminiaturization & Molecular Electronics," R. D. Alberts, Wright Air Dev. Div. 11/15/60.

CLEVELAND

"Trends in Engineering Education," C. B. McEachron Jr., Case Inst. of Tech. 11/10/60.

COLUMBUS

"Thermionic Converters," G. M. Branch, GE Res. Lab. 11/10/60.

"Fundamental Design Considerations for Linear Transistor Amplifiers," D. E. Thomas, Bell Tele. Lab. 12/6/60.

DALLAS

"Impressions of Soviet Semiconductor Research," R. L. Petritz, Texas Instruments, Inc. 12/6/60.

DETROIT

"What We have Learned at Fermi," R. W. Hartwell, Power Reactor Dev. Corp. 11/15/60.

"Tunnel Diodes, Operation & Application," I. A. Lesk, GE Co. 12/9/60.

ELMIRA-CORNING

"Ultrasonic Cleaning Techniques," E. A. Walter, Narda Ultrasonics Corp. 9/19/60.

"The Outgassing of Tube Materials," John Turnbull, RCA. 10/24/60.

"Magnetohydrodynamics," Werner Emmerich, Westinghouse Res. Labs. 11/14/60.

"The Esaki Diode," Leo Esaki, IBM Corp. 12/5/60.

(Continued on page 124A)

Section Meetings

(Continued from page 123A)

EL PASO

"New Trends in Data Processing," Frank Lent, IBM. 11/17/60.

EMPORIUM

"Electrical Space Propulsion Systems," Mr. Szymanowski, Curtiss-Wright. 6/21/60.

"Telephony's Technical Progress," F. D. Reese, Automatic Electronics Labs., "A Linear Sweep System for Television," George Kent, Sylvania Home Electronics, "Reliability—Entertainment Electronic Eq.," Richard O'Fallon & J. R. Belville, Motorola Inc. 8/26/60.

"The TIMMS Project," A. P. Haase, GE Co.; "Emission From Semiconductor Materials," Paul Carroll, Carborundum Co.; Seminar Picnic. 8/27/60.

"Use of Ultrasonics In Industry," Robert McMillen, Electro-Tel Corp. 9/20/60.

"Nuclear Time Dating Techniques," David Bandel, American Machine & Foundry Co. 10/18/60.

"New Design of Manufacturing Eq.," S. J. Gartner, Sylvania Electron Tubes. 11/15/60.

EVANSVILLE-OWENSBORO

"Television Stereophonic System," R. B. Dome, GE Co. 11/14/60.

FORT HUACHUCA

"Propagation of VHF Signals by Scattering from Meteor Trials & Ionization Aligned Along the Earth's Magnetic Field," J. L. Heritage, Smyth Res. Assocs. 11/18/60.

GAINESVILLE

"IRE Opportunities for Members," Dr. Ronald L. McFarlan, IRE President. 11/8/60.

HOUSTON

"A New Low Cost Approach to Closed Circuit TV," Harry Keep, Gulf Coast Electronics. 11/15/60.

HUNTSVILLE

"A Survey of Some Recent Developments in Reliable FM Communication," F. J. Bagadady, Mass. Inst. of Tech. 9/29/60.

"The Solid State, A Technological Revolution," R. A. Ramey, Westinghouse. 10/27/60.

Panel Discussion—"Seimiconductor Growth vs Thin Film Deposition," J. R. Black, Motorola, H. H. Henkels, Westinghouse, Roy Currie, Marshall Flight Center, C. R. Seashore, Redstone Arsenal. 11/17/60.

INDIANAPOLIS

"The Thermionic Diode as an Energy Converter," Mr. D. Dresser, GM. 10/27/60.

Tour of Avon Big 4 Railroad Yard. 12/1/60.

ISRAEL

"The International Symposium on Information Theory 1960," Dr. J. Shekel, Ministry of Defence. 10/19/60.

"Digital Differential Analyzers," Alexander Fuchs, Ministry of Defence. 11/9/60.

ITHACA

"Microwaves—A Discovery Ignored by Two Generations," Sven Olving, Cornell Univ. 11/3/60.

KANSAS CITY

MAECON—(Mid-America Electronics Conference) 11/15-16/60.

(Continued on page 126A)

Section Meetings

(Continued from page 124A)

KITCHENER-WATERLOO

"Soviet Research in Circuit Theory & Automatic Control," M. E. Van Valkenberg, Univ. of Illinois. 11/21/60.

LAS VEGAS

"Peaceful Uses of Atomic Energy," Lewis Fussell, Edgerton, Germeshausen & Grier, Inc. 12/7/60.

LITTLE ROCK

"Low Frequency Atmospherics," J. W. Green, Jr., Univ. of Ark 11/21/60

LONG ISLAND

"Symposium on Adaptive Control Systems." 10/7-18-19/60.

"Electronic & Mathematical Approaches to Living Cells," Britton Chance. Univ. of Pa. School of Medicine. 11/1/60.

LUBBOCK

"The Role of the Jr. College in Technical Education," R. S. Burks, South Plains College. 11/15/60.

MOBILE

"Communications Engineering for BMEWS," A. W. Shaefer, Western Elec. Co.; "Aims & Accomplishments of NW Florida Section IRE," Kenneth Huntley, USAF; Election of Officers. 10/28/60.

NORTH CAROLINA

"IRE Activities—Present & Future," Dr. R. L. McFarlan, IRE President. 11/29/60.

(Continued on page 128A)

Section Meetings

(*Continued from page 126A*)

NORTHERN ALBERTA

Film—"The Saint Lawrence Seaway." 11/15/60.

NORTHERN NEW JERSEY

"Tiros Weather Satellite," Maury Staton, RCA. 11/9/60.

NORTHWEST FLORIDA

"History & Application of Semiconductors," Mr. Linhart, Motorola, Inc. 11/1/60.

OMAHA-LINCOLN

Installation of Officers; Section Business Meeting. 12/2/60.

QUEBEC

"Discussion of Some Basic Problems When Planning Point-to-Point Radio Systems," K. H. Zement, Quebec Tele. Co. 11/21/60.

ROCHESTER

"Education: The Ultimate Weapon," G. S. Brown, Mass. Inst. of Tech. 11/3/60.

ROME-UTICA

"Optical Masers," R. C. Mack, Hughes Aircraft Co. 11/15/60.

SALT LAKE CITY

"The Rocket Static Test Program at Bacchus," Joseph Arbogast, Hercules Powder Co. 10/15/60.

"Semiconductor Developments & Devices," G. S. Horsley. 11/10/60.

SAN FRANCISCO

"Experiences of Recent Visitors to the Soviet Union," David Packard, Hewlett-Packard Co., Panel: Irmgard Flugge-Lotz, Stanford Univ., G. L. Pearson, Stanford Univ., Bernard Widrow, Stanford Univ., Roy Amara, Stanford Res. Inst., Fred Kurzweil, Jr., IBM. 10/24/60.

SHREVEPORT

"The IRE Member and His Organization," J. N. Dyer, Vice-President IRE. 10/25/60.

"Thermoelectricity," A. R. Orsinger, Texas Instruments. 12/6/60.

SOUTH CAROLINA

"The Engineer and His Professional Society," Dr. R. L. McFarlan, IRE President. 1/11/60.

TORONTO

"Acoustics & the Ear," W. E. Hodges, Univ. of Toronto, Consultant. 11/14/60.

Tour of Studios & Facilities, K. J. Easton, Trans-Canada Telemeter. 12/5/60.

TWIN CITIES

"Report on First Congress of the International Federation of Automatic Control—Moscow," Hugo Schuck, Minneapolis-Honeywell Regulator Co. 11/17/60.

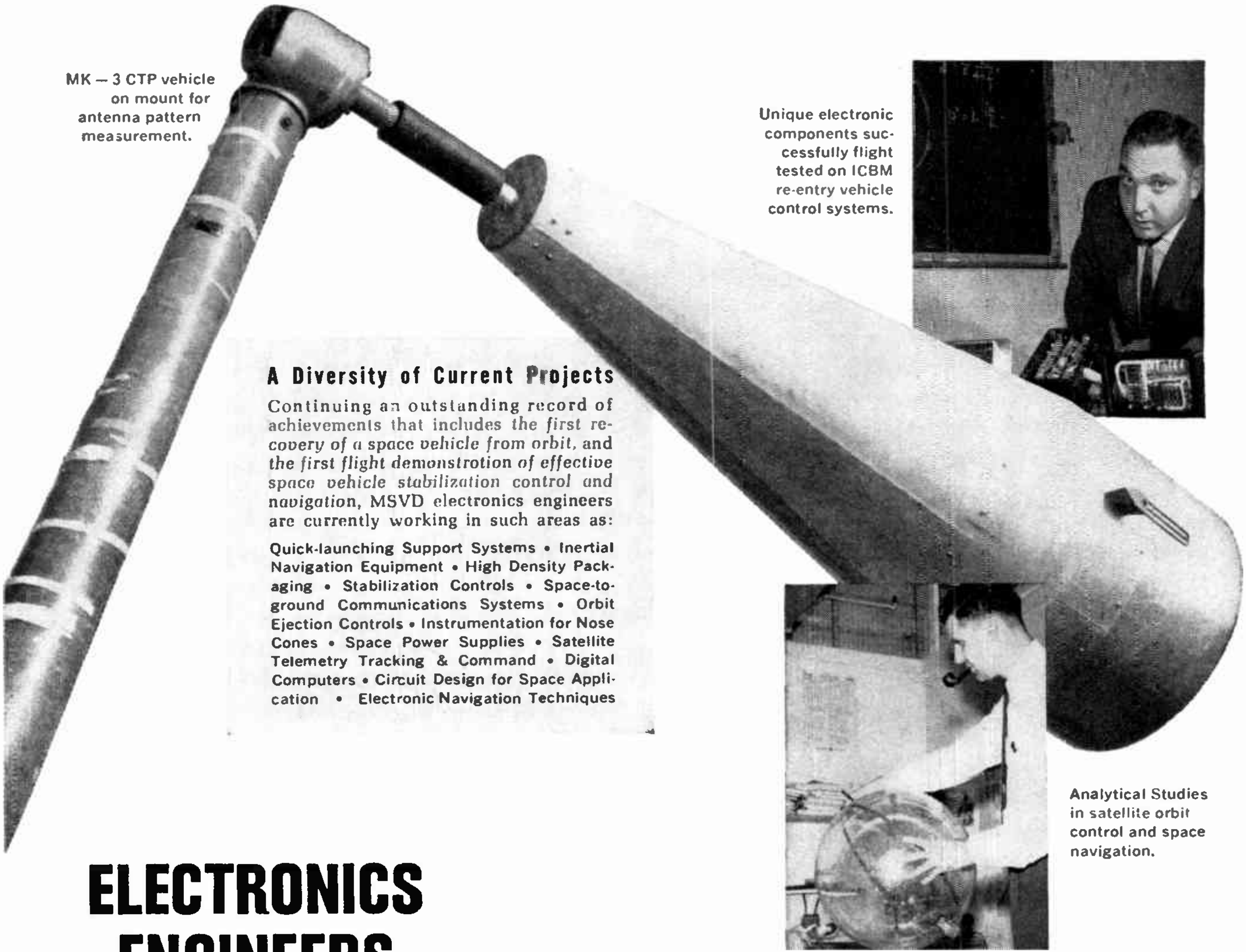
VANCOUVER

Panel Discussion—"Electronics in Education," L. R. Kersey, University of B. C., Panel: A. D. Moore, Univ. of B. C., R. Ridsdale, B. C. Vocational School, K. Wheeler, Vancouver Vocational Inst. 10/17/60.

"Taming 50 Kilowatts," C. R. Smith, C.K.W.X. 11/21/60.

(*Continued on page 130A*)

MK — 3 CTP vehicle on mount for antenna pattern measurement.



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Analytical Studies in satellite orbit control and space navigation.

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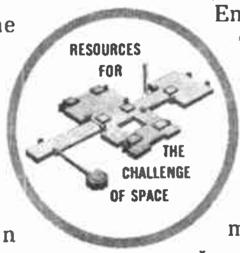

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GENERAL ELECTRIC

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Section Meetings

(Continued from page 128A)

VIRGINIA

"Coaxial or Strip Line Semi-Conductor Switch for Microwave," R. V. Garber, Diamond Ordnance Fuze Lab. 11/4/60.

"U. S. Navy Space Surveillance Program," M. S. Maxwell, Naval Proving Grounds. 11/11/60.

"An Airborne Platform Powered by Microwave Fuel," Dr. R. L. McFarlan, IRE President. 11/30/60.

WASHINGTON

"Project Mercury," W. J. North, NASA. 11/14/60.

WESTERN MASSACHUSETTS

"Micro Electronic Semiconductor Networks," C. H. Phipps. Texas Instruments. 11/9/60.

SUBSECTIONS

BURLINGTON

"Electronics in Japan," Shigeyuki Nagasawa, Tama Elec., Tokyo, Japan. 11/15/60.

EASTERN NORTH CAROLINA

"Application of Nuclear Techniques to Industry," R. L. Ely, H. G. Richter, Res. Triangle Inst. 11/11/60.

"Carrier Telephone Systems," G. E. Gatliff, ITT Kellogg. 12/9/60.

LANCASTER

"The Role of Electronics in the Highways & Autos of the Future," George Gray, RCA Res. Labs. 10/25/60.

LEHIGH VALLEY

"Determining the Vulnerability of Military Equipment to Electronic Countermeasures," R. H. Sugarman, U. S. Army Signal Res. & Dev. Lab. 11/2/60.

MERRIMACK VALLEY

"Applications of Computers," C. L. Gold, IBM. 11/14/60.

"Unity of the Engineering Profession," Panel: John Hitt, Howard Evirs, I. Middleton, Frederick Bacon 12/5/60.

NEW HAMPSHIRE

"Space Surveillance," E. W. Wahl, A. F. Cambridge Res. Center. 11/16/60.

NORTHERN VERMONT

"Automation Engineering, Rules & Principles," L. M. Bundy, IBM. 11/21/60.

PANAMA CITY

"Investment & Management of Securities," John Espy, Espy & Smith Investments Securities Inc. 11/15/60.

PIKES PEAK

"INCONORAMA Data Transmission & Display System," Ben Fischer, Fenske, Fedrick & Miller, Inc. 11/9/60.

SANTA ANA

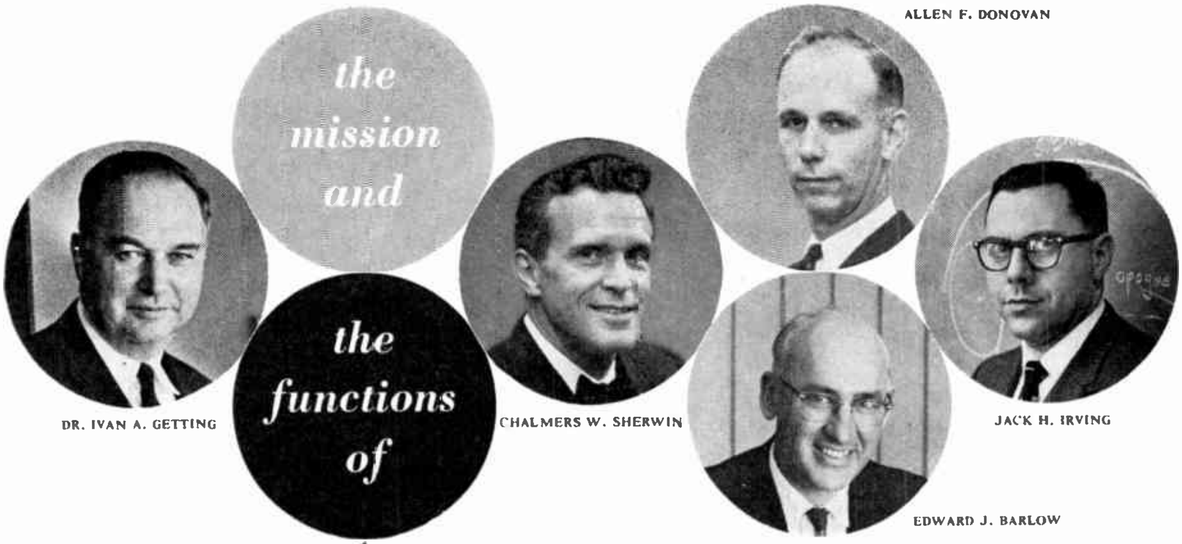
"Industrial Design & Human Engineering in Man-Machine Systems," Henry Dreyfuss, Henry Dreyfuss Co. 11/15/60.

SANTA BARBARA

"Observations on Education Abroad," William Everitt, Univ. of Illinois; "Operation Greenhouse," Roderick Morrison, Edgerton, Germeshausen & Grier; Tour of Hoffman Science Center. 11/17/60.

WESTCHESTER

"Recent Developments in the Theory of Error Correcting Codes," R. T. Chien, IBM. 11/16/60.



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Aerospace Corporation scientists and engineers are already engaged in a wide variety of specific systems projects and forward research programs, under the leadership of scientist/administrators including corporation president Dr. Ivan A. Getting, senior vice president Allen F. Donovan, and vice presidents Edward J. Barlow, William W. Drake, Jr., Jack H. Irving, and Chalmers W. Sherwin.

Aerospace Corporation is currently seeking scientists and engineers capable of meeting genuine challenge and with proven ability as:

- SENIOR ELECTRONICS ENGINEERS
  - Communications systems
  - Guidance electronics
  - Data processing systems
  - Radio techniques
  - Electromechanical design
  - Information theory
  - Sensing systems
- SPACE VEHICLE SPECALISTS:
  - Senior power systems engineer
  - Sr. flight performance analyst
  - Re-entry aerodynamicist

Those qualified and experienced in these and related fields are urged to direct their resumes to:

Mr. James M. Benning, Room 109
P.O. Box 95081, Los Angeles 45, Calif.

AEROSPACE CORPORATION

A new and vital force engaged in accelerating the advancement of space science and technology

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

Two-Channel Oscillograph

The mobile Model 297 two-channel oscillograph is now available from **Sanborn Co.** Industrial Div., 175 Wyman St., Waltham 54, Mass. This complete system, which uses plug-in interchangeable "850" style preamplifiers, is housed in a 35-inch-high wheel-mounted grey steel cabinet that permits easy movement between rooms or into and out of elevators. Wide application versatility is provided by the preamplifiers, available in carrier, dc coupling, phase sensitive demodulator, and low level types.

The recording mechanism, with low impedance, enclosed galvanometers, uses the heated stylus method to produce traces on 50 mm rectangular coordinate charts moving at a choice of four pushbutton-selected speeds. The system includes two transistorized current-feedback power amplifiers, their power supply, and the preamplifier power supply. System cooling is provided by forced filtered air from an integral blower.

Major specifications of the Model 297 include frequency response to 125 cps, within 3 db at 10 mm peak-to-peak amplitude; gain stability better than 1% over 20°C temperature changes and 20 volt power line variations; maximum nonlinearity 0.25 mm. Mechanical features include chart speeds of 1, 5, 20 and 100 mm/sec; individual stylus heat controls; marginal timer-marker stylus with internal 1 second timer; and a paper take-up assembly.

An optional MOPA is available to provide carrier and chopper excitation required by certain preamplifiers. The 297 System can also be supplied in a portable carrying case or for standard 19" rack mounting. Complete information is available on request from the firm

Noise Generator

A new reasonably priced random-noise generator has been announced by **Gonset Div., Young Spring & Wire Corp.,** 801 S. Main St., Burbank, Calif. Intended for operation in the VHF range this new

(Continued on page 134A)

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 132A)

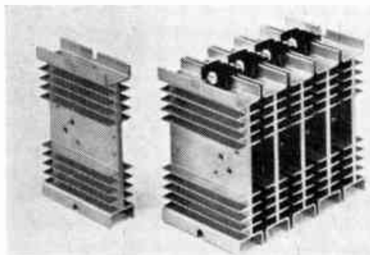
generator provides direct readings of sufficient accuracy for average laboratory use. It is, however, well suited for production line testing.

The noise-generating head is a separate assembly which connects to the main housing by a flexible cable. This arrangement allows the head to be attached directly to the receiver under test; this eliminates stray signal pickup from long connecting leads. The type 5722 noise diode and associated circuitry are located within this external head which fits into a clamp on the side of the housing when not in use.

The instrument provides direct reading to 25 db into 50 ohm impedances. The large-scale panel meter facilitates accurate reading of low noise figures. Filament of the noise generating diode and random-noise output is controlled by a variable autotransformer. Rapid checking of residual starting noise is expedited by a panel switch which temporarily cuts off diode current without need to change filament voltage settings.

Self-contained power supply operates from 117 volts ac, uses diode rectifiers and provides regulated dc operating voltage.

Transistor Heat Sinks



Heat sinks for power transistors and diodes, designed for ease of mounting and stacking, are now available for immediate delivery from **Invar Electronics Corp.,** 323 W. Washington Blvd., Pasadena, Calif. Offering 25 square inches of surface area for each inch of length, these heat sinks provide a large heat dissipating surface while maintaining overall size. The units were specifically designed to be used singly or in multiple heat sink arrays and are

(Continued on page 136A)

NEWS New Products

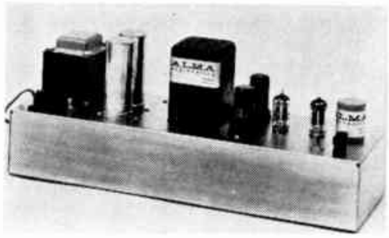
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 134A)

furnished with holes for mounting the power transistor, for mounting or "stacking" the heat sinks in a compact array, and for attaching a terminal or resistor board to the heat sink. The finish is electrically conductive and meets MIL C-5541. Special hole patterns and finishes can be furnished. For all specifications write to Sales Dept. P. at the firm.

Line Amplifier

Alma Engineering, 901 W. Washington St., San Diego 3, Calif., has just introduced Model 200 line amplifier, offering impressive response and distortion characteristics. This unit has frequency response within ± 0.5 db from 20 to 20,000 cps. Total harmonic distortion does not exceed 0.080% at any test frequency from 20 to 20,000 cps, measured at +28 dbm output. Hum and noise are 70 db below +8 dbm.

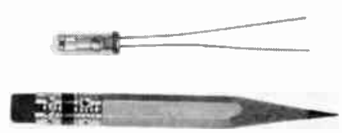


Designed primarily for the FM broadcasting and recording industries, Model 200 exceeds all NAB specifications. The amplifier uses silicon diode rectifiers, Mylar capacitors and features modular construction.

For further information write to the firm.

Cadmium Cells

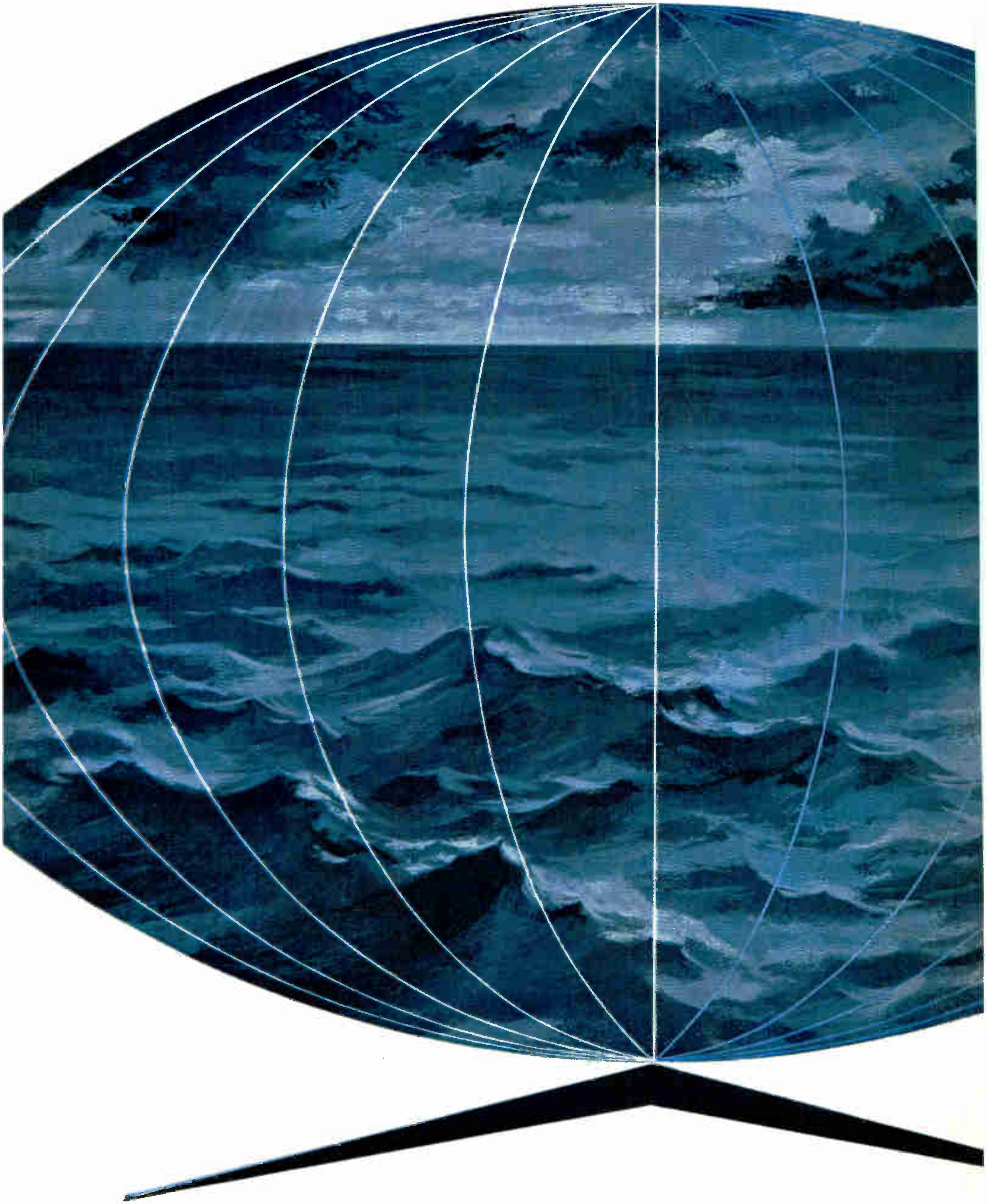
Amperex Electronic Corp., 230 Duffy Ave., Hicksville, L. I., N. Y., has announced a diversified line of photoconductive cadmium sulphide cells, including microminiature types.



One of the items in this line is the ORP 60, a microminiature cell mounted in a sealed, all-glass transistor envelope measuring 0.59 inch in length and 0.25 inch in diameter. It is top sensitive and has a 70 mw dissipation rating.

The ORP 60 is suited for use in comput-

(Continued on page 140A)



(Continued from page 136A)

ers where it functions as a noiseless resistor. It is also used to advantage in light-operated flip-flop circuits.

The ORP 60 has also been utilized for the automatic control of contrast and brightness in TV receivers, whereby the intensity of the picture beam varies with the ambient illumination. Another application is automatic street light control, switching lights on and off when the daylight rises or falls below a certain level.

Priced at $0.35 to $0.75 depending on ordering quantity, the ORP 60 is immediately available in mass production lots.

The type ORP 61 is the side sensitive version of the ORP 60.

Cadmium sulphide cells are being used in place of gas-filled phototubes, particularly in on-off applications. With a sensitivity 10,000 times greater than conventional phototubes, as well as high dissipation ratings, they replace the phototube, and its amplifier.

Variable Attenuator

It is now possible to obtain greater linearity over a wider band width in the L-Band range than heretofore with the development of the Model 1001 waveguide beyond cut-off type attenuator by **General Communication Co.**, 677 Beacon St., Boston 15, Mass.

(Continued on page 112A)

Why don't you **Talk** with Westinghouse about Advanced Research and Development

For 75 years, Westinghouse has pioneered new ideas . . . ideas reinforced by analysis and development. At Westinghouse-Baltimore, engineers in the Advanced Development subdivision engage in studies of such projects as: electronic detection of submerged submarines; wall-size electroluminescent military situation displays; satellite and space communication systems; ICBM detection and interception systems; long-range radars with high definition; and solid state feasibility studies.

By engaging in such vital, advanced and varied projects and by associating with pioneering engineers and scientists, you can make rapid strides toward your own professional goals.

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Parametric Microwave Amplifiers
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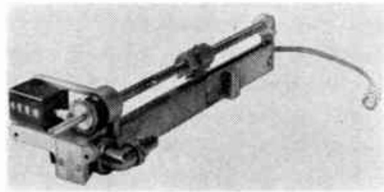
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 140A)



Its general specifications and characteristics are as follows: Linearity: −0.1 db per 30 db. Attenuation: adjustable over a 100 db range. Insertion Loss: approximately 25 db. Power Dissipation: to 5 kw peak. Frequency Range: 925–1225 mc. Attenuation Rate: 4 db per turn of drive shaft. Standing Wave Ratio: 1.25 to maximum. Type BNC Connectors.

Stops are provided to prevent operation in the non-linear region. No Load is supplied, but GCC's Model 3001, 3051 or any other suitable termination may be employed. Other models, with knobs, varying cable lengths, and other types of RF connectors, are available.

For specific information, write to the firm.

McClenahan Named by Sperry

William J. McClenahan has been named Marketing Manager of **Sperry Rand Corp.'s Sperry Electronic Tube Div.**, Gainesville, Fla

Division General Manager John R. Whitford, who announced the promotion, stated that creation of the new position reflects the division's increased emphasis on a market oriented, rather than a product oriented approach to microwave tube sales and development.

McClenahan's new responsibilities include contracts, sales planning, advertising, and promotion of the division's line of klystron and traveling wave tubes. These responsibilities cover both U. S. and foreign markets.

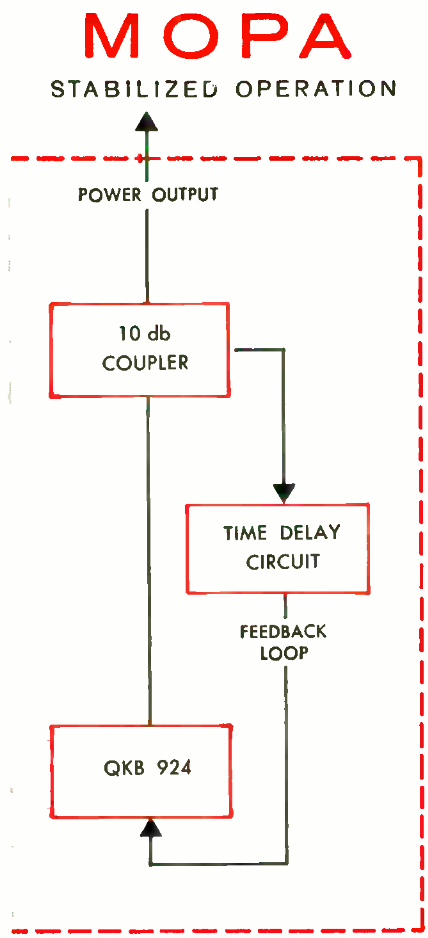
Before transferring to the Sperry sales organization in 1956, McClenahan was Field Engineering Supervisor in Sperry's Southwest District. In that capacity, he had experience with both military and commercial electronic systems. He was named Sales Manager of the Gainesville operation in 1958.

Digital Display Devices

These new digital display instruments from **Servo Development Corp.**, 2 Willis

(Continued on page 144A)

QKB 924

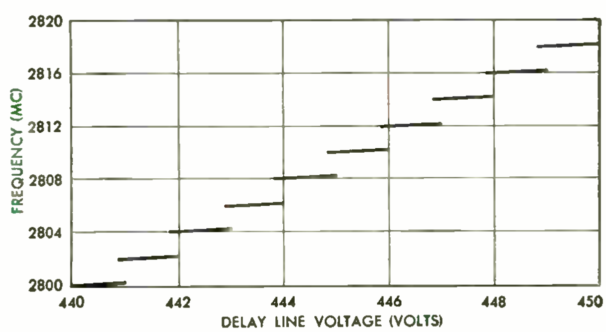


New Raytheon master oscillator assures extreme stability in frequency diversity transmitters

The QKB 924 voltage tunable "O" type backward wave oscillator with the feedback circuit shown above provides a highly stable master driver—local oscillator for S-Band MOPA chains.

External feedback through a delay line provides a 10:1 or more increase in frequency stability—performance that is particularly suitable for frequency diversity MTI applications. The frequency vs. voltage curve of the circuit is essentially flat at discrete steps over the entire 2,700 to 3,200 Mc range. Power output is typically 100 milliwatts with a delay line tuning voltage of 350 to 700 volts. Models are also available at frequencies through X-Band.

Write for detailed application information to Raytheon Company, Microwave & Power Tube Division, Waltham 54, Massachusetts. In Canada: Waterloo, Ontario.



TYPICAL STABILIZED OPERATION

RAYTHEON COMPANY

RAYTHEON

MICROWAVE AND POWER TUBE DIVISION

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DAYTON, OHIO, BAldwin 3-8128 • LOS ANGELES, CALIF., PLymouth 7-3151 • CANADA: Waterloo, Ont., SHerwood 5-6831

NEWS New Products

(Continued from page 142A)

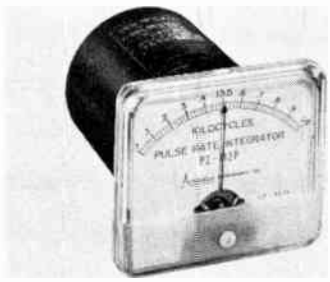
Court, Hicksville, L.I., N. Y., incorporate precision gearing and a wide selection of rotating components for introducing heading, speed, elevation, bearing, or other voltage settings, into servo systems. The units are available with a wide selection of single and dual speed synchros, potentiometers, or shaft encoders.

They are contained in 2" diameter MS33639 sealed cases and the assemblies qualify to MIL specs. The counters are new designs and are available in special configurations and calibrations, and with red or white integral lighting. Backlash is reduced to approximately 2 minutes through use of AGMA precision 2 gears and ABEC-7 ball bearings.

Further information on specifications or applications will be supplied upon request.

Frequency Deviation Meter

A new meter for measuring frequency deviation is announced by **Anadex Instruments, Inc.**, 14734 Arminta St., Van Nuys, Calif. Called the Series PI-111 Pulse Rate Integrator, it is designed for monitoring power line frequencies, turbine flow sensors, tachometers, rotating machinery, repetition rate picks-ups, radiation detectors, and other frequency generating devices.



The integrator features an input sensitivity as low as 10 mv rms. Standard models are available to indicate 57 to 63 cps and 380 to 400 cps. Variations are also available with center frequencies from 50 cps to 50 kc. Resolution is 0.1%. The Pulse integrator features an integrally mounted panel-type meter. Standard meter sizes are 3½ wide or 4½ inches wide. Special meters are available.

The models are priced starting at $150.00 each. Delivery is thirty days after receipt of order. Quantity prices are available upon request.

For additional information write direct to the manufacturer.

Curve Follower

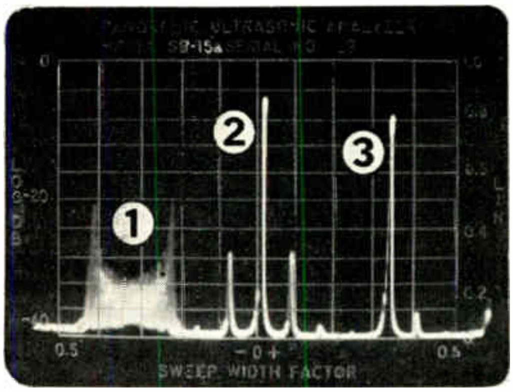
It is only necessary to plot a repetitive function variation once with the new DATA-TRAK curve follower offered by **Research, Inc.**, Box 6164, Minneapolis 24, Minn. Continuous rotation of the data drums in the new models produces cyclical output signal variations from a pencil drawn "pattern."

Model FGE 5048 Data-Trak interprets function curves drawn as a double line with ordinary pencil on common graph paper. No special paper, pencils, or conductive inks are required. The pencilled chart, mounted on Data-Trak's revolving drum, guides a servo driven capacitive probe that follows the centerline between the plotted lines. A potentiometer, geared to the probe drive mechanism, divides any impressed voltage in precisely the same proportion as the drawn curve

(Continued on page 146A)

NOT JUST ULTRASONIC...

•ULTRA-Compact (only 8¾" high)
•ULTRA-Versatile (so many applications)
•ULTRA-Fast·Easy-to-read·Economical



Lab setup shows SB-15a versatility. (1) FM display measures dynamic deviation. (2) & (3) are AM and SSB signals, respectively, with sine wave modulation.

PANORAMIC's SB-15a SPECTRUM ANALYZER 0.1 KC TO 600 KC

Panoramic's advanced Model SB-15a automatically and repetitively scans spectrum segments from 1 kc to 200 kc wide through the entire range (0.1 kc to 600 kc) . . . plots frequency and amplitude along the calibrated X and Y axes of a long persistence CRT, or on a 12 x 4½" chart (optional RC-3a/15). Sweep rates are adjustable from 1 to 60 cps.

Adjustable resolution enables selection and detailed examination of signals as close as 100 cps. Self-checking internal frequency markers every 10 kc. Also internal amplitude reference • Only 8¾" high, the SB-15a is completely self-contained, needs no external power supply or regulator.

PANORAMIC PRESENTATION MEANS

- quick signal location, minimum chance of missing weak signals or holes in spectrum
- faster measurements—no tedious point-by-point plots
- reliable spotting of low level discrete signals in noise
- positive identification and dynamic analysis of all types of modulation

ALL THESE APPLICATIONS . . .

- Noise, vibration, harmonic analysis
- Filter & transmission line checks
- Telemetry analysis
- Communication System Monitoring . . . and more
- Power Spectral Density Analysis (with Model PDA-1 Analyzer)
- Frequency Response Plotting (with Model G-15 Sweep Generator)

Write now for specifications, other applications of PANORAMIC's Model SB-15a. Get on our regular mailing list for THE PANORAMIC ANALYZER, featuring application data.

SUMMARY OF SPECIFICATIONS

Frequency Range: 0.1 kc to 600 kc.

Sweepwidth: Variable, calibrated from 1 kc to 200 kc.

Center Frequency: Variable, calibrated from 0 to 500 kc.

Markers: Crystal controlled, 10 kc and 100 kc plus harmonics.

IF Bandwidth: Variable, 100 cps to 4 kc.

Sweep Rate: Variable, 1 cps to 60 cps.

Amplitude Scales: Linear, 40 db log (extendable to 60 db) and 2.5 db expanded.

Sensitivity: 200 µv to 200 v full scale deflection.

Accuracy: ± 0.5 db.

Input Impedance: 55,000 ohms.

eem Sec. 2900

dependable CERTIFIED SPECIFICATIONS for accurate data

PANORAMIC RADIO PRODUCTS, INC.

522 So. Fulton Avenue, Mount Vernon, N. Y.

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the pioneer is the leader

THE MARK OF THE ORIGINAL PANORAMIC

Panoramic RADIO PRODUCTS, INC.

(Continued from page 144A)

divides the graph scale. Since the probe does not actually touch the paper, no chart wear occurs and the charts can be stored and reused indefinitely. Data drum rotation time is adjustable from 300 to 10 seconds per revolution (12 to 360 cycles per hour) on standard models, with other ranges available. Operation may be continuous, or limited to an arbitrary number of cycles by a predetermined counter. Units are available with one, two, or three channels. Drum rotation is synchronized on the 2- and 3-channel models. A new, smaller case has hinged cover for easy access to chart drums.

Principal uses are in guiding automated processes and in programming analog computers and controllers.

For complete information write to the firm.

Relay-Potentiometer

A full report on the Raytheon Raysistor, an electro-optical, relay-potentiometer, is contained in a brochure published by **Raytheon Company's Industrial Components Div.**, 55 Chapel St., Newton 58, Mass.

The brochure, utilizing cross-sectional diagrams, circuit diagrams, charts and photographs, explains the electrical and mechanical characteristics of the new basic-circuit component and how it is designed to replace relays, switches, and potentiometers for low-noise commutation, switching, and controlling circuits.

Use of the Raysistor as a control link in either the tube or transistor version of the AGC loop, as a suppressed carrier modulator; a single-pole, double-throw relay; a commutator; or as a negative resistance generator; is indicated in the free brochure.

The brochure may be obtained from William H. Weed at the firm.

Plug-In Chopper

The Solid State E ectronics Co., 15321 Rayen St., Sepulveda, Calif., announces the availability of its new Model 66 plug-in line-driven chopper (or modulator).

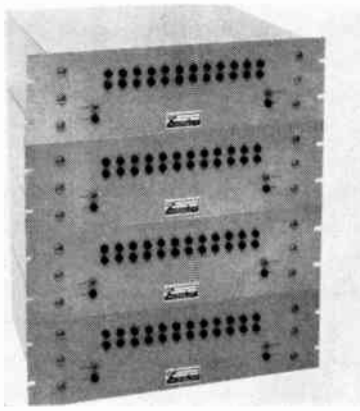
The unit incorporates the addition of a transformer-coupled isolating drive network and clipping diodes so that it can, for example, be driven from a 60 or 400 cps sine wave power line or from a drive source that is common to the dc voltage being chopped.

Sinusoidal or square wave drive may be utilized over a frequency range extending from 20 cps to 12 kc sine wave or 30 cps to 5 kc square wave. This unit may also be used as a synchronous demodulator to convert an ac signal to dc.

It is capable of linearly switching or chopping voltages over a wide dynamic range which extends down to a fraction of a millivolt and up to 5 volts.

Sonic Delay Line

Control Electronics Co., 10 Stepar Place, Huntington Station, L. I., N. Y., has developed a sonic type delay line, with good phase and attenuation characteristics as well as delay of 0.1 second.

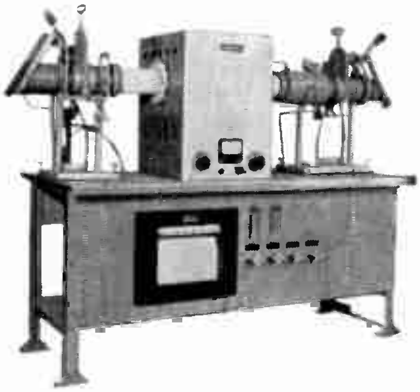


(Continued on page 150A)

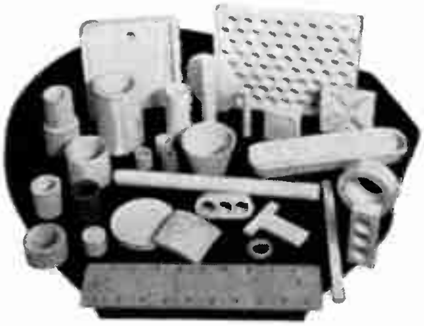
Horizontal Zone Scanner for zone purification of metallic germanium and other metallic and organic materials.

Low Temperature Furnace for solar cells and treatment of metals.

Vertical Floating Zone Scanner for accurate and precise production of hyper-pure semiconductor materials and metals.



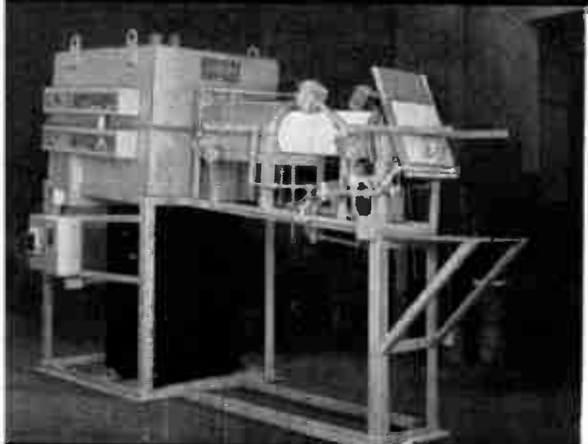
Batch Type Atmosphere Furnace for low temperature alloying.



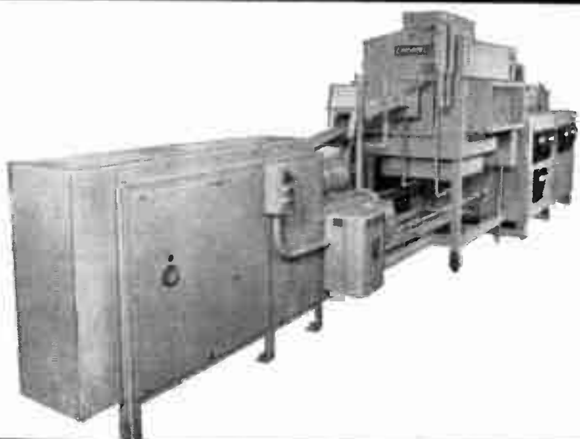
We can engineer and produce refractory designs and mixes—dishes, boats, crucibles of zircon, zirconia, alumina, pure oxides, etc., for specific needs.

High Temperature Crucible Pot Furnace for melting. Maximum temperature 1600°C.

for Semiconductors Lindberg now comprehensive



Hydrogen Atmosphere Furnace for reduction of germanium oxides and other materials.



Hump Type Mesh Belt Conveyor Furnace for precision alloying of transistors and diodes.

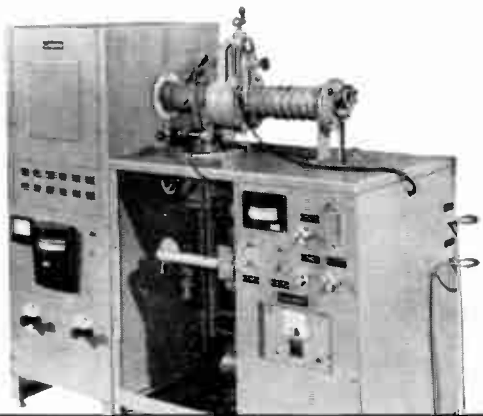
High Temperature Tube Furnace for germanium ingots.

Dual frequency unit for zone refining, one of a complete line of induction heating units.

Mechanical Pusher Type Furnace for production of transistors by gas diffusion.

Precision Gas Diffusion Furnace with multiple controls for quality transistor and semiconductor bodies.

High Temperature Gas-Tight Box Furnace for solar cells.



Vacuum unit for gas diffusion in alloy muffle or refractory.

offers the most advanced designs, the most line, in efficient, precise, production equipment.

Semiconductor and electronics manufacturers now can find in Lindberg one experienced source for dependable equipment for production of semiconductors, solid state and other electronic devices. The past several years a special staff of our experienced electronic, mechanical and electrical engineers, metallurgists, ceramicists and chemists has concentrated on the design and development of improved equipment for these uses. This group of experts, with many year's experience in all phases of the application of heat to industry, plus Lindberg's comprehensive research facilities provide the right combination to keep pace with the expanding requirements for reliable, efficient, precision equipment in these rapidly growing industries.

Shown here are some Lindberg units now in operation in plants throughout the country. We believe they provide collectively the most comprehensive line available today.

All of this equipment is operating in our research laboratories. We invite you to visit us, meet our able research staff, see at first hand—from crystal growing, to purifying materials, to finished product—how Lindberg offers a complete and outstandingly competent source for consultation on your semi-conductor or electronic production problems. For additional information please get in touch with your local Lindberg representative (See your classified phone book) or write directly to us.

Semiconductor Division

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Los Angeles plant: 11937 S. Regentview Avenue, Downey, California. In Canada: Birlefco-Lindberg Ltd., 15 Pelham Ave., Toronto 9, Ont. Also, Lindberg plants in Argentina, Australia, England, France, Italy, Japan, South Africa, Spain, Switzerland and West Germany.

PROFESSIONAL CONSULTING SERVICES

NEWS New Products

(Continued from page 146A)

Designated as Model FA-573, the delay line has an attenuation of 20 db at 300 cps and impedance of 1K ohm. With multiple taps, every 1 millisecond, this delay line has 100 taps with tap accuracy of better than 0.5%. Additional taps can be supplied upon request. Cut-off frequency is at 400 cps and phase linearity is better than 0.75% up to 300 cps.

The delay line is rack mounted and measures 19 long, 10 deep and 21 inches high. Weight is 200 pounds. This delay line which has geological auto-correlation applications can be cascaded with some sacrifice in bandwidth and attenuation.

Availability is 6 to 8 weeks. Prices are available upon request. For additional information write to the firm.

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The industry's broadest line of COMMUNICATIONS TRANSISTORS

NEW! Transistor Guide

for Communications Circuit Designers

To make it faster and easier for you to find the Philco communications transistor that best meets your needs, put this new guide to work for you. 12 pages of application information, descriptions of major types and a complete selector chart will speed your work. For your copy, write Dept. IRE261.

You'll find one Exactly Right for your requirements

Whatever your specific requirements . . . in gain, noise figure, AGC and other electrical characteristics . . . you'll find a Philco communications transistor that meets your needs precisely.

Unlike other manufacturers, who offer only limited lines of "general-purpose" transistors, Philco designs and produces transistors to meet specific requirements. Philco can do this only because the exclusive Philco Precision-Etch* process of manufacture permits precise control of all parameters and absolute uniformity.

To increase circuit efficiency and reliability, to reduce cost and to make your design easier . . . look first to the Philco line, which offers you the industry's widest range of communications transistors. You'll find one *exactly right* for your requirements.

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Philco Transistors are immediately available in quantities 1-999 from your Philco Industrial Semiconductor Distributor.

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TRANSISTOR CENTER U.S.A.

BASIC

because...

It Measures

VOLTAGES
from 0.5 mv to 10v, dc

CURRENTS
from 0.005 $\mu\mu$a to 1 ma, dc

RESISTANCES
from 0.3 MΩ to 500 MMΩ

. . . it can also be used as a d-c amplifier with a current gain of over 16 x 10^9 . . . use it as a "front end" to drive recorders and other instruments

- True direct-coupled amplifier design — no choppers, vibrators, or reeds to give mechanical trouble or get out of adjustment.
- High input resistance: 100 mega-meg-ohms — does not load down sensitive circuits. Input resistance can also be set by switch to other values from 10^4 to 10^{11} ohms.
- Constant low voltage across unknown (9.1v) — no need for voltage coefficient corrections when making high resistance measurements.
- Output for 5-ma recorders.

Type 1230-A D-C Amplifier and Electrometer. . . . $440

Write For Complete Information

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WEST CONCORD, MASSACHUSETTS

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