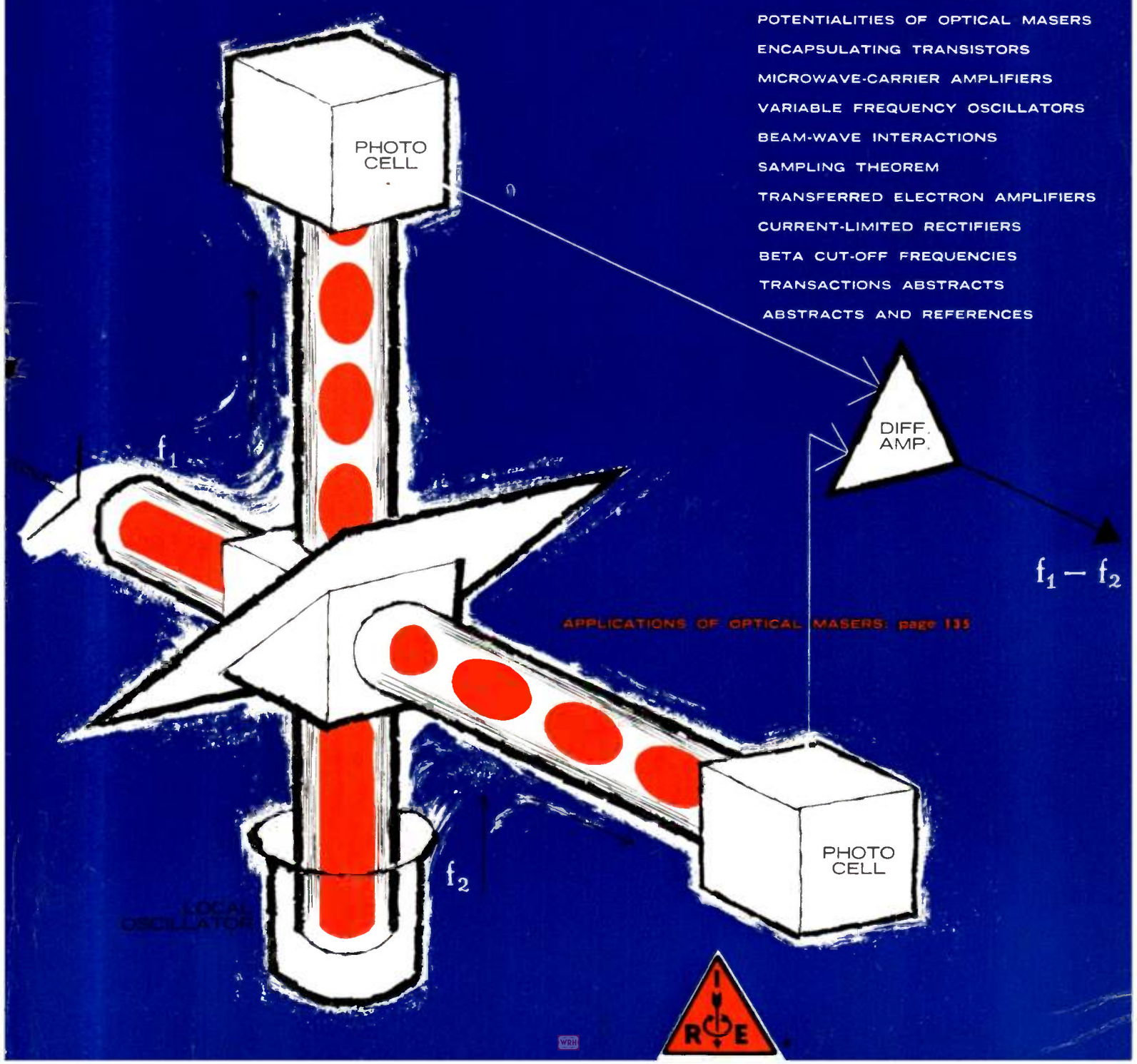
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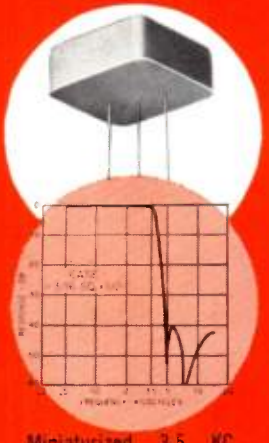
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VARIABLE FREQUENCY OSCILLATORS
BEAM-WAVE INTERACTIONS
SAMPLING THEOREM
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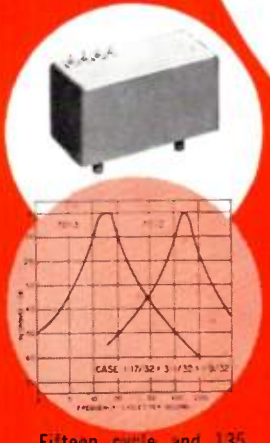
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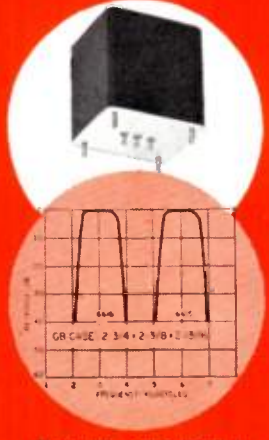


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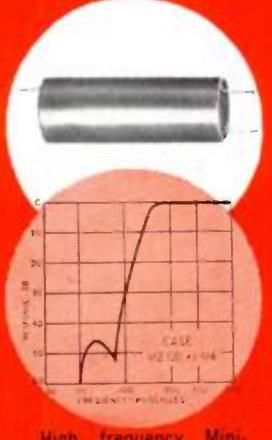


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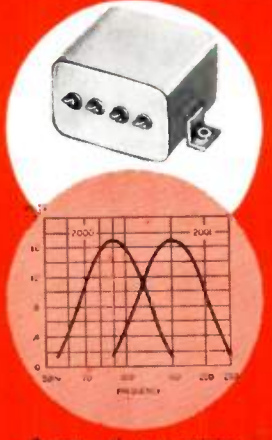
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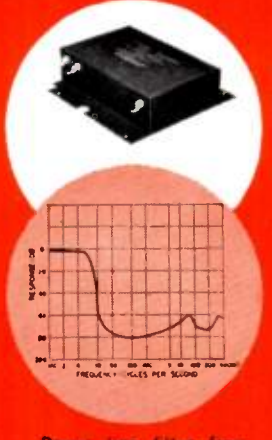
Three KC and 6 KC flat top band pass filters. 400 ohms to 20K ohms. MIL-T-27A: each filter 1.7 lbs.



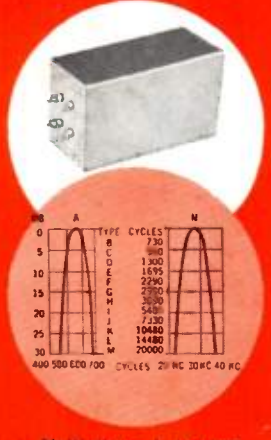
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The coherent beams produced by optical masers can be converted to radio frequencies by heterodyne techniques. Shown on the cover is a balanced photoelectric mixer consisting of a half-reflecting half-transmitting mirror, two photocells and a differential amplifier. For an introduction to optical maser principles and applications, see page 135.

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On Getting Something for Almost Nothing

For a lightweight, low-power-consuming KU-band superhet receiving system for a new combat surveillance radar equipment, we were confronted with the need for getting good sensitivity and overall performance without being able to support from a power standpoint the usual klystron local oscillator (LO) that one might expect to use. As a result of other developmental work on this project that we are doing for the U.S. Army Signal Corps, we found that we could conveniently, and with very little power consumption, generate between 10 and 100 microwatts of LO power, and the problem then became one of developing some means of getting the maximum sensitivity at these low power levels.

An old trick that is probably familiar to all of us from one time or another is the use of forward DC bias as a means of getting a mixer crystal to the right operating point without depending on the LO signal itself. Experiments made here at X-, KU-, and K-band confirmed that, with proper implementation of DC biasing, we could obtain close to normal sensitivity with the available LO input power.

Figure 1 shows some typical data taken at KU-band for tangential sensitivity as the function of LO power. As can be seen from the figure, the operation with DC bias compares favorably with conventional operation of the mixer. Since the system under design required good linearity, an important consideration was the available dynamic range for the mixer as a function of LO drive. Figure 2 shows some data taken with DC bias at various LO powers; it also shows that good linearity and dynamic range can be obtained to −30 dbm input signal level with only −20 dbm of LO power.

All the experiments were run using a balanced mixer configuration, since this allows us to put the LO signal into the mixer without the losses associated with the loose coupling customarily used with single-ended mixers.

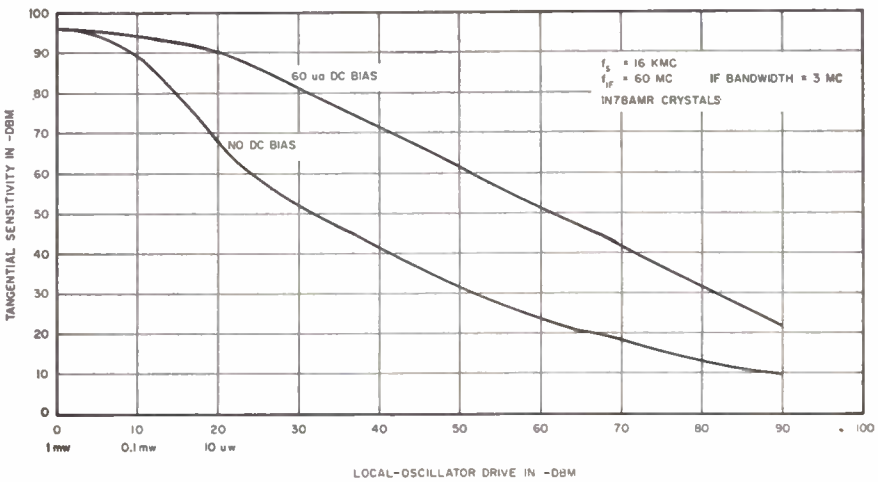


FIGURE 1. TANGENTIAL SENSITIVITY AS A FUNCTION OF LOCAL-OSCILLATOR DRIVE

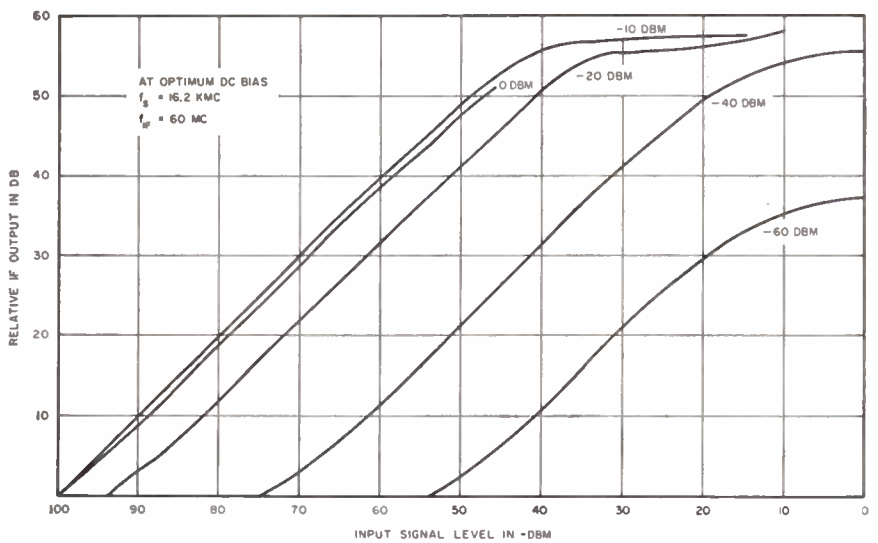


FIGURE 2. DYNAMIC RANGE AS A FUNCTION OF LOCAL-OSCILLATOR DRIVE

The low LO technique is currently being used in the final equipment and has resulted, for this particular equipment, in reducing the overall input power required by about one-third (a saving of about 10 watts) without any significant losses in performance; therefore, it has permitted a considerable reduction in the size and weight of the self-contained power source as well as a reduction in the size and weight of the power converter circuits. The excellent performance obtained by using the low LO DC biasing technique makes it very useful to systems where equipment size, weight, and power consumption are of paramount importance but optimum system performance must be maintained.

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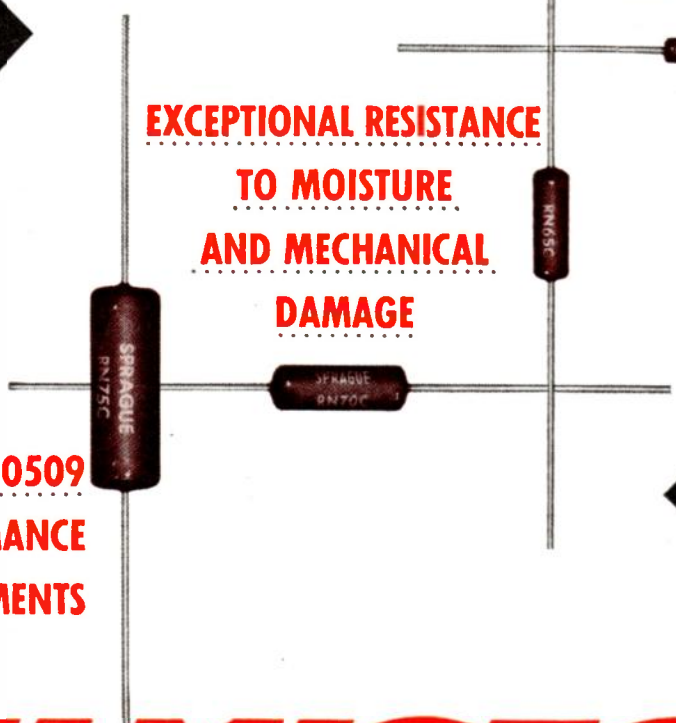
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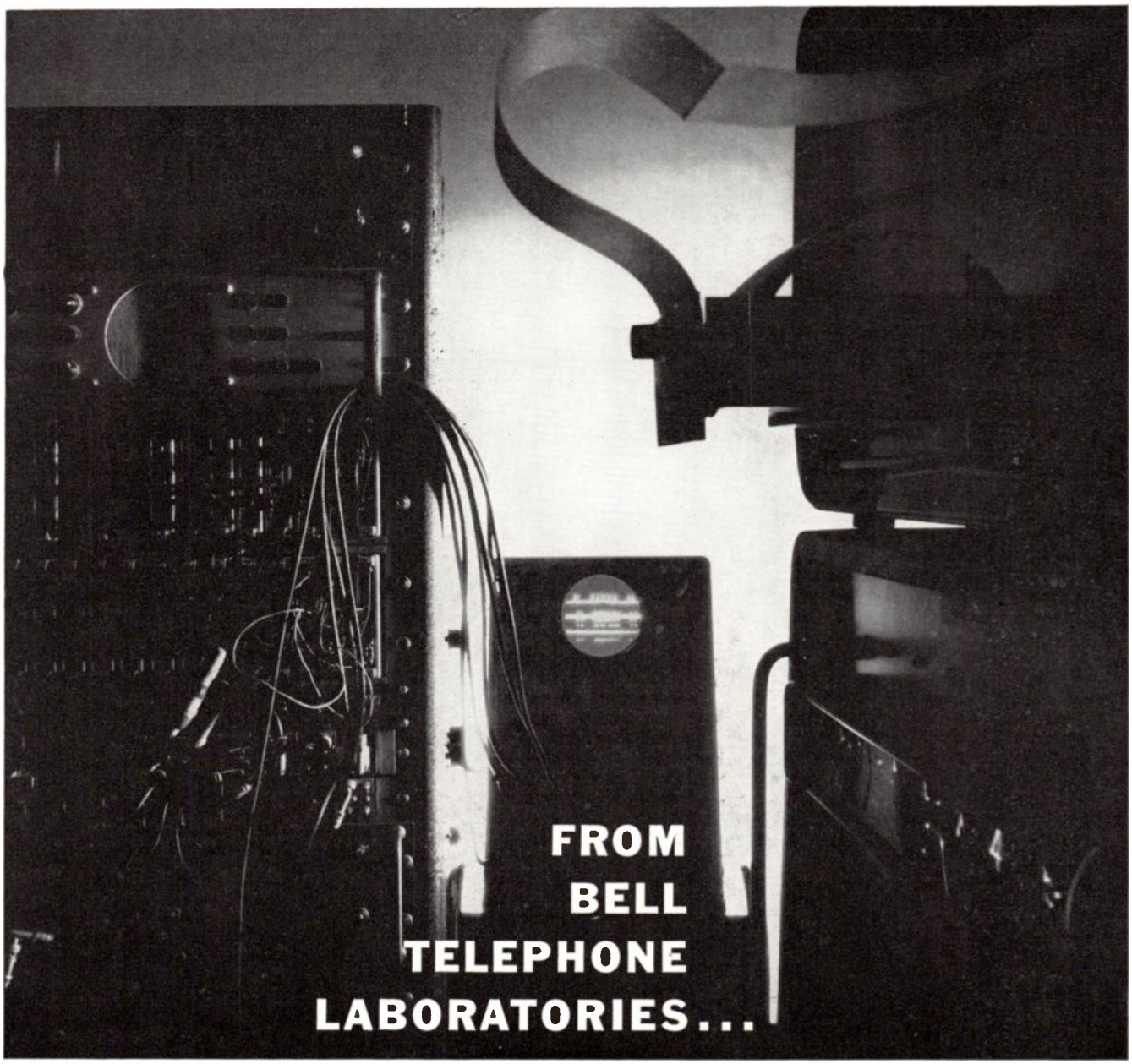
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The fundamental capabilities of pulse transmission are under study at Bell Laboratories. At a transmission rate of 200 million bits per second, for example, PCM could simultaneously transmit 3000 telephone conversations on a single circuit.

AN INTRIGUING DEVELOPMENT IN TELEPHONE TRANSMISSION

Bell Laboratories engineers have applied a method of transmitting telephone conversations which uses a series of ON-OFF pulses rather than the continuous electrical signals generally used since the time of Alexander Graham Bell's first famous message.

The method is called Pulse Code Modulation. With PCM the telephone caller's voice is sampled every 1/8000th second. Each sample is then encoded into a series of ON or OFF pulses, and these pulse groups are sent over the regular telephone line. Spaced periodically along the line are repeaters which clean up and amplify the pulses. At the receiving end the pulse groups are decoded and the caller's voice is reconstructed.

Since the pulses are of very short duration, it is possible to interlace many different voice messages and send them all over one line. For example, in a PCM system now operating between Newark and Passaic, N. J., a single pair of wires carries as many as 24 one-way voice signals.

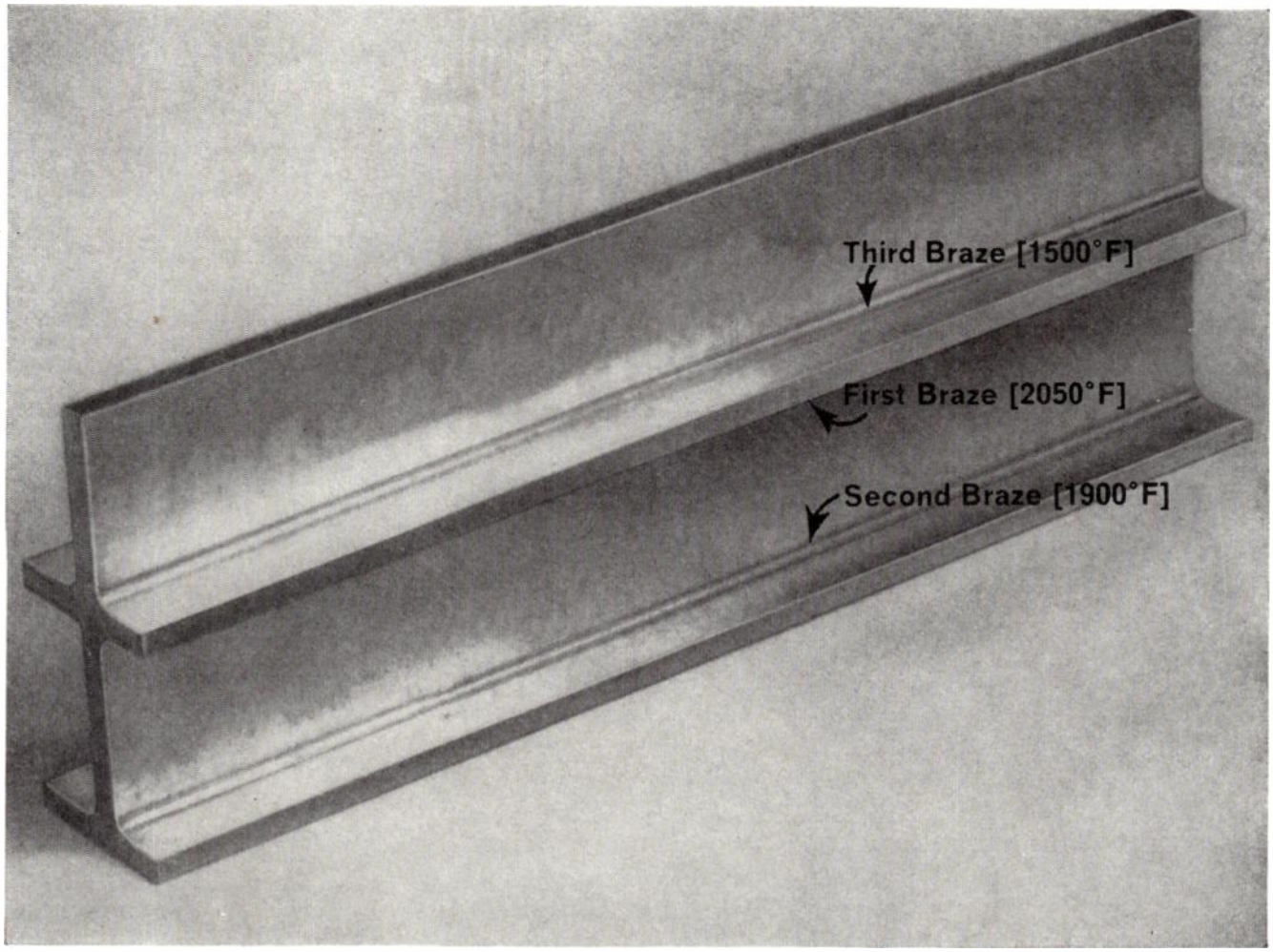
Other systems for carrying more than one voice signal over a single telephone line have been developed and are in widespread use. PCM, however, provides special advantages, for example, in cable circuits connecting telephone offices in a congested metropolitan area.

PCM in its present practical form for cable circuits has been made feasible by Bell Laboratories' invention and development of the transistor, the key element necessary for a small economical system.

Currently, PCM systems carrying much larger bundles of communication channels are under study at Bell Laboratories. The goal as always is the improvement of Bell System communication services.

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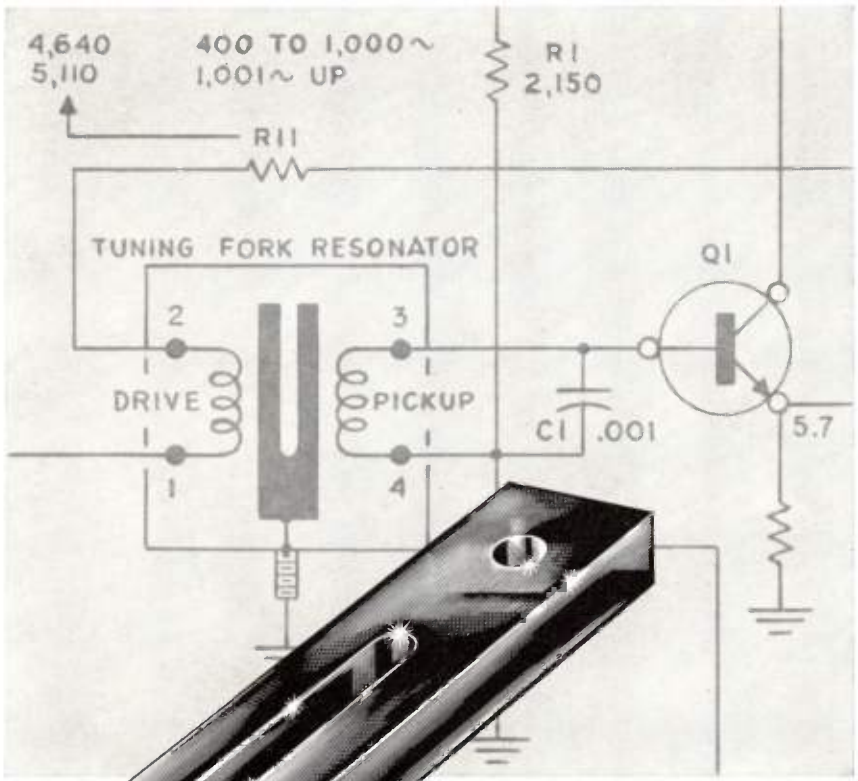
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Meetings with Exhibits

● 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.

Δ

March 1-3, 1962

Eighth Scintillation and Semiconductor Counter Symposium, Shoreham Hotel, Washington, D.C.

Exhibits: Dr. George A. Morton, RCA Labs., Princeton, N.J.

March 26-29, 1962

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

Exhibits: Mr. William C. Copp, IRE Advertising Dept., 72 West 45th St., New York 36, N.Y.

April 11-13, 1962

SWIRECO (South West IRE Conference & Electronics Show), Rice Hotel, Houston, Texas

Exhibits: Mr. J. C. Robinson, P.O. Box 1505, Houston, Texas

May 1-3, 1962

Spring Joint Computer Conference, Fairmont Hotel, San Francisco, Calif.

Exhibits: Mr. John W. Ball, Pacific Telephone Co., 3240 Arden Way, Sacramento, Calif.

May 14-16, 1962

NAECON (National Aerospace Electronics Conference), Biltmore Hotel, Dayton, Ohio

Exhibits: Mr. Robert J. Stein, 19 Beverly Place, Dayton 19, Ohio

May 23-25, 1962

National Telemetering Conference, Sheraton Park Hotel, Washington, D.C.

Exhibits: Mr. H. W. Royce, Martin Company, Baltimore, Md.

June 12-14, 1962

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 18-19, 1962

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

Exhibits: Mr. John H. Landeck, Admiral Corp., 3800 W. Cortland St., Chicago 47, Ill.

June 25-27, 1962

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

Exhibits: Mr. Ralph I. Cole, Melpar, Inc., 3000 Arlington Blvd., Falls Church, Va.

(Continued on page 10A)

ML-8087 Precision

Scan Conversion Tube
Fast Erase — High Resolution



The Machlett Laboratories, Inc. announces a new precision manufactured Scan Conversion Storage tube, the ML-8087. Successor to the Machlett made 403X type tubes, which have seen over one million hours use in airways control service, the ML-8087 provides these principal advantages and important features:

1. High resolution: a minimum of 180 range rings/diameter at 50% amplitude modulation; equivalent to 900 TV lines.
2. Fast erase: less than 2 seconds erase cycle to reduce stored information to noise level.
3. Wide storage range: to meet FAA 1213b specification and beyond.
4. High signal/noise ratio, typically 80:1 (peak signal to rms noise).
5. Rapid set-up time: A few minutes installation is all that is required to adjust tube for optimum operation. No need for critical dynamic focussing of electron beams.
6. No variation of output signal with size of written area.
7. Only simple video circuits are needed for readout.

Available now from The Machlett Laboratories, Inc. . . . send for complete data on the ML-8087.

MACHLETT

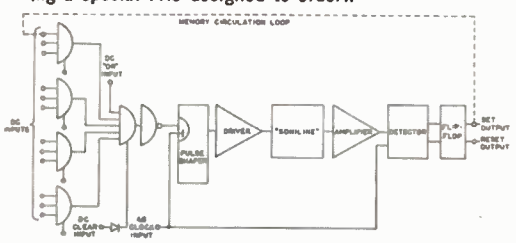
The Machlett Laboratories, Incorporated
A division of Raytheon Company
Springdale, Connecticut

WHAT ARE S-PACS?

Concisely, **S-PACS** are an integrated family of high-speed, compatible digital modules for **dc** to **five** megacycle operation!!

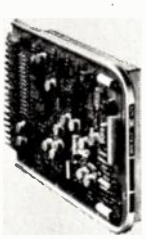
Outstanding quality and **maximum economy** are combined in this series of **S-PAC** digital circuit packages!! Specifically, **S-PACS** are noted for their consistently **reliable** operation — a most significant factor in the successful, **direct** implementation of digital systems or computers!!

Noteworthy among this high-quality **S-PAC** series is the **Model SM-30***— a plug-in serial magnetostrictive delay storage element with a maximum storage capacity of **1500 bits** at a **1 megacycle** rate!! **Serial Memory** is achieved by feeding the **SM-30** output **back** to the input and **recirculating** the stored data! **Increased storage** capacity can be obtained by **cascading several SM-30 units directly**, or by using a **special** PAC designed to order!!



SM-30 FUNCTIONAL DIAGRAM

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Meetings with Exhibits

(Continued from page 8A)

August 21-24, 1962

Western Electronics Show and Conference (WESCON), Ambassador Hotel & Memorial Sports Arena, Los Angeles, Calif.

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

August 29-September 1, 1962

Second International Conference on Information Processing, Munich, Germany

Exhibits: Dr. Heinz Billing, Max Planck Institute, Munich, Germany

October 1-3, 1962

Eighth National Communications Symposium, Hotel Utica & Utica Municipal Auditorium, Utica, N.Y.

Exhibits: Mr. R. E. Gaffney, General Electric Co., Light Military Electronics Dept., Utica, N.Y.

October 8-10, 1962

National Electronics Conference, Exposition Hall, Chicago, Ill.

Exhibits: Mr. Rudy Napolitan, National Electronics Conference, 228 N. LaSalle St., Chicago, Ill.

October 29-31, 1962

15th Annual Conference on Electronic Techniques in Medicine & Biology, Edgewater Beach Hotel, Chicago, Ill.

Exhibits: Dr. J. E. Jacobs, 624 Lincoln Ave., Evanston, Ill.

November 13-15, 1962

Northeast Research & Engineering Meeting (NEREM), Somerset Hotel & Commonwealth Armory, Boston, Mass.

Exhibits: Mr. Stewart K. Gibson, c/o IRE Boston Office, 313 Washington St., Newton, Mass.

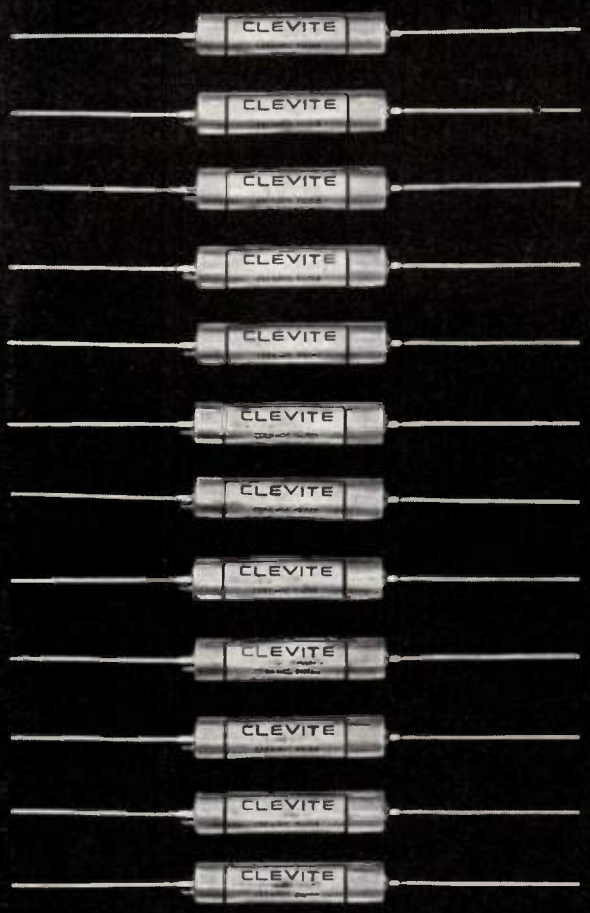
December 4-5, 1962

Eastern Joint Computer Conference, Bellevue-Stratford Hotel, Philadelphia, Pa.

Exhibits: Mr. Charles Phillips, 5603 Jordan Rd., Washington 16, D.C.

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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.



MODEL	6db BANDWIDTH
TL-2D5A	2 kc
TL-4D9A	4 kc
TL-6D12A	6 kc
TL-8D16A	8 kc
TL-10D18A	10 kc
TL-16D25A	16 kc
TL-20D32A	20 kc
TL-30D45A	30 kc
TL-40D55A	40 kc
TL-45D60A	45 kc
TL-32E48C	32 kc
TL-50D85C	50 kc

(suffix "A" denotes 455 kc center frequency;
suffix "C" denotes 500 kc center frequency)

Clevite Ceramic Ladder Filters

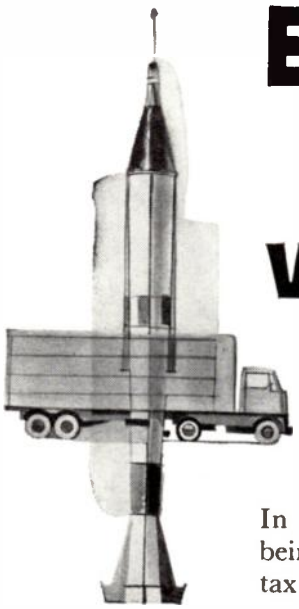
Now in stock in 12 bandwidths... 80 db rejection in 0.1 cu. in.

Clevite ceramic ladder filters provide more selectivity for their size than any conventional i-f filter. They are fixed tuned and need no alignment—are non-magnetic and non-microphonic. Leading manufacturers now have Clevite ladder filters in their communications equipment. Improve your newest design with these unique filters. Write now for complete specifications—Bulletin 94012, or for selectivity curves available on each stock model. ■ Dimensions: 5/16" diameter x 1½" long. ■ Selectivity: 60 db/6db shape factor from 1.3:1 to 2.6:1. ■ Center Frequency Stability: within 0.2% for 5 years, and within 0.2% from −40° to +85° C. ■ Impedance: 1200-1500 ohms. ■ Designed for military environment.

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In our world, a defense system cannot be merely adequate. We cannot risk being second best. We cannot afford to demand less than full value for every tax dollar spent.

Industry's job is to grasp the full implication of the mission, to design and build the system to meet cost requirements and delivery schedules. And to *insure* that it will perform when the buttons are pressed.

The company which can do this job has a total capability. At Hughes, this capability has many parts—a record of 15 years of scientific achievement, a working force of over 28,000, facilities covering 5.3 million square feet, vigorously applied Value Engineering and Cost Improvement programs, a fund of over 100,000 *man-years* of systems experience.

These are impressive statistics even when taken piecemeal. But they are most meaningful when seen as parts of a functional unity. For Hughes can be likened to a dynamic, *creative system* which has proved its competence and spirit of purpose many times. Examples:—*The "Missile Monitor" truck-mounted mobile command and control air defense system — 16,000 airborne fire control systems — Surveyor soft lunar landing spacecraft — 33,000 Falcon air-to-air missiles.*

The broad capabilities reflected here fully equip Hughes to assume complete management responsibility for systems assembly, check-out and integration. This competence is backed by Hughes abilities in manufacturing, assembly, operator training, operational activation, field support and maintenance.

Expect more from Hughes. Hughes is ready to deliver.

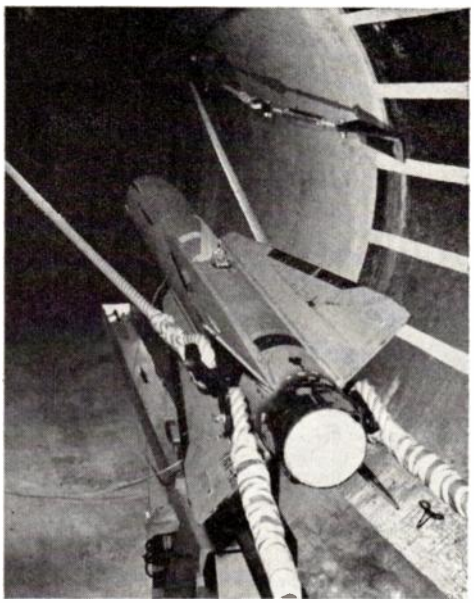
Creating a new world with electronics

HUGHES

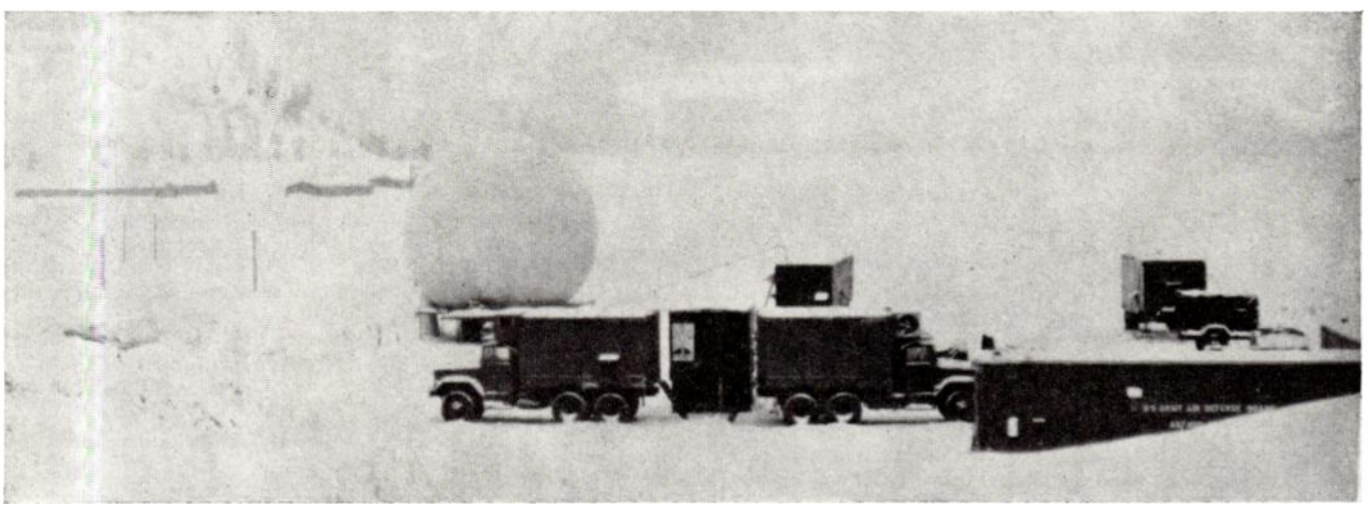
HUGHES AIRCRAFT COMPANY

Systems integration, as well as design and development was Hughes responsibility in the MA-1/F-106 program. Nine similar Hughes systems have preceded and followed it—a total of 16,000 airborne systems having beer produced. Hughes has been the sole supplier to the U.S. Air Force of all-weather interceptor fire control systems for 12 years.

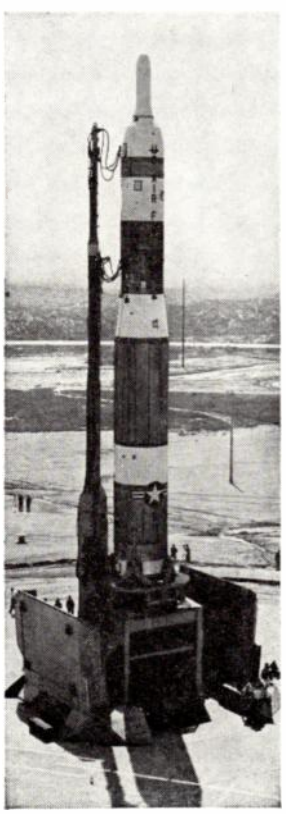




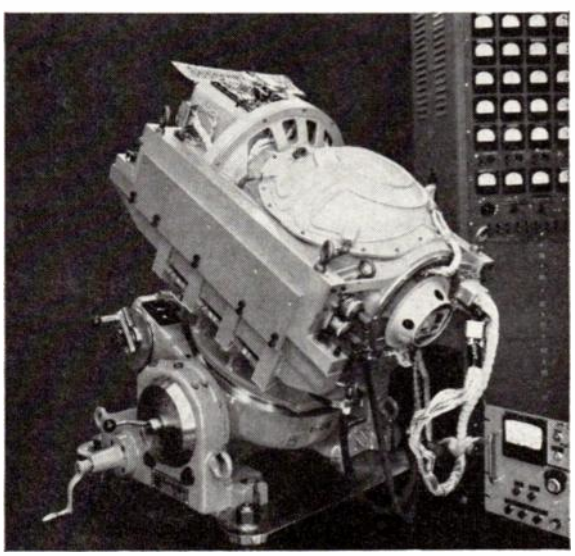
Propulsion facilities—Hughes has tested over 1,500 missiles in every conceivable mode of operation. The Hughes Tucson plant has extensive facilities for solid propellant and explosives storage and final missile assembly. The company maintains special propellant and propulsion capabilities to support procurement of solid fuel engines.



Command control experience—The "Missile Monitor" mobile electronic air defense system performs target acquisition and missile directing functions for Nike Ajax, Nike Hercules and Hawk batteries. At Hughes Fullerton, some 400 acres are devoted to special field test ranges for electronics systems and the Munsen course for full-scale testing of instrumented vehicles under the most rugged conditions. The Environmental Laboratory has one of the nation's largest low frequency and shock facilities.



Field service and support experience — Hughes large force of experts is permanently assigned in NATO, SEATO and the free countries of the world to insure that every Hughes missile and electronics system will deliver what it should. Hughes engineers work with the U. S. Air Force on the MA-1/Falcon system, with the U. S. Navy on the Polaris, with the U. S. Army on "Missile Monitor." A Hughes team is managing the Titan site activation (photo) at Mountain Home, Idaho.



Guidance experience—A Polaris missile guidance unit undergoing check-out. Other Hughes work in the area includes: Polaris Submarine Fire Control Systems, special ARPAT studies, airborne tactical attack systems, inertial platforms and components, stellar tracking systems and digital computers.

IRE News and Radio Notes

Current IRE Statistics

(As of December 31, 1961)

Membership—96,551
Sections—111
Subsections—31
Professional Groups—29
Professional Group Chapters—291
Student Branches†—217

† See June, 1961, issue for a list.

Calendar of Coming Events and Author's Deadlines*

1962

Feb. 6–7: Symp. on Redundancy Techniques for Computing Systems, Dept. of Interior Auditorium, Washington, D. C.

Feb. 7–9: 3rd Winter Conv. on Military Electronics, Ambassador Hotel, Los Angeles, Calif.

Feb. 14–16: Internat'l. Solid State Circuits Conf., Philadelphia, Pa.

Feb. 14–16: Annual Meeting of Biophysical Society, Sheraton Park Hotel, Washington, D. C.

Feb. 19–24: Winter Inst. on Optimum and Adaptive Control Systems Theory, Univ. of Florida, Gainesville.

Mar. 1–3: 8th Scintillation and Semiconductor Counter Symp., Shoreham Hotel, Washington, D. C.

Mar. 26–29: IRE Internat'l. Conv., Coliseum and Waldorf Astoria Hotel, New York, N. Y.

Mar. 28–29: 3rd Symp. on Engrg. Aspects of Magnetohydrodynamics, Univ. of Rochester, Rochester, N. Y.

Apr. 10–12: 2nd Symp. on the Plasma Sheath, New England Mutual Hall, Boston, Mass.

Apr. 11–13: SWIRECO (S. W. IRE Conf. and Electronics Show), Rice Hotel, Houston, Tex.

Apr. 24–26: Symp. on the Mathematical Theory of Automata, United Engrg. Ctr., New York, N. Y.

May 1–3: Spring Joint Computer Conference, Fairmont Hotel, San Francisco, Calif.

May 3–4: 1st International Congress on Human Factors in Electronics, Lafayette Hotel, Long Beach, Calif.

May 4–5: Bay Area Symp. on Reliability and Quality Control, U. S. Naval Postgraduate School, Monterey, Calif.

May 8–10: Electronic Components Conf. Marriott Twin Bridges Hotel, Washington, D. C.

May 14–16: NAECON, Dayton, Ohio.

May 14–16: Symp. on Thermionic Power Conversion, Antlers Hotel, Colorado Springs, Colo. (DL*: Feb. 1, 1962, Dr. V. C. Wilson, GE Co., Research Lab., Schenectady, N. Y.)

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Sarnoff to Address IRE 50th Anniversary Banquet

General David Sarnoff, Chairman of the Board of RCA, will be the principal speaker at the Golden Anniversary banquet during the IRE International Convention. The banquet will be held on March 28, 1962, in the Grand Ballroom of the Waldorf Astoria Hotel.

Participants in the special banquet program, commemorating the fiftieth anniversary of the founding of the IRE, will include Patrick E. Haggerty, IRE President and President of Texas Instruments Inc.; Lloyd V. Berkner, IRE Junior Past President and President of the Graduate Research Center; and Alfred N. Goldsmith, co-founder and Editor Emeritus of the IRE. Donald G. Fink, Vice President of Research of the Philco Corporation, will act as toastmaster.

During the banquet, President Haggerty will present the annual IRE Awards for 1962 to their recipients. The newly elected Fellows of the IRE will also be honored on this occasion, with Thomas F. Jones, Jr., Head of the School of Electrical Engineering, Purdue University, acting as spokesman for the Fellows.

General Sarnoff, Fellow and Life Member of the IRE, joined the Institute in 1912, the year of its formation, and played a leading role in its early activities as IRE Secretary (1915–1917) and Director (1915–1920). He was the first recipient of the IRE Founders Award, established in 1953.

In view of the large attendance expected, members are urged to make their banquet reservations immediately. Tickets may be ordered from IRE Headquarters, 1 East 79 Street, New York 21, N. Y., at $15.00 each.

Advisors Named For MIL-E-CON 1962

The Honorable Harold Brown, Director of Defense Research and Engineering, Department of Defense, heads the list of eminent civilians and military officers who will serve as advisors for the Sixth National Convention on Military Electronics, to be held in Washington, D. C., June 25–27, 1962. This annual meeting is sponsored by the IRE Professional Group on Military Electronics. Other advisors for MIL-E-CON are: Admiral G. Anderson, USN, Chief of Naval Operations; Honorable J. H. Rubel, Deputy Director, Defense Research and Engineering, Department of Defense; Dr. H. L. Dryden, Deputy Administrator, National Aeronautics and Space Administration; General B. A. Schriever, USAF, Commander, Headquarters, Air Force Systems Command; Lieutenant General A. G. Trudeau, USA, Chief of Research and Development, Department of the Army; Vice Admiral J. T. Hayward, USN, Deputy Chief of Naval Operations; H. Davis, Deputy for Research, Assistant Secretary of the Air Force (Research and Development); Major General F. L. Ankenbrandt, USAF (Ret), MIL-E-CON 1961 President, and Manager, Defense Products Assurance, Defense Electronic Products, RCA, Camden, N. J.; W. L. Doxey, Chairman PG on Military Electronics, and Director, Electronic Components Research Department, U. S. Army Signal Research and Development Labs., Fort Monmouth, N. J.; B. S. Melton, Chairman, Washington Section, IRE, and Chief, Systems Engineering Group, Air Force Technical Applications Center, Headquarters, USAF; J. Awramik, Jr., Electronics Scientist, U. S. Naval Research Laboratory.

J. J. Slattery, General Manager, The Martin Co., Baltimore, Md., is Chairman of the Technical Program Committee, which has set a deadline of February 1, 1962 for technical papers in the various fields of military electronics. L. D. Whitelock, 5614 Greentree Road, Bethesda 14, Md., is Exhibits Chairman.

More than 5700 engineers, scientists, and executives from industry, Government agencies and laboratories, the Armed Forces, and universities attended MIL-E-CON 1961. They listened to and discussed more than 100 technical papers and looked at exhibits of new developments in military electronics by 125 companies. This branch of electronics includes such fields of interest as: space technology, electronic propulsion, molecular electronics and systems, guidance and control, data handling systems, and operational analysis. Attendance has increased at each of the five MIL-E-CONs held, and is expected to show another increase at MIL-E-CON 1962.

Symposium on Thermionic Power Conversion

A Symposium on Thermionic Power Conversion, co-sponsored by the Technical Society and the Department of Defense, will be held at the Antlers Hotel, Colorado Springs, Colo. on May 14–16, 1962. Participating groups are the AIEE, IRE, American Nuclear Society, American Rocket Society, and the Electron Physics Division of the American Physical Society.

Because of the interdisciplinary nature of the technical problems associated with thermionic power conversion, there is need for communication among members of the serveral technical societies and between industrial, university, and military programs. This Symposium has been arranged in an effort to provide this communication efficiently. It will be an open technical meeting with the program supplied and conducted by the participating societies. Topics under discussion will be: Electrode Surface Processes, Transport Phenomena, Materials and Fabrication Techniques, Experimental Methods, and Device Construction and Performance. No registration fee will be charged, and no preregistration is required.

Call for Papers

1962 WESTERN ELECTRONICS SHOW AND CONVENTION

August 21–24, 1962

Los Angeles Memorial Sports Arena and Statler-Hilton Hotel, Los Angeles, Calif.

The following materials should be submitted by April 15, 1962:
100 to 200 word abstract, including title of paper, name and address of author.
500 to 1000 word summary of the paper.
Indication of the technical field in which the paper belongs (use IRE Professional Group classification).

Any necessary military or company clearance of papers must be granted before submission of the paper.

Address all materials to WESCON Business Office, c/o Technical Program Chairman, 1435 S. La Cienega Blvd., Los Angeles 35, Calif.

CALL FOR PAPERS FOR IFAC CONGRESS ON AUTOMATIC CONTROL

The second Congress of IFAC on Automatic Control will be held in Basle, Switzerland during September, 1963. Authors intending to present papers are requested to submit a 200 word abstract no later than March 1, 1962. Nearly all of the papers should deal with theory, or with applications of automatic control. A few papers will be adopted on components of control devices and on general items, and should be on one of the following topics: Theory: discrete systems, stochastic systems, optimal systems, learning systems, or systems reliability; Applications: process dynamics, computer studies of applications, on or off line, or optimizing or adaptive control applications; Components: new and effective devices, or measurement of the reliability of components.

The length of the paper should not exceed 20 typewritten pages, including drawings and tables. Four copies should be submitted no later than June 1, 1962, in English, French, Russian, or German. Authors are requested to inform their national member organization of IFAC of their intention to present a paper. This will assure their receiving all available information pertaining to the Congress. Correspondence relating to Applications of Automatic Control may be addressed to Professor I. Lefkowitz, Case Institute of Technology, 10900 Euclid Ave., Cleveland 6, Ohio. Information regarding papers in other areas of automatic control may be had by writing the chairman of the appropriate committee.

AIR FORCE MARS ANNOUNCES SCHEDULE

The Air Force MARS Eastern Technical Net broadcasting on 3295, 7540, and 15715 kc announces the following schedule of broadcasts on Sundays, 2–4 P.M. EST:

Feb. 4—"Advances in Broad Band Communications," K. Worthen, Communications Products Dept., General Electric Company.

Feb. 11—"Modern Electronics," D. M. Bray, Advanced Electronics Center, General Electric Company.

Feb. 18—"High Power Amplifiers Using Low-Cost Drift Power Transistors," C. F. Wheatley, RCA Laboratories.

Feb. 25—"Application of Rectifiers," R. Pollack, RCA Laboratories.

Mar. 4—"Tunnel Diodes; What They Are and What They Can Do," R. L. Watters, General Electric Research Laboratory.

Mar. 11—"Tunnel Diode Circuitry," E. Gottlieb, Semiconductor Products Dept., General Electric Company.

INTERNATIONAL EXHIBITION OF ELECTRONIC COMPONENTS

The twenty-fifth International Exhibition of Electronic Components will be held on February 16–20, 1962, at the Parc des Expositions, Porte de Versailles, Paris, France. This Exhibition, the oldest show specializing in the presentation of Electronic Products, has been a National French Exhibition since 1934. Since that time, the size and number of participants has more than doubled. It is sponsored by the Federation Nationale des Industries Électroniques, and organized by the Syndicat des Industries de Pieces Détachées et Accessoires Radioélectriques et Électroniques with the help of the following organizations: Syndicat des Industries de Tubes Électroniques et Semiconducteurs, Chambre Syndicale des Constructeurs de Compteurs, Syndicat des Constructeurs Français de Condensateurs Électriques Fixés, and Syndicat des Constructeurs de Relais Électriques.

Fifteen countries are represented in the Exhibition, as well as the Electronic Technical Press from various countries, the International Union of Radio Electronic Press, and the Syndicate of the French Radio Electrical Press. Since this is a specialized Exhibition, it is opened only to firms manufacturing the following materials: electronic spare pieces (capacitors, coils, transformers, switchings, safety devices, resistors, racks and chassis, aerials, relays, equipment and materials for electronic purposes), electronic tubes and semiconductor devices, and dial meters. General Administrator of the show is Robert Foucault.

Calendar of Coming Events and Author's Deadlines*

(Continued from page 14A)

May 22–24: Nat'l. Microwave Theory and Techniques Symp., Boulder, Colo.

May 22–24: Conf. on Self-Organizing Principles, Museum of Science and Industry, Chicago, Ill.

May 23–25: Nat'l. Telemetering Conf., Sheraton Park Hotel, Washington, D. C.

May 24–26: IRE 7th Region Conf. on "Space Communications," Olympic Hotel, Seattle, Wash.

June 18–19: Chicago Spring Conf. on Broadcast and Television Receivers, O'Hare Inn, Chicago, Ill. (DL*: Feb. 15, 1961, A. Cotsworth, Zenith Radio Corp., 6001 W. Dickens Ave., Chicago 39, Ill.)

June 25–27: 6th Nat'l. Convention on Military Electronics, Shoreham Hotel, Washington, D. C. (DL*: Feb. 15, 1962, J. J. Slattery, The Martin Co., Baltimore 3, Md.)

June 25–30: Symp. on Electromagnetic Theory and Antennas, Tech. Univ. of Denmark, Copenhagen.

June 27–29: Joint Automatic Control Conf., New York Univ., New York, N. Y.

June 27–28: 9th Annual Symp. on Computers and Data Processings, Elkhorn Lodge, Estes Park, Colo.

June 28–29: 4th Nat'l. Symp. on Radio Frequency Interference, Town House Hotel, San Francisco, Calif.

July 18–19: Data Acquisition and Processing in Medicine and Biology, Whipple Aud., Strong Mem. Hosp., Rochester, N. Y.

Aug. 14–16: Internat'l. Conf. on Precision Electromagnetic Measurements (former title: Conf. on Standards and Electronic Measurements), Nat'l. Bur. Standards, Boulder, Colo. (DL*: Mar. 15, 1962, Dr. G. Birnbaum, Hughes Res. Lab., Malibu, Calif.)

Aug. 21–24: WESCON (Western Electronics Show and Conf.) Los Angeles, Calif.

Aug. 27–Sept. 1: 2nd Internat'l. Federation of Information Processing Societies Congress, Munich, Germany.

Sept. 3–7: Internat'l. Symp. on Information Theory, Brussels, Belgium

Sept. 3–7: Nat'l. Advanced Technology, Management Conf., Seattle, Wash.

Sept. 3–7: 4th Internat'l. Congress on Microwave Tubes, Kurhaus Hotel, Scheveningen, Netherlands (DL*: Mar. 15, 1962, Congress Office, Postbox 62, Eindhoven, Netherlands.)

Sept. 13–14: 10th Annual Engineering Management Conf., Hotel Roosevelt, New Orleans, La.

Sept. 13–14: Nat'l. Symp. on Engineering Writing and Speech, Mayflower Hotel, Washington, D. C.

Sept. 19–20: 11th Annual Industrial Electronics Symp., Chicago, Ill.

Sept. 28–29: 12th Annual Broadcast Symp., Willard Hotel, Washington, D. C.

* DL = Deadline for submitting abstracts.

J. P. Eckert and Dr. J. Mauchly Receive John Scott Medal

J. Presper Eckert (S'41–A'47–SM'53–F'56) and Dr. John W. Mauchly (A'42–M'56–F'57) were awarded the John Scott Medal at the banquet of the annual Eastern Joint Computer Conference, held at the Sheraton Park Hotel, Washington, D. C., on December 13, 1961. They were presented with the award for their contribution to the welfare of mankind by the invention of the first electronic computer. The presentation was made by Donald C. Rubel, Vice President of the Philadelphia Board of Directors of City Trusts, administrators of the bequest under which the award is made. John Iliff, secretary of the Scott Award advisory committee, read the citation. A cash award of $1000.00, to be shared equally, also was presented to the inventors.

Mr. Eckert and Dr. Mauchly submitted their plan for the first electronic computer, ENIAC, in 1943. The project was undertaken during World War II, under a contract between the University of Pennsylvania and the U. S. Army Ordnance Department. Designed to compute firing tables, the computer was also put to work on other scientific problems. In 1950, they built UNIVAC, the first commercial electronic computer, which revolutionized business methods and started the computer industry on its way.



J. Presper Eckert (*left*) and Dr. John W. Mauchly (*right*), recipients of the John Scott Award presented at the annual Eastern Joint Computer Conference, Washington, D. C.

Mr. Eckert received the B.S.E.E. degree in 1941, and the M.S.E.E. degree in 1943, both from the Moore School of Electrical Engineering of the University of Pennsylvania. In 1946, he and Dr. Mauchly formed the Electronic Control Company, in Philadelphia, later incorporated as the Eckert-Mauchly Computer Corporation. In 1950, Remington Rand assumed control of the corporation, renaming it the UNIVAC Division. Mr. Eckert has served in this division as Head of commercial product engineering and technical consultant, and is currently Vice President and Special Assistant in the Office of the President, Remington Rand Division. He was honored for his pioneering work in electronic computer developments with presentation of the Howard N. Potts Medal from the Franklin Institute, Philadelphia, in 1949.

Dr. Mauchly studied electrical engineering and physics at Johns Hopkins University from 1925 to 1932. He joined the staff of the Physics Department in 1928 and received the Ph.D. degree in physics in 1930. From 1933 to 1941, he taught in the Physics Department of Ursinus College. After Eckert-Mauchly became a part of Remington Rand, he became Head of the UNIVAC Applications Research Center, a post which he held until the formation of Mauchly Associates. Dr. Mauchly has been past President of the Association for Computing Machinery, a trustee of the Society for Industrial and Applied Mathematics, and a member of the American Physical Society, American Geophysical Union, Institute of Mathematical Statistics, American Statistical Association, Phi Beta Kappa, and Sigma Xi. He was a recipient of the Howard N. Potts Medal from the Franklin Institute.

Colloquium on Microwave Communication

The Second Colloquium on Microwave Communication, sponsored by the Hungarian Academy of Sciences, department of Technical Sciences and the Scientific Society of Telecommunication, will be held in Budapest, June 5–8, 1962. The Colloquium will consider subjects concerning microwave communication in the following fields: systems analysis, microwave circuits and antennas, microwave electronics, electronic circuits, and systems measurements. G. Bognár, member of the Hungarian Academy of Sciences, is Chairman of the Organizing Committee.

Papers may be presented in English, German, Russian or Hungarian. A simultaneous translation via earphones will be arranged for the English, German and Russian languages. Those interested in participating are asked to communicate with the Secretary before March 31, 1962. Contributors are asked to enclose a summary of their proposed paper.

IRE Seventh Region Conference

The Seattle Section of the IRE will host the IRE Seventh Region Conference on May 24–26, 1962, in Seattle, Wash. The theme of the Conference will be "Space Communications," stressing the impact of electronics on the cultural and social life of the future. Because of the international importance of this Conference, and to celebrate the 50th Anniversary of the founding of the professional society, the Board of Directors of the IRE will hold their annual meeting during the Conference. In honor of the occasion, the Seattle Worlds Fair will designate Friday, May 25, 1962 as "Electronics Day."

Call for Papers

Special Issue of Proceedings of the IRE on Quantum Electronics

A special issue of the Proceedings of the IRE will be devoted to Quantum Electronics, with emphasis on Optical and Ultramicrowave Masers (Molecular Amplifiers operating by Stimulated Emission of Radiation). This volume will be published in the Winter of 1962. The undersigned has been asked to coordinate this issue and endeavor to obtain outstanding papers in the field for incorporation. Tentatively, it has been decided to collect papers in the following categories:

A. Broad survey papers in quantum electronics
B. Original Papers on
   1) Optical masers
   2) Optical pumping
   3) Studies of molecular and atomic transitions in gases, liquids, and solids which bear on maser application
   4) Wave-matter interaction phenomena
   5) Resonant structures for ultramicrowave and shorter waves
   6) Phonon interactions as related to maser phenomena
   7) Studies of semiconductors as applied to coherent electromagnetic wave generation
   8) Information and noise theory for maser systems
   9) New accomplishments in the microwave region using maser systems
   10) New ideas and accomplishments in the broad field of quantum electronics.

Manuscripts should be submitted to the undersigned as soon as they are available. The tentative deadline for the complete manuscript is July 1, 1962. A time extension will be granted for work in progress where a suitable detailed abstract is submitted. The refereeing of papers will be facilitated if manuscripts are sent in early. Three copies of the paper should be submitted, including one original typed copy with one set of reproducible illustrations (and two copies of all illustrations), and a photograph and biography of the author. Further inquiries may also be directed to Mr. E. K. Gannett, Managing Editor, Proceedings of the IRE, One East 79 Street, New York 21, N. Y.

Professor J. R. Singer
Editor, IRE Quantum Electronics Issue
Cory Hall
University of California
Berkeley 4, Calif.



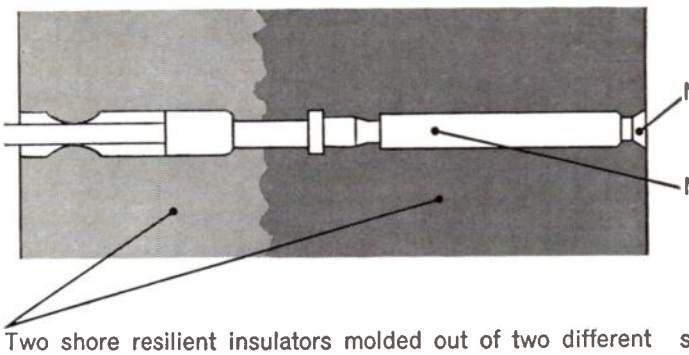
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NAECON Plans Announced

Raymond J. Nordlund, President of NAECON, has announced plans for the Fourteenth Annual National Aerospace Electronics Conference to be held at the Dayton Biltmore Hotel and the Memorial Hall May 14–16, 1962. NAECON is jointly sponsored by the Dayton Section of the IRE and the Professional Group on Aerospace and Navigational Electronics.

The technical papers program, headed by George Langston, will be built around the Conference theme, "Electronics-Aero/Planetary Science." Some of the planned subjects are: New Cryogenic Electronics Techniques, Application of Coherent Optical Radiation Devices, Coherent Optical Radiation Theory and Techniques, Biological Electronics, Electronic Simulation of Living Systems, Computer Technology in Engineering Design, Aerospace Medical Electronics, Inertial Technology for Aerospace Navigation, Radio Propagation Phenomena, New FM Stereo Techniques, and Millimeter Wave Technology. The Institute of Aerospace Sciences is again participating in the technical program. W. Donald Dodd is preparing four technical sessions on Materials for Space, Magnetohydrodynamics, Electronic Propulsion, and Reentry Energy Management.

Last year's innovation, the classified session, contributed by the Aeronautical Systems Division of AFSC, will be repeated this year. It will be held in the ASD Auditorium at Wright-Patterson Air Force Base on Wednesday, May 16th, from 1:00 P.M. to 4:30 P.M. Free transportation will be provided from and to the Biltmore Hotel. Attendance will be limited to the first 500 who register at the special registration desk in the NAECON registration room on May 16. Security clearance forms will be included with the NAECON Advance Brochures and must be mailed to the indicated address before April 25, 1962. Additional forms will be available from the NAECON Office, 1414 E. Third Street, Dayton, Ohio. The classified program will include such papers as "Survey of the USAF Limited War Program," "The Aerospace Plane," "Navigation and Guidance," "Control of Aerospace Vehicles," etc.

Exhibits will be displayed in Memorial Hall before the Air Force Systems Command, the Air Force Logistic Command, and the participating Aerospace and Electronic Industries. Those interested should contact Robert J. Stein, 19 Beverly Place, Dayton 19, Ohio.

Symposium on Computers And Data Processing

The Denver Research Institute of the University of Denver will hold its Ninth Annual Symposium on Computers and Data Processing at the Elkhorn Lodge in Estes Park, Colo., on June 27 and 28, 1962. The continuing theme of this series of meetings has been the advanced treatment of basic problems in computer technology. Papers will be presented in the fields of Components, Logic Design, Philosophy of Computer Design, and Inter-Relation of Analog and Digital Techniques. The fourth topic last year was Computers and Education. Session IV this year will continue this policy of having one session which is, in effect a probe into a growing area. Such subjects as simulation of analog functions on digital equipment, digital control of analog operations, and analog to digital conversion will be appropriate.

Although it is anticipated that the program will be comprised largely of invited papers, a limited number will be selected from papers submitted without invitation. Authors wishing to submit papers may send abstracts of approximately 150 words, no later than March 1, 1962, to: W. H. Eichelberger, Denver Research institute, University of Denver, Denver 10, Colo.

NEC Elects Officers

James H. Kogen, Chief Engineer of GPE Controls, Inc., Chicago, Ill., has been elected President of the National Electronics Conference, Inc., for 1962. Past Vice President of NEC, he succeeds J. J. Gershon as President. He has been particularly active in AIEE affairs, representing the AIEE on the NEC Board of Directors, and is also a member of the Instrument Association of America, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi, and is a registered professional engineer.

Other officers named for the 1962 NEC, which will be held at McCormick Place, Chicago, on October 8–10, 1962, are: J. S. Powers, Kenneth Meyers Co., Vice President; Dr. E. W. Ernst, University of Illinois; Secretary; Dr. J. S. Aagaard, Northwestern University, Treasurer; C. F. Valach, Amphenol Borg Electronics Corp., Assistant Treasurer Elected as Chairman of the Board is V. H. Disney, Director of Electronic Research, Armour Research Foundation. Mr. Disney is a past President of NEC and represents the Illinois Institute of Technology on the NEC Board of Directors.

More than 15,000 engineers are expected to attend the 18th annual Conference, and more than 500 electronic firms will be exhibiting their products. A record attendance and number of exhibitors is expected. The National Electronics Conference, Inc. is a non-profit organization with headquarters in Chicago, serving as a national forum for the presentation of authoritative technical papers on electronic research, development and application. Sponsors include the IRE, AIEE, Illinois Institute of Technology, Northwestern University, and the University of Illinois. Participants are Electronic Industries Association, Electronic Representatives Association, 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.



Members of the Fourteenth Annual NAECON Committee:
Sitting left to right: F. Kuppin (Registration); L. Brewer (Secretary); R. J. Nordlung (President); M. Cushing (Ladies Committee); M. Roberts (Secretarial); E. Malogorski (Printing). *Standing left to right:* W. C. Eppers, Jr. (Dayton Section); J. J. Magill (Arrangements); A. G. Bissonette (Publicity); W. J. Cushing (Finance); J. L. Dennis (Executive Vice President); G. A. Langston (Papers); T. Rynda (Proceedings); W. D. Dodd (Institute of Aerospace Sciences); R. J. Stein (Exhibits); C. Ashley (Arrangements).

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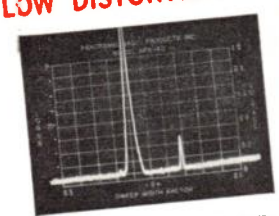
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7- 18.0 — 26.4 KMC	— 70 to — 90 dbm
8- 26.4 — 44.0 KMC	— 60 to — 85 dbm

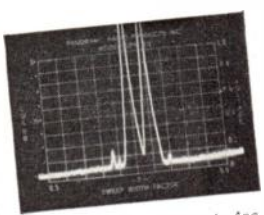
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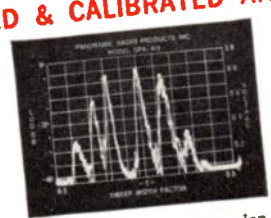
Extended dynamic range comparison of 2 signals on SPA-4a. Larger is + 15 db over full scale log. Smaller is at —28 db on scale or —43 db from larger. Note exceptional freedom from spurious. (Photo not retouched)



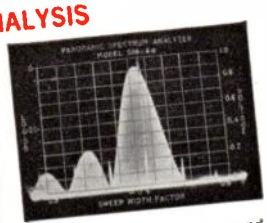
Distortion analysis illustrates SPA-4a wide range linearity. Odd-order distortion here is measured more than 50 db below level of 2 main tones (deflected 20 db above full scale). Photo unretouched.

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Narrow band 20 kc dispersion analysis shows unique resolution capability. Here, a 1000 mc FM signal with 2 kc modulation is seen near first carrier null. Photo unretouched.



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PROFESSIONAL GROUP NEWS

At its meeting on November 14, 1961, the IRE Executive Committee approved the following new Chapters: PG on **Antennas and Propagation** and **Microwave Theory and Techniques** (Joint Chapter)—Toronto Chapter; PG on **Instrumentation**—Philadelphia Chapter; PG on **Space Electronics and Telemetry**—Denver Chapter.

On December 12, 1961, the IRE Executive Committee also approved the following new chapters: PG on **Bio-Medical Electronic**—Milwaukee Chapter; PG on **Reliability and Quality Control**—Binghamton Chapter.

NEREM REPORT AVAILABLE

A record breaking attendance of over 17,000 registered at the Northeast Electronics Research and Engineering Meeting (NEREM '61). One of the features of the annual meeting, held this year at the Somerset Hotel and Commonwealth Armory, Boston, Mass. was the distribution of a letterpress-printed conference report—the *NEREM Record*—with illustrated condensations of the 95 papers. Copies are now available from the Boston Section of the IRE, 313 Washington St., Newton 58, Mass., at $7.50 per copy.

CALL FOR PAPERS
IRE SYMPOSIUM ON
RADIO FREQUENCY INTERFERENCE

Authors are invited to submit summaries of technical papers for the Fourth National Symposium sponsored by the IRE Professional Group on Radio Frequency Interference, to be held on June 28–29, 1962, at the Del Webb Towne House, San Francisco, Calif. The theme of the Symposium, which will be conducted on an unclassified basis, is "Design—The Answer to RFI." It is hoped that the emphasis on design will serve the following purposes: establish some requirements for interference-free design at the circuit, equipment, and system levels; bring to the attention of designers the importance of freedom from interference as a design criterion, and stress its importance in preliminary planning and design; bring new applications in communication and control to the attention of working engineers in the field of RFI, so that some new problems can be anticipated; focus attention on new components and materials and encourage their use in eliminating interference; and help in making useful the results of current efforts in interference prediction and data collection such as the DOD Radio Frequency Compatability Program. Time will also be allowed for papers in other areas within the field of particular current interest.

Summaries should be 500 words in length and should be furnished in triplicate with the title of the proposed paper, and the author's name and corporate or government affiliation no later than February 16, 1962. Summaries should be sent to R. G. Davis, Technical Program Chairman, Department 58-25, Lockheed Missiles and Space Co., Sunnyvale, Calif.

OBITUARIES

Dr. Albert Hoyt Taylor (M'16–F'20–L'51), a pioneer in the development of radar and a past President of the IRE (1929), died on December 11, 1961. Born on January 1, 1879, he received the B.S. degree from Northwestern University in 1902, and the Ph.D. degree from the University of Göttingen, Germany, in 1908. He was an Instructor at Michigan State College (1900) and at the University of Wisconsin (1903), and Professor and Head of the Physics Department at the University of South Dakota (1903–1907).



A. H. TAYLOR

Dr. Taylor was a Lieutenant in the Naval Researve in 1917. In 1918 he was promoted to the rank of Lieutenant Commander, in charge of the aircraft radio laboratory. He was appointed Commander in 1919, remaining in active service until 1922. Upon the organization of the Naval Research Laboratory, he was made superintendent of its radio division. In this capacity, he directed the American development of radar and installed the first combat radar unit on the battleship *New York* in 1938. At the end of World War II, in recognition of his contributions in the field of radar, he received the Medal of Merit and was hailed as a "father of radar." In October, 1959, he received the Stuart Ballantine Medal from the Franklin Institute, for more than fifty years of scientific work. In 1927, he received the IRE Morris Liebmann Memorial Prize. He was also the recipient of the IRE Medal of Honor, the John Scott Medal, and the Pioneer Award of the Professional Group on Aeronautical and Navigational Electronics.

Dr. Taylor was a fellow of the AIEE and the American Physical Society. He was the author of thirty five papers and held fifty four patents.

Dr. Ray H. Manson (M'23–F'30–L'58), former President and Chairman of the Board of Stromberg-Carlson Company, died on December 23, 1961. He was born in Bath, Maine, on August 25, 1877, and was graduated from the University of Maine in 1898. He was awarded the degree of Doctor of Engineering in 1933. Before joining the staff of Stromberg-Carlson, he was associated with the Western Electric Company and the Kellogg Switchboard and Supply Company, both in Chicago, and the Dean Electric Company and Garford Manufacturing Company in Elyria, Ohio.



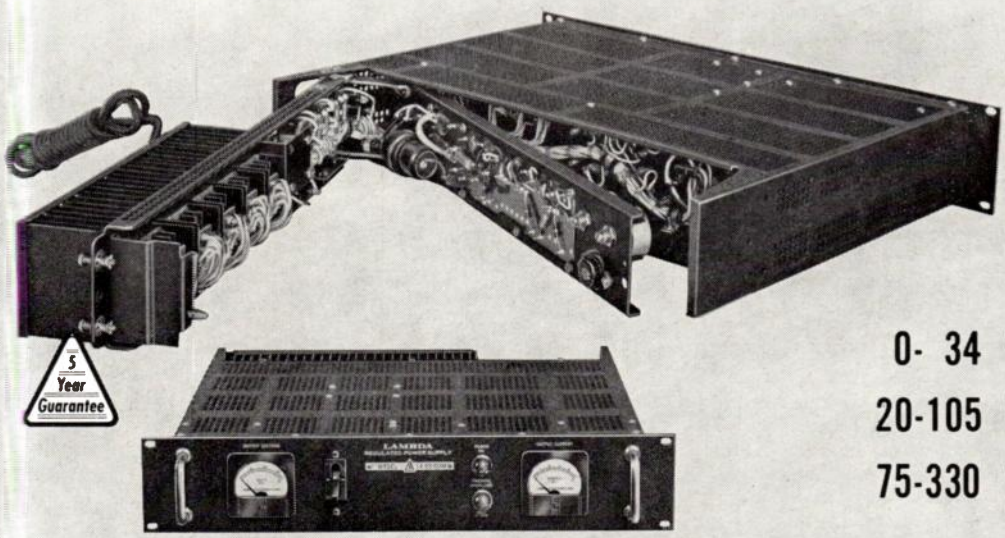
DR. R. H. MANSON

In 1920, he was elected a Director of Stromberg-Carlson, becoming Vice President in charge of engineering in 1924, and General Manager in 1940. In 1945, he was appointed President. One of the nation's leading authorities in the field of communications, and creator of more than 100 inventions in this field, Dr. Manson was President of the IRE in 1931. He was a Director of the Radio Manufacturers Association, an executive committee member of the association's set division, and representative of that organization of the administrative committee of the Radio Technical Planning Board. He was also a Fellow of the American Institute of Electrical Engineers, a Fellow of the Radio Club of America, and a member of the Acoustical Society of America, the Rochester Engineering Society, and the American branch of the Newcomen Society of England.

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Industrial Engineering Notes*

Industry Sales Pass $10 Billion

The electronics industry will make history by the end of 1961 by passing the $10 billion mark in total factory sales, EIA President L. Berkley Davis disclosed at a press conference preceding the Winter Conference in Los Angeles last week. The figure for the industry at the close of business would be $10.15 billion, a new all-time high, he said. He forecast $10.8 billion total factory sales during 1962. Electronics' rise to $10.15 billion from 1960's $9.75 billion was achieved through continued expansion of military and industrial markets, Mr. Davis said. Dollar sales of consumer products declined somewhat, he observed, partially because of trends toward purchases of smaller models in some areas, although consumers are shifting toward quality buying in the phonograph field. "The uptrend of the national economy during the last half of this year is expected to continue, thereby providing healthful conditions for further expansion of our industry during 1962," Mr. Davis said. "It is reasonable to assume that a larger gain in total electronics output will be forthcoming next year, stimulated by the need for more advanced weapon systems, expanding space and missile programs, a continuing rise in plant modernization and automation, and increasing consumer spendable incomes." In the industrial electronics field, Mr. Davis said, computing and data processing systems, testing equipment, microwave apparatus, and navigational aids are continuing to enlarge in sales as a result of more sophisticated ideas and techniques being applied to industrial and business processes. "Increases generally in industrial electronic sales are anticipated during 1962 and are expected to continue throughout the next decade," Mr. Davis said. "More capital investments in plant modernization, increasing defense production, and efforts by military equipment manufacturers to diversify into the commercial fields are factors strongly stimulating growth of this sector." In discussing salient aspects of the military market with total 1961 factory sales estimated at $5.3 billion, Mr. Davis called attention to the ever-accelerating rate of technological change. He noted that total expenditures for research and development and for new weapons have risen faster than the military budget as a whole and said the trend is expected to continue. "Despite the increasing emphasis on missiles and satellite systems, military aircraft will continue to remain the largest procurement item in the defense product mix, at least until 1964," Mr. Davis said. "Thereafter, missiles and space systems will dominate the procurement category." Total dollar sales of components during 1961 are expected to match or move slightly above the 1960 figure of $3 billion, Mr. Davis stated. While receiving and TV picture tube sales are somewhat below last year, he noted that increases are being recorded for such components as capacitors, fixed resistors, relays, and transformers. "Although unit sales of transistors and diode-rectifiers are up appreciably from last year, price declines in these maturing sectors of the semiconductor industry will probably prevent the total dollar volume from surpassing the $525 million high-level mark established in 1960," Mr. Davis said. "However, if the newer emerging semiconductor devices, such as solar cells, infrared sensors, thermoelectric, and similar components are considered, then the over-all semiconductor industry is showing continuing growth."

Association Activities

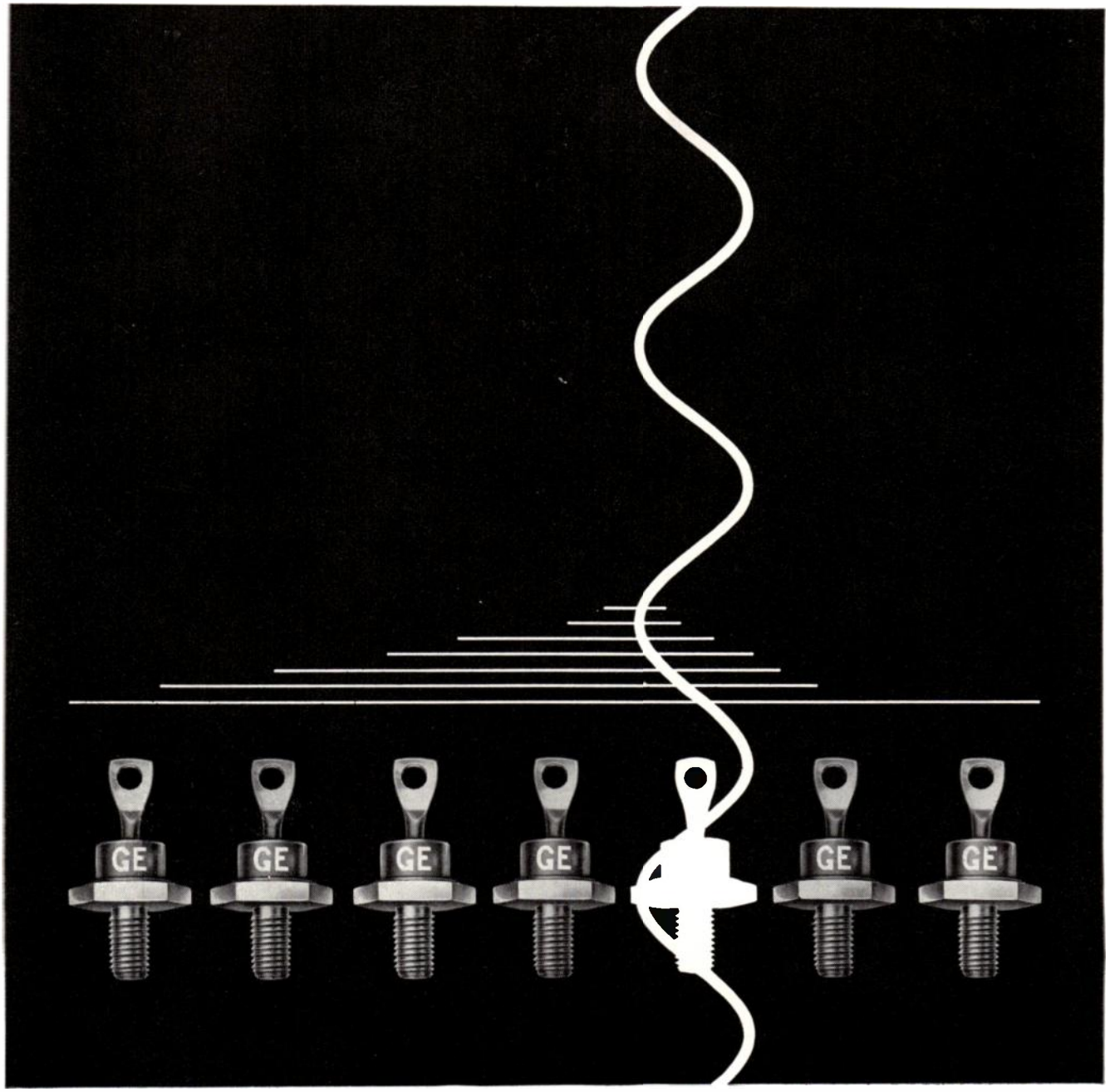
A comprehensive bibliography of publications and articles on value engineering and value analysis has been prepared by a task force of EIA's Value Engineering Committee headed by Fred S. Sherwin (Raytheon Co.). Listed are titles of nearly 250 publications and articles in the field, along with the authors or originating firms and a brief description of what each covers. A limited number of copies of the bibliography is available to representatives of EIA member-companies from the Office of Information, EIA Headquarters. Ask for "Value Engineering and Value Analysis: A Bibliography."

Governmental and Legislative

The National Science Foundation last week released the proceedings of an NSF-sponsored conference held at Princeton University in November 1960 on academic and industrial basic research. Purpose of the conference was to bring into closer association the scientific interests of government, universities, and industry. The first of its kind sponsored by NSF, the conference drew upon the faculty at Princeton and the industrial and other academic interests nearby. Papers were presented on four major themes: The role of government, industry, and the university with respect to basic research; the interdependence of academic and industrial basic research; managerial and personnel problems in research laboratories; and

(Continued on page 24A)

* The data on which these Notes are based were selected by permission from *Weekly Reports*, issues of November 27, December 3 and 18, 1961, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.



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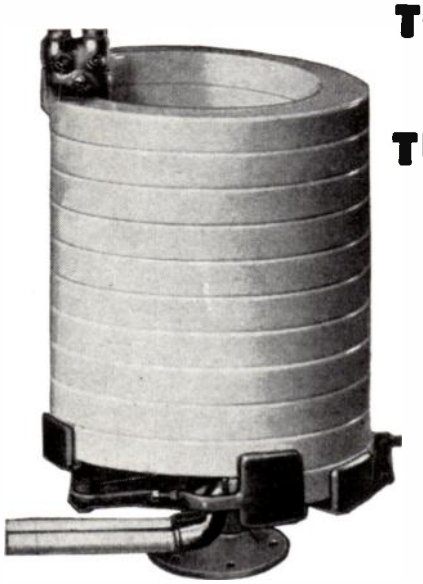
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Lapp

Industrial Engineering Notes

(Continued from page 22A)

examples of industrial experience in basic research. Copies of the "Proceedings of a Conference on Academic and Industrial Basic Research" may be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., for 55 cents.

Industry Marketing Data

A record monthly total of 19,386,202 transistors sold at factory outlets during September sent the number of units sold during the first nine months of this year surging ahead of the total for the entire 12 months of 1960, according to month-end statistics released last week by the EIA Marketing Data Department. Through September, 136,490,332 transistors were sold with a total value of $227,002,035, compared with 90,263,352 valued at $222,198,961 sold during the first nine months of 1960. The year-end totals last year were 127,928,586 units worth $301,432,285. . . . **Despite a dip in output of television and radio receivers, October was the second highest production month for both types of consumer products, ranking only behind September.** Monthly totals released last week by the EIA Marketing Data Department showed 620,815 TV sets produced in October, against 694,580 during the record month of September. Total TV output during the 10 months of 1961 was 5,014,583, just ahead of the 4,873,120 cumulative total recorded for the same period in 1960. October radio production stood at 1,796,391, compared with the 2,048,698 peak reached in September. Cumulative output of radio sets this year through October remained behind the total for last year. The year-to-date totals were 13,797,879 this year against 14,135,937 in 1960.

Section Meetings

ALAMOGORDO-HOLLOMAN

"Parametric Amplifiers and Up-Grading Radar Performance," J. Saloom, SFD Laboratories; 11/21/61.

ALBUQUERQUE-LOS ALAMOS

"A New Concept in Analog Computers," M. Gilliland, Beckman Instruments; 11/14/61.

ATLANTA

"The Attraction of Scientifically Oriented Industry to Georgia," J. F. Byrne, Motorola Corp.; 11/17/61.

"Proposed Industrial Research and Development Center for Georgia," K. Wagner, Georgia Tech.; 12/14/61.

BALTIMORE

"Application of Airborne Television to Public Education," F. G. Mullins, Westinghouse Electric Corp., Joint IRE-AIEE; 11/13/61.

(Continued on page 26A)



Model 440
$250

Widest range–highest stability!

Improved NARDA Solid-State POWER METER –portable unit with rechargeable battery– measures rf power (0.1 to 10 mw) within 3%!

Compensated or not—you can't buy a more stable Power Meter for lab use than the new, improved version of Narda's Model 440! Self-balancing, it provides accurate direct reading measurements of either CW or pulsed power. Fully transistorized and powered by a nickel-cadmium battery (rechargeable overnight), it provides complete freedom from deviations due to line voltage fluctuations. Moreover, zero-set controls have been especially re-designed for fine resolution and ease of control on lowest scales.

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Section Meetings

(Continued from page 24A)

"Some Studies in Machine Learning Using the Game of Checkers," A. L. Samuel, IBM Corp., T. J. Watson Research Center; 12/11/61.

BAY OF QUINTE

"Semiconductor Radiation Detectors," H. Pullan, RCA Corp; 11/15/61.

BEAUMONT-PORT ARTHUR

"VHF Vehicular Communication," D. Land, Motorola, Inc.; 11/28/61.

"Telemetering and Control Systems," J. T. Hargett, Texas Instruments; 12/12/61.

BENELUX

"Simofoon," G. M. Uitermark, Netherlands P.T.T.; 11/29/61.

BOSTON

"Dynamic Behavior of R. & D. Organizations, Management by Feedback Control Techniques," R. B. Wilcox & Panel, Joint PGAC-PGEM-AIEE; 10/24/61.

BUFFALO-NIAGARA

"Fundamentals of Reliability Prediction," H. McIntyre and R. Costello, Sylvania Engineering Labs.; 11/15/61.

"Miniaturization of Physiologic Equipment," N. W. Greatbatch, N. W. Greatbatch Inc.; 12/13/61.

CENTRAL FLORIDA

"Orbiting Astronomical Observatory," W. C. Coulbourn and J. Russell, Grumman Aircraft Engr. Corp.; 10/19/61.

"Engineering Today," Dr. L. V. Berkner, President IRE; 11/28/61.

CENTRAL PENNSYLVANIA

"Radio Propagation Folklore," Mrs. M. Lindeman Phillips, Consultant; 9/19/61.

"Modern Developments in Oscilloscopy," T. Smith, Tektronix, Inc., Joint with AIEE, E. F. Society; 10/19/61.

"The Electrical Two-Speed Synchro," L. S. Kahn, Eclipse Pioneer Div., Bendix, Joint with AIEE; 11/21/61.

CHINA LAKE

"High Speed Digital Computer Usage at NOTS," J. Zaharias, USNOTS; 9/7/61.

"A Direct Plotter for Frequency Analog Data," F. Pierce, U. S. Naval Ordnance Test Station; 10/19/61.

CHICAGO

"The Engineer—Threat to Society?" E. J. Tangerman, McGraw-Hill Co., Inc.; 12/8/61.

CINCINNATI

"The Dual Channel Transistor Power Amplifier," A. Bereskin, University of Cincinnati; 11/21/61.

COLUMBUS

Panel meeting, Discussion of possible IRE and AIEE merger. M. Thurston, Ohio State Univ., M. E. Boone, Ohio State Univ., H. R. Chope, Ind. Nucleonics, K. E. Cochran, Battelle Memorial Inst., P. L. Cosgriff, Ohio State Univ., R. H. Gulcher, North American Aviation, C. A. Levis, Ohio State Univ., G. Tamplin, Battelle Memorial Inst., F. C. Weimer, Ohio State Univ.; 11/28/61.

DALLAS

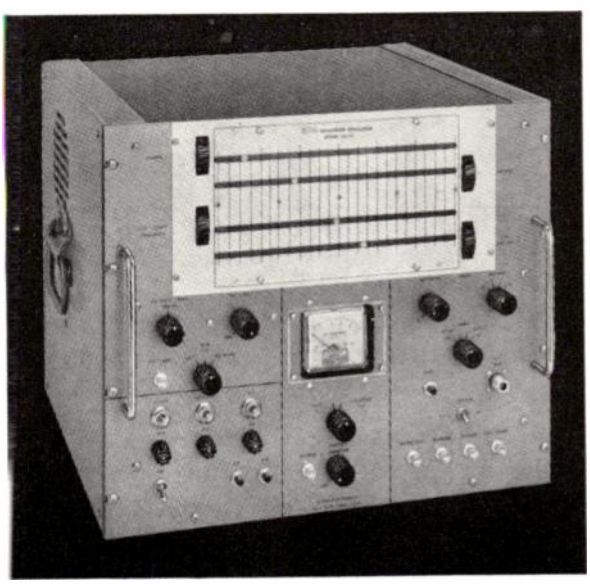
"How To Build a Brain," J. E. Steele, Aeronautical Systems Div., WPAFB; 11/27/61.

(Continued on page 28A)

Microwave Sweep Oscillators

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Alfred Model 623BK Microwave Oscillator, featuring built-in leveler, drift less than ±.02% per hour, adjustable frequency markers and Quick Look Readout.

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Section Meetings

(Continued from page 26A)

DAYTON

"Skybolt Guidance Subsystem Management," R. F. Miller, Nortronics Div., Joint with PGEM Dayton Chapter; 9/6/61.

"The Ruby Laser," M. R. Curre, Hughes Research Labs.; 10/4/61.

"History and Development of the Space Detection and Tracking System," W. E. Kelly, 496 L System Program office, Hanscom Field; 11/1/61.

EL PASO

"The El Paso Electric Co. Microwave System," J. R. Stuart, El Paso Electric Co., Tour of El Paso Electric Co.; 10/26/61.

"The Language of Computers," J. R. Young, New Mexico State University; 11/30/61.

EMPORIUM

"Space," J. E. Steele, U.S. Air Force; 11/21/61.

ERIE

"Application and Selection of Semiconductor Devices," D. H. Lynch, Westinghouse Electric Co.; 11/21/61.

"Project Fastest," D. J. Puyalie, Erie Electronics Div. Erie Resistor Corp., Joint with Presque Isle Section, Instrument Society of America; 11/28/61.

FORT WAYNE

"Fundamentals of Reliability and Applications," A. O. Plait, Magnavox Co.; 11/30/61. 1961 Engineers Holiday Ball; 12/2/61.

FORT WORTH

"Solion Devices," J. Cox, Texas Research & Electronic Corp.; 11/14/61.

HAMILTON

"Theory and Design of Silicone Power Rectifiers," A. J. Carlan, Syntron Co.; 12/4/61.

INDIANAPOLIS

"Electrical Optics in Semiconductor Devices," A. Middleton, Harshaw Chemical Co.; 11/30/61.

KANSAS CITY

"Symposium—Electronic Systems Reliability," H. T. Gruber, Battelle Memorial Institute, A. J. Finocchi, I.T.T. Labs., W. E. Boyes, Sandia Corp., G. W. Lindsay, Wright-Patterson Air Force Base, A. L. Floyd, Hoffan Elect. Corp.; 11/14/61.

KITCHENER-WATERLOO

"The Theory of Ultrasonic Cleaning," M. A. Wilson, Canadian Westinghouse Corp.; 11/20/61.

LITTLE ROCK

"The World's Most Powerful Transmitter, 2,000,000 Watts VLF, at U. S. Naval Radio Station, Cutler, Maine," M. W. Bullock, Continental Elec. Manufacturing Co.; 10/16/61.

"Microwave Communications Problems," F. V. Long, Texas Eastern Transmission Co.; 11/13/61.

LONDON

"Design Considerations of Satellite-Borne T. V. Systems," D. J. Ritchie, Bendix Corp.; 11/13/61.

"Satellite Communications," D. J. McDonald, Bell Telephone Co. of Canada; 12/11/61.

LOS ANGELES

"Some Potentialities of Optical Masers," B. M. Oliver, Hewlett-Packard Co., Meeting hosted by Pasadena IRE; 11/16/61.

(Continued on page 30A)

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amplifier gain either 100 or 1000, adjustable to precise values.

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Frequency of high cutoff point	Maximum noise, microvolts RMS referred to input	
	Normal (10 meg impedance)	Low Noise (100 k impedance)
100 kc	3.0	1.9
30 kc	1.9	1.1
10 kc	1.4	0.8
3 kc	0.9	0.6
1 kc	0.7	0.4
300 cps	0.5	0.3
100 cps	0.4	0.25

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Section Meetings

(Continued from page 28A)

"Extra-Sensory Perception and the Engineering World," J. B. Rhine, Parapsychology Lab., Duke University; 12/7/61.

LUBBOCK

"The Impact of Semiconducting Devices on the Problem of Network Synthesis," R. H. Secat, Texas Technological College; 10/24/61.

"Use of IBM 1401 Computer for Accounting," C. Kight, South Western Bell Telephone Co.; 11/28/61.

MIAMI

"Computers: Application and Design," P. Collins, IBM Corp.; 10/25/61.

"The History and Development of Underwater Detection," H. A. Cook and A. A. Elwood, Airpax Electronics, Inc.; 11/15/61.

MOBILE

"Space Communications," R. E. Gudgen, Southern Bell Tel. & Tel., Installation of officers for 1961–1962 term; 10/27/61.

NEW ORLEANS

"Ramifications of Electronics in Missile Production," V. E. Haywood, Avondale Shipyards; 11/17/61.

NORTH CAROLINA

"Present Status of Solid State Development and Devices," M. L. Embree, Bell Tel. Labs.; 11/17/61.

NORTHERN NEW JERSEY

"Optical Masers, Their History and Future," W. R. Bennet, Jr., Bell Tel. Labs.; 11/8/61.

OMAHA-LINCOLN

"Microwave Radiation Hazards," W. W. Mumford, Bell Tel. Labs.; Joint with Neb. Section AIEE and U. Of Neb. joint student branch; 11/17/61.

OTTAWA

"RCAF Tacan Air Navigational System," R. R. Sly, and J. Coffell, RCAF; "Experimental Results on Fading of Jamming Signals in L & S. Frequency Bands," J. M. MacDonald, RCAF; "The Design of Related Parts for Modern Air Weapons," K. Thomassen, RCAF; Sponsored by PGMIL Chapter; 12/7/61.

PHILADELPHIA

"MIA--The Metal Interface Amplifier," J. M. Spratt, Philco Research Center; "MIA As a Circuit Component," S. K. Ghandi, Philco Research Center; 11/8/61.

PITTSBURGH

"Status of Standards in Physical Measurements," A. V. Astin, National Bureau of Standards; 11/16/61.

Tour of the Pitt Acoustic Lab. K. C. Stewart, Univ. of Pittsburgh; 12/4/61.

PRINCETON

"Optical Properties of the Helium-Neon Maser," D. R. Herriott, Bell Tel. Labs.; 11/14/61.

QUEBEC

"Electronics: Its Use in the Study of the Nervous System," G. Lamarche, Laval Univ.; 11/14/61.

ROCHESTER

1961 Conference on Data Acquisition and Processing in Biology and Medicine; 7/13/61.

(Continued on page 32A)



An NEC semiconductor device for every circuit application

The range of semiconductor products at NEC is perhaps the widest of any manufacturer. Entertainment or industrial, standard broadcast frequencies to microwave — NEC has a semiconductor device with the ratings and characteristics for your application. There's a wide selection of current ratings, operating temperatures, and mountings.

These are produced in NEC's new semiconductor plant where crystal surfaces are cleaned by 100 tons of hyper pure water daily. Every seal is tested in krypton isotopes, providing a failure rate suitable for the most critical applications.

A single source for semiconductors can mean a saving in time and costs. Just let NEC know your requirements and full technical data will be sent.

Types of NEC semiconductor devices

• PNP super-grown Germanium transistors • Germanium photo transistors • PNP alloy junction Germanium transistors • Silicon Mesa transistors • Germanium Mesa transistors • Germanium gold-bonded diodes • Silicon rectifiers • Silicon controlled rectifiers • Silicon capacitors • Zener diodes • Germanium point-contact mini-diodes • Silver-bonded diodes • Microwave mixer diodes • Silicon junction mini-diodes

Reliability The first NEC transistorized carrier telephone system was an NT & T installation in 1958 between Toyama and Takaoka, a distance of 15 miles. It consists of two terminal stations and a repeater station with 240 channels using 1,600 transistors.
During last 14,000 hours of operation transistor failure has caused only two channel faults. This corresponds to a failure rate of 0.009% per 1,000 hours.

Communications Systems / Electronic Components NEC *Nippon Electric Co., Ltd.* Tokyo, Japan

IS *YOUR* COMPANY ON THE *OFFENSE* FOR *DEFENSE?*

SIGNAL is your introduction to the men who control government's growing $6.5 billion radio-electronics spending

Never before have our armed forces so badly needed the thinking and products of the electronics industry. Advertising in SIGNAL, the official journal of the *Armed Forces Communications and Electronics Association*, puts you in touch with almost 11,000 of the most successful men in the field—every one a prospect for your defense products!

Share in the defense and the profits! Company membership in the AFCEA, with SIGNAL as your spokesman, puts you in touch with government decision-makers!

SIGNAL serves liaison duty between the armed forces and industry. It informs manufacturers about the latest government projects and military needs, while it lets armed forces buyers know what *you* have to offer to contribute to our armed might. SIGNAL coordinates needs with available products and makes developments possible.

But SIGNAL is more than just a magazine. It's *part of an over-all plan!*

A concerted *offensive* to let the government, which has great faith in industry and the private individual producer, know exactly what's available to launch its far-sighted plans. Part of this offensive is the giant AFCEA National Convention and Exhibit (to be held next year in Washington, D.C., June 12-14). Here, you can *show* what you have to contribute directly to the important buyers. Your sales team meets fellow manufacturers and military purchasers and keeps "on top" of current government needs and market news.

Besides *advertising* in SIGNAL which affords year-round exposure by focusing your firm and products directly on the proper market . . . besides *participation* in the huge AFCEA National Convention and Exhibit . . . the over-all plan of company membership in the AFCEA *gives your firm a highly influential organization's experience and prestige to draw upon.*

As a member, you join some 175 group members who feel the chances of winning million dollar contracts are worth the relatively low investment of time and money. On a local basis, you organize your team (9 of your top men with you as manager and team captain), attend monthly chapter meetings and dinners, meet defense buyers, procurement agents and sub-contractors. Like the other 59 local chapters of the AFCEA, your team gets to know the "right" people.

In effect, company membership in the AFCEA is a "three-barrelled" offensive aimed at putting your company in the "elite" group of government contractors —the group that, for example in 1957, for less than $8,000 (for the full AFCEA plan) made an amazing total of *459.7 million dollars!*

This "three-barrelled" offensive consists of

(1) Concentrated advertising coverage in SIGNAL, the official publication of the AFCEA;

(2) Group membership in the AFCEA, a select organization specializing in all aspects of production and sales in our growing communications and electronics industry; and

(3) Attending AFCEA chapter meetings, dinners and a big annual exposition for publicizing your firm and displaying your products.

If *you're* in the field of communications and electronics . . . and want prestige, contacts and exposure . . . let SIGNAL put your company on the *offense* for *defense!* Call or write for more details—now!

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Boston · Chicago · Minneapolis · Los Angeles · San Francisco

AFA

Section Meetings

(Continued from page 30A)

"The Dynamics of Change," C. H. Linder, General Electric Co.; 10/5/61.

"Computer Model of Arterial Fibrillation" G. K. Moe, Masonic Research Lab.; 10 16/61.

"Varactor Harmonic Generators as Power Sources at Microwave Frequencies," M. E. Hines, Waveguide Systems Div., Microwave Assoc., Inc.; 11/9/61.

"The Nature of Biological Data," S. N. Swisher, Univ. of Rochester; 11/13/61.

SACRAMENTO

"U. S. Air Force Ionospheric Scatter Radio Communications System for the North Atlantic Area," J. B. Austin, McClellan Air Force Base; 11/29/61.

ST. LOUIS

Tour of Southwestern Bell Telephone Co.; 9/21/61.

"Space and Engineering," L. C. Jones, St. Louis University; 10/17/61.

"Lighting Control Today," E. L. Lustig, Vickers Inc.; 11/9/61.

SAN ANTONIO-AUSTIN

"Medical Research in Connection with Space Flight," J. P. Stapp, USAF; 11/16/61.

"Radio Astonomy," W. M. Sherrill, Southwest Research Institute; 11/17/61.

"Use of Airborne Radar on the B-52," J. B. Everette, USAF; 12/7/61.

SAN DIEGO

"Electronics—What's Next," R. T. Silberman, Electronics Capital Corp.; 12/6/61.

SEATTLE

"Tilt—Tactics Assessment by Infrared," R. A. Watkins, Raytheon Co.; 9/21/61.

"The IRE and the Space Age," L. V. Berkner, President of IRE; 10/13/61.

"Logical Thinking," V. A. Clark, Jr., Sippican Corp.; 11/16/61.

"Prediction of Radio Frequency Interference by Computer," J. H. Vogelman, Capehart Corp.; 12/14/61.

SHREVEPORT

"Localizer and Its Antenna System Used in Instrument Landings," R. Payne, Greater Shreveport FAA Field Office; "Glide Path and Its Place in the Instrument Landing System," A. Bukowsky, Sheveport Systems Maintenance District Office; 11/7/61.

SOUTH BEND-MISHAWAKA

"Digital Circuits and Techniques," A. Young, Bendix Research Labs.; 10/26/61.

SOUTH CAROLINA

"The Place of the Professional Engineer in A Large Industrial Organization," C. W. Reynolds, Western Electric Co.; 9/8/61.

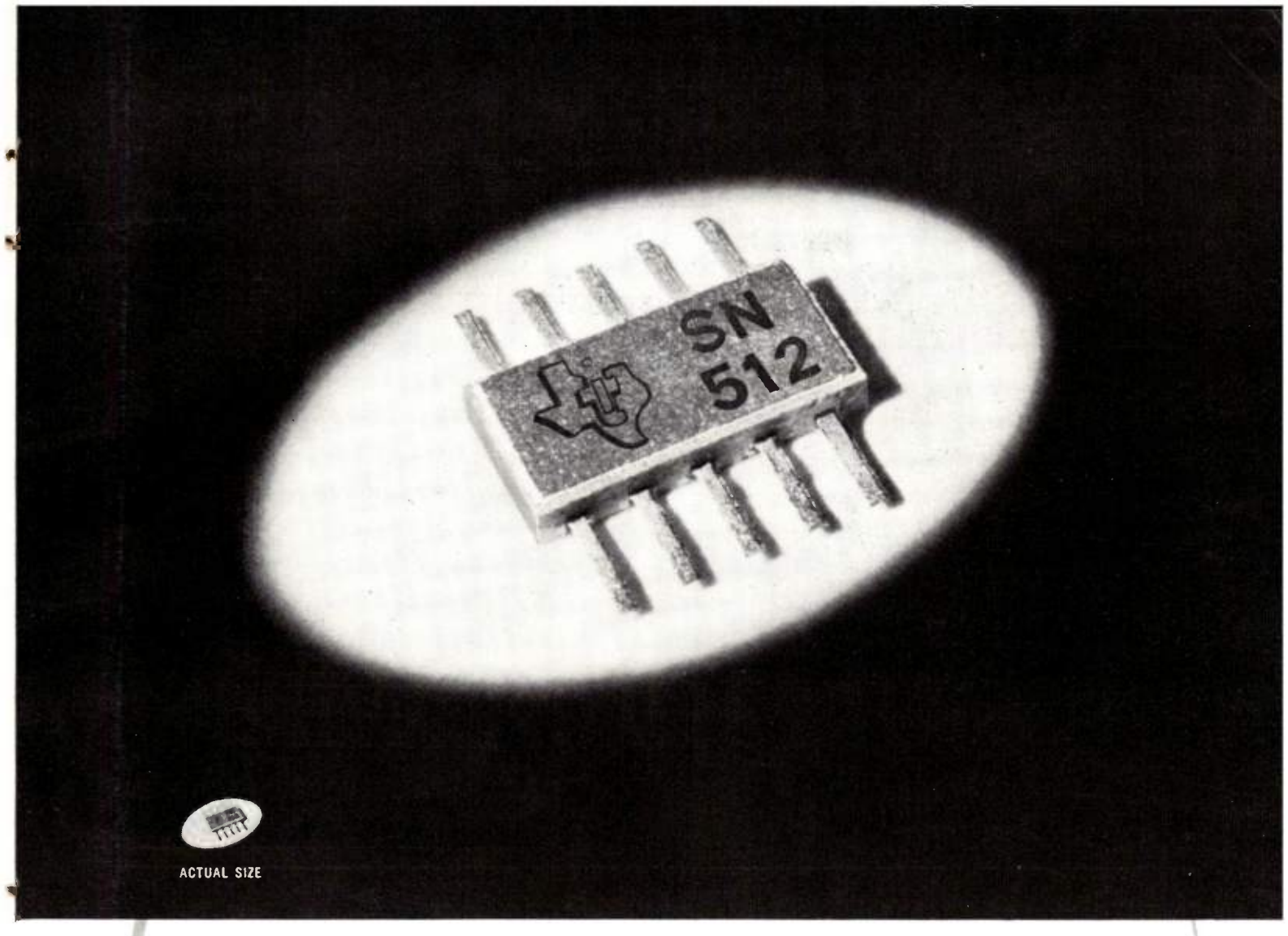
Tour of Telephone Building, H. Creeck: 10/19/61.

"Fundamentals of Radio Astronomy," R. T. McFarland, Western Electric Co.; Talk by B. J. Dasher, IRE Director of Region 3, on proposed IRE/AIEE consolidation; 11/17/61.

TOKYO

"Recent Developments on Microwave Techniques," H. E. Weppler, A.T.T., U.S.A., G. Dawson, S.T.C., U.K., D. G. Jones, G.P.O., U.K., G. Rosman, P.G.O., Australia, R. Kaiser, F. T. Z., Germany; 11/10/61.

(Continued on page 34A)



new shape for the electronics industry!

This tiny electronic device is the **Solid Circuit*** semiconductor network that has ushered electronics into a new era. It was developed and is now in production at Texas Instruments. Beyond its obvious size and weight reduction, it promises reliability and lower costs beginning *now* . . . not three, five or ten years from now!

It already has been used to build a working digital computer with equivalent component density of 12 thousand per cubic inch! Analog type networks for military and industrial uses have been demonstrated and will be in production in the near future. For high reliability military and industrial equipments, the cost already is comparable to the overall cost of circuitry assembled from conventional components.

Even though tiny (1/32" x 1/8" x 1/4" – only a thousandth of a cubic inch!), single semiconductor networks can perform circuit functions requiring some two dozen conventional components – transistors, diodes, resistors, and capacitors. They are inherently more reliable than equivalent circuits using conventional components because network components and interconnections are created within a tiny wafer of solid semiconductor material, which is then hermetically sealed.

The **Solid Circuit** illustrates TI's continuing ability not only to anticipate industry needs but also to explore, develop and produce workable solutions to these needs. The industries served by Texas Instruments will continue to benefit by further such advances in electronics, and also continued leadership in other major fields of TI technology – materials, metallurgy, components, and geophysics.

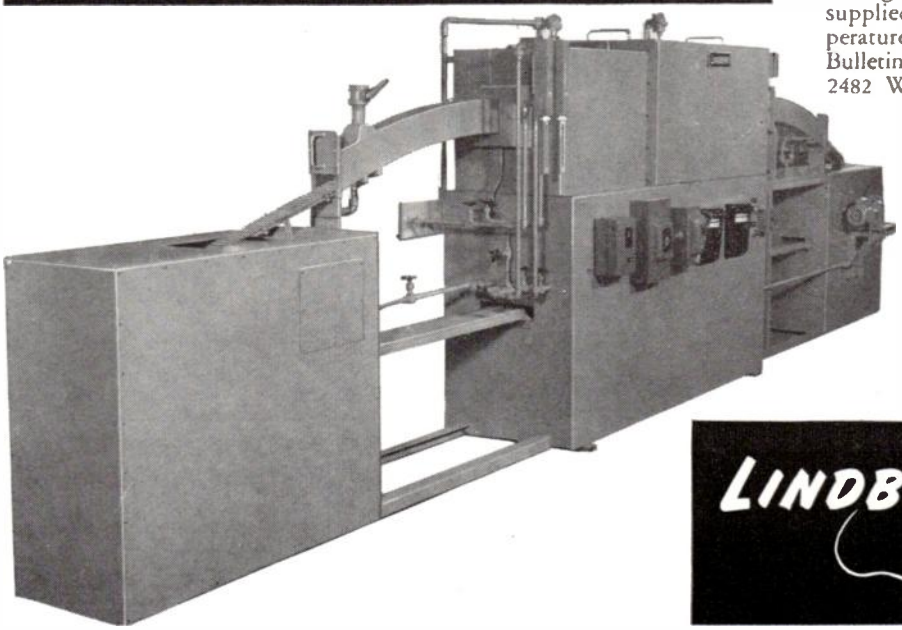
For information about this latest TI advance and the organization which made it possible, write to Merchandising Department. **Trademark of Texas Instruments*

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Lindberg Inclined Mesh Belt Furnaces for Alloying

This type furnace has been specifically designed by Lindberg for alloying, soldering and brazing applications for the semiconductor industry. Special metal resistor element design gives long element life and precise temperature control. Functional furnace construction and adequate insulation insure low shell temperatures. Alloy muffle and belt and precision drive provide precise control and product purity during processing cycle. Inclined or straight through mesh belt furnaces of this type can be supplied for any desired muffle dimension or temperature uniformity. For full information write for Bulletin No. 1083. Lindberg Engineering Company, 2482 West Hubbard Street, Chicago 12, Illinois.



This Lindberg Inclined Mesh Belt Furnace provides two individually heated chambers, preheat, 24" long (max. temp. 400°C.) rated at 7KW and high heat, 36" long (max. temp 1000°C.). Water-jacketed cooling chamber is 72" long. Inconel alloy muffle and furnace hearth. Variable AC drive. Overall dimensions, 66" high, 25'10" long and 30" wide. Working dimensions 4" high over a 3" wide belt.

LINDBERG heat for industry

Section Meetings

(Continued from page 32A)

TOLEDO

Tour of WTOL-TV; 11/16/61.

TUCSON

"The Atlantic Missile Range," L. T. Taylor, Patrick Air Force Base; 10/19/61.

VANCOUVER

"The Polarcardiograph," G. E. Dower, D. Moore, P. Harding, U.B.C.; 11/20/61.

VIRGINIA

"The Navy's Participation in Guided Missile Testing at Cape Canaveral," R. F. Sellars, U. S. Navy; 11/10/61.

"Molecular Electronics—A Key to Reliability," H. H. Steenbergen, Wright Patterson Air Force Base; 12/8/61.

WESTERN MASSACHUSETTS

"Optical Masers," D. A. Park, Williams College; 12/6/61.

WINNIPEG

Annual IRE Student Papers Competition; 11/15/61.

Tour of HMCS Chippewa Barracks, lecture by B. Beare; 12/10/61.

SUBSECTIONS

BUENAVENTURA

"Lasers and Propagation of Light in an Ocean Environment," C. H. Wilcox, Hughes Research Lab.; 10/11/61.

"Nike Zeus," R. Isenson, U. S. Army; 11/8/61.

CATSKILL

"The Effective Use of Marginal Checking in Transistorized Computers," L. W. Snell, IBM Corp.; 12/7/61.

CRESCENT BAY

"Electronic Engineering Aspects of the New Los Angeles Airport Tower," R. Anderson, Federal Airways Administration—L. A. International Airport; "Wescon—1961," W. Peterson, Micro Gee Products, Inc.; 11/20/61.

EASTERN NORTH CAROLINA

"Power Transistor Applications, Ratings, and Problems," B. Gershen, Delco Radio Div., General Motors; 11/10/61.

FAIRFIELD COUNTY

"Recent Meteorological Observations and Interpretations from Weather Satellites," R. Hanel, Natl. Aeronautics Space Administration; 12/7/61.

LANCASTER

"MIA—The Metal Interface Amplifier," J. P. Spratt, Philco Research Center, "MIA as a Circuit Component," S. K. Ghandhi, Philco Research Center; 11/8/61.

MID-HUDSON

"Space Communications," L. Gray, I.T.&T; 11/21/61.

MONMOUTH

"Electromagnetic Wave Interaction with Plasmas," R. Buser, U. S. Army Signal Res. and Dev. Lab.; 11/15/61.

PALM BEACH

"IBM Type 704 EDP System," D. Beutenmuller, Pratt & Whitney Aircraft; 5/16/61.

"A Seismograph System for Off-Shore Oil Exploration," L. C. Pasley, Marine Inst. Co.; 6/27/61.

Tour of local RCA facility and demonstration of 301 Computer, F. Irvin; 9/19/61.

"Training & Education & Techniques of General Analog Computing," I. Hay, Elect. Assoc.; 10/17/61.

"Manufacturing Processes for Silicon Mesa Transistors," P. M. LaHue, Minneapolis-Honeywell; 11/15/61.

SANTA ANA

"Lasers, Principles and Applications," H. Lyons, Electro-Optical Systems, Inc.; 11/15/61.

WESTCHESTER

"Advances in Electron Tubes," S. Gertzis, Amperex Corp.; 10/18/61.

"Optical Systems for Data Processing," R. Williams, Columbia Univ. Hudson Labs.; 11/14/61.

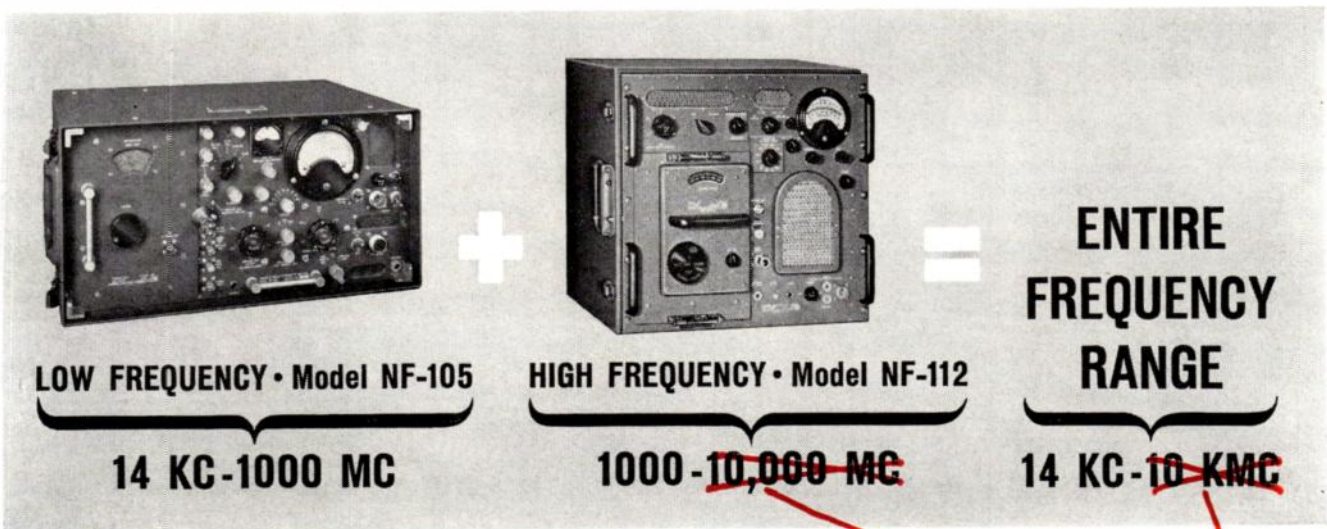
WESTERN NORTH CAROLINA

"Transistor Devices in Motorola Communications Equipment," J. Blanton, Motorola Co., J. Wood, Motorola Co.; 12/1/61.

THE ONLY SIMPLE FORMULA FOR



COMPLETE INTERFERENCE MEASUREMENTS



Through years of hard field use Model NF-105 has acquired an outstanding reputation as a noise and field intensity meter for the frequency range from 150 kilocycles to 1000 megacycles. The versatility of this instrument has now been expanded through the introduction of a new tuning unit extending its coverage down to 14 kilocycles. What is more, this unique measuring equipment has been joined by Model NF-112 which covers the frequency range from 1000 to 10,000 megacycles. The same simplicity, accuracy and speed of operation and reliability of performance which made Model NF-105 so successful have been designed into Model NF-112. Each instrument uses an impulse generator as its calibrator; each combines in one basic unit the components common to all frequency ranges, including the power supply, calibrator, attenuators and metering circuits. All frequency determining components and circuits are contained in plug-in tuning units. Model NF-112, incidentally, uses one single antenna for the entire range from 1000 to 10,000 megacycles.

15,000 megacycles

You save considerably in SIZE, WEIGHT and COST by letting these two instruments do your entire interference measuring job from 14 KC all the way up to 10,000 MC. 15,000 MC

Approved for testing to 10 KMC under Specifications MIL-I-11748, MIL-I-6181D, Category A and MIL-I-26600, Category A.

For complete technical information, send for Catalog 614.

Plan to attend our next seminar on interference instrumentation. Details upon request.

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MANUFACTURERS OF: FIELD INTENSITY METERS • DISTORTION ANALYZERS • IMPULSE GENERATORS • COAXIAL ATTENUATORS • CRYSTAL MIXERS

COINCIDENT CURRENT CORE MEMORIES WITH CYCLE TIMES OF 1.5 MICROSECONDS MADE POSSIBLE BY NEW 30 MIL CORE PLANES

High speed 30 mil O.D. ferrite memory cores are today being wired in 4-wire coincident current matrices on a production basis by Ferroxcube Corporation of America, Saugerties, New York.

The tiny cores, capable of switching at .35 microseconds, make possible read, write cycle times of 1.5 to 2 microseconds and provide outputs of 35 to 45 millivolts when driven by currents comparable to those customarily used with 1 microsecond 50 mil cores.

Tightly spaced on 39 mil centers, Ferroxcube's 30 mil planes also offer the opportunity for size reductions and reduce drive impedances. Standard 30 mil planes are presently offered in 256, 1,024, 4,096 and 16,384 bit sizes, all in 4-wire coincident current patterns. This advance in wiring technique allows design and manufacture of high speed memories at substantially lower cost through utilization of the less expensive coincident current circuitry and lower drive currents.

With cores already in large scale production, and wiring operations building-up rapidly, these high speed stacks are being offered by Ferroxcube at prices comparable to competitors' slower 50 mil stacks. Ferroxcube 50 and 80 mil stacks are priced comparably lower. The company's unexcelled high volume core making facilities and unique, highly reliable wiring techniques have made possible lower and lower costs while increasing uniformity and reliability.

Ferroxcube has rapidly broadened its line of products and its capability in the computer memory field over the past year after a previous history of making memory cores in very high volume for a limited list of customers.

In addition to the 30/18 memory cores in coincident current matrices, Ferroxcube has recently announced such offering as Biax cores wired in planes and stack, stacks with internal temperature controlled over wide ranges of ambient temperature and many other products.

Ferroxcube is also a major producer of ferrite inductive components which include pot cores, E, U, and I cores and toroids—in addition Ferroxcube offers a complete line of thermistors, varistors and light dependent resistors.

Detailed specifications of any product can be obtained by writing to:

FERROXCUBE CORPORATION OF AMERICA
East Bridge St.
Saugerties, New York

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

Calibrating Potentiometer

A low-cost, compact laboratory potentiometer covering a broad field in which the potentiometer method is applicable is now available from **Central Scientific Co.**, a division of **Cenco Instruments Corp.**, 6450 W. Cortland St., Chicago 35, Ill. The Student Potentiometer is suitable for the calibration of meters, checking standard cells and standard resistances and the measuring of temperature by thermocouples.



A printed circuit switch controls fifteen 10-ohm resistors for maintaining a constant potentiometer current for all positions. A ratio-multiplier switch gives the potentiometer three full scale ranges: 0–1.6 0–0.16 and 0–0.016 volt. A calibrated slidewire of stable Karma wire is arranged so that it may be used alone or with 45-ohm "end-coils" as a variable ratio in a bridge circuit or for electrolytic conductivity measurements by the Kohlrauch method. The unit weighs six pounds and its dimensions are $8\frac{1}{2}\times12\times5\frac{1}{2}$ inches. For details contact the firm.

Polarized Relay

Oak Manufacturing Co., Crystal Lake, Ill., has introduced a new high speed Polarized Relay, Type 510, designed for low level switching and sampling in instrument, integrating, computer and multiplexing applications.



A number of exclusive construction features are incorporated for stamina of operation and contact reliability. Among these are the use of ceramic magnets and glass fused-to-metal for all internal insulation and contact supports. Magnetic properties of the ceramic magnet are not affected by temperatures as high as 150°C.

Design of the relay reduces the number of parts to a minimum, resulting in lighter weight, smaller size, plus increased ruggedness in wide extremes of environmental conditions. Its lightweight feature makes it highly adaptable for airborne applications.

The hermetically sealed relays have been subjected to a series of exhaustive reliability tests to establish their ratings. Life expectancy is up to 1 billion operations, pull-in and drop-out time is 700 microseconds maximum and 150 microseconds minimum. Bounce time, both pull-in and drop-out is 100 microseconds maximum.

Noise is at the low level of less than 100 microvolts peak-to-peak into 1 megohm with preamplifier bandpass of 8-1000 cps.

Blackbody Infrared Calibration Standard



Operating in the temperature range of 150° to 1000°F the new blackbody calibration standard is product of **Radiation Electronics Co., Division of Comptometer Corp.**, 5600 Jarvis Ave., Chicago 48, Ill., features a proportional temperature controller with a control precision of ±2°F. The new unit also features $2\frac{1}{8}$ inch diameter aperture that permits calibration to be accomplished without critical alignment problems.

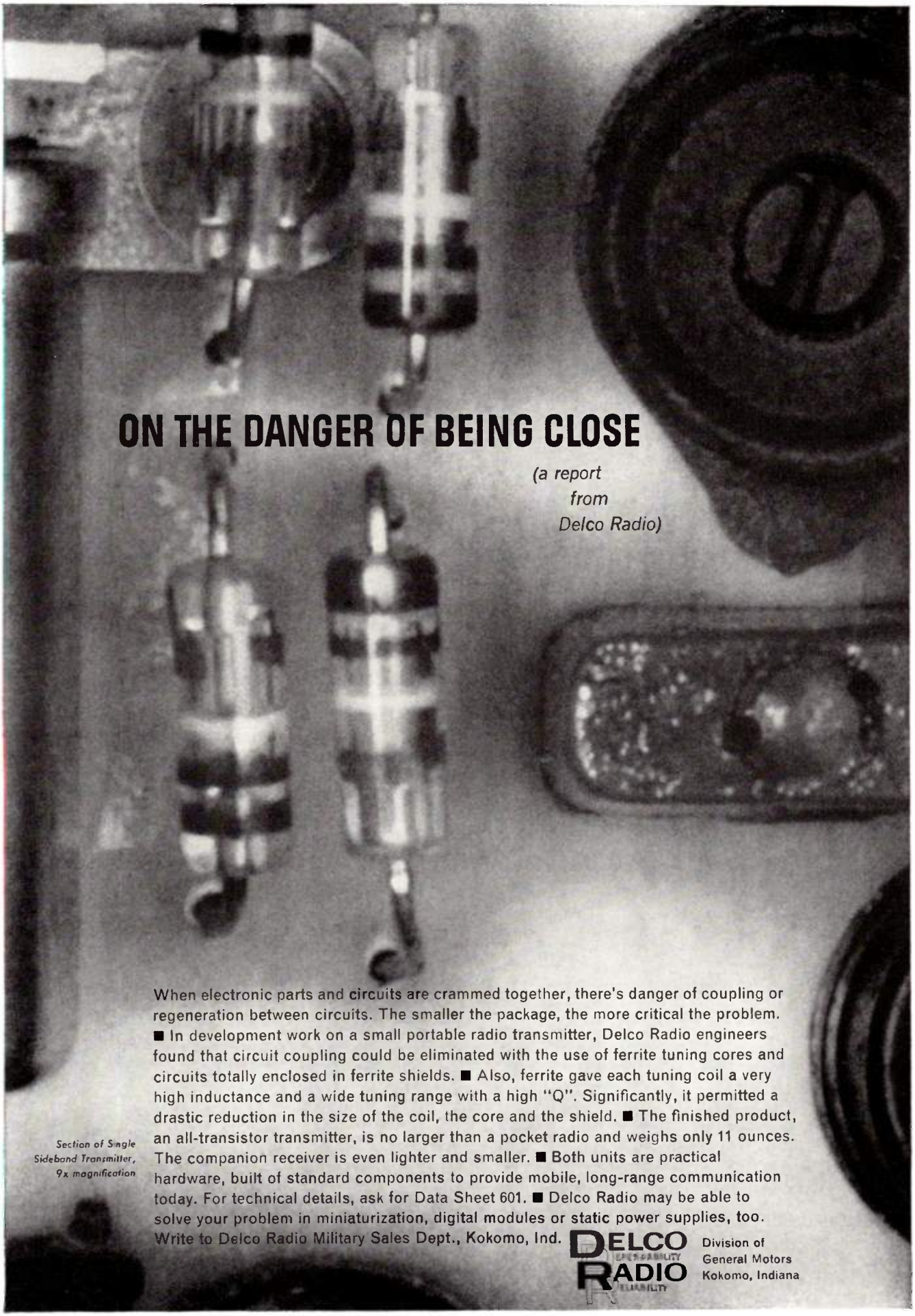
Price of the new blackbody is $950.00 and delivery is two weeks.

Low Cost Thermoelectric Devices

Many special applications of cooling module designs adapted to customer products coupled with new manufacturing techniques have resulted in substantial reductions in prices for standard thermoelectric devices according to an announcement from **Ohio Semiconductors**, 1025 Chesapeake Ave., Columbus 12, Ohio.

O.S.'s substantial price reductions are expected to make possible many T. E. applications once considered but ruled out due to cost alone. This re-evaluation should be especially true of applications in which relatively small numbers of cooling units are required.

(Continued on page 38A)



ON THE DANGER OF BEING CLOSE

(a report from Delco Radio)

When electronic parts and circuits are crammed together, there's danger of coupling or regeneration between circuits. The smaller the package, the more critical the problem. ■ In development work on a small portable radio transmitter, Delco Radio engineers found that circuit coupling could be eliminated with the use of ferrite tuning cores and circuits totally enclosed in ferrite shields. ■ Also, ferrite gave each tuning coil a very high inductance and a wide tuning range with a high "Q". Significantly, it permitted a drastic reduction in the size of the coil, the core and the shield. ■ The finished product, an all-transistor transmitter, is no larger than a pocket radio and weighs only 11 ounces. The companion receiver is even lighter and smaller. ■ Both units are practical hardware, built of standard components to provide mobile, long-range communication today. For technical details, ask for Data Sheet 601. ■ Delco Radio may be able to solve your problem in miniaturization, digital modules or static power supplies, too. Write to Delco Radio Military Sales Dept., Kokomo, Ind.

Section of Single Sideband Transmitter, 9x magnification

DELCO RADIO

Division of General Motors
Kokomo, Indiana



TYPE LF relay shown (cover removed) is 2-coil design which controls entire latching operation within relay.(Actual size).

CLARE LATCHING SUBMINIATURE crystal can RELAY

The new CLARE Type LF, magnetic latching subminiature relay offers designers simplified circuitry in small space by providing latching effect without transistors. Magnetic latching results in power economy.

The Type LF is available with either 2-coil or 1-coil configuration. The 2-coil relay allows complete control of the latching operation within the relay and provides an extremely compact operating unit. The 1-coil relay is somewhat more sensitive; it is adaptable to existing circuits where outside control is provided. The Type LF provides the same wide range of mounting arrangements and terminals as the CLARE Type F relay.



FOR NON-LATCHING OPERATION

CLARE *Type F* SUBMINIATURE CRYSTAL CAN RELAY

The CLARE Type F relay is extremely fast and more than moderately sensitive. It is built to withstand temperature extremes, heavy shock and extreme vibration. Contacts, rated at 3 amperes, are excellent for low-level circuit operations. Send for Design Manual 203.

For coil and mounting data on CLARE Type LF relay send for CPC-12. Address: C.P. Clare & Co., 3101 Pratt Blvd., Chicago 45, Illinois. In Canada: C. P. Clare Canada Ltd., 840 Caledonia Road, Toronto 19, Ontario. Cable address: CLARELAY.

C. P. CLARE & CO.

Relays and related control components

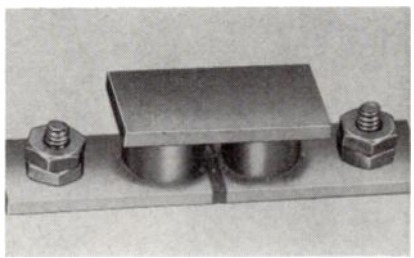
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

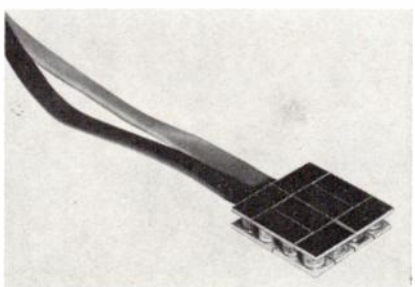
(Continued from page 36A)

Included in the categories of increased availability and price reductions are three TM series thermoelectric devices:

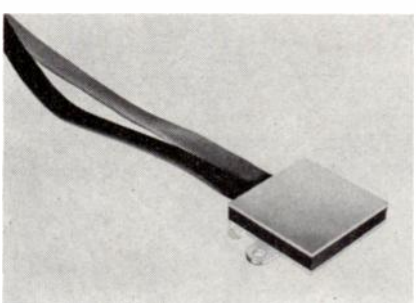
The TA-12 "Thermo-Cell" thermoelectric junction, capable of pumping as much as 3 watts or typically attaining a temperature differential in excess of 85°C. The TA-12 measuring $\frac{5}{16}$" high, $1\frac{13}{32}$" wide and $\frac{11}{32}$" deep, operates in a temperature range of −100°C to +125°C. It can be used for spot cooling, as in cases where overheated components otherwise impair the efficiency of a circuit, and to meet many other special thermal requirements. The TA-12 is available off-the-shelf in small quantities and requires two weeks' delivery in reasonable quantities. Prices of the TA-12 are $10.00 (1 and 2); $7.50 (3 to 9); $5.00 (10 to 29); $4.50 (30 to 99); $4.00 (100 to 299); $3.50 (300 to 999); $3.00 (1000 to 2999); and $2.00 (3000 or more).



The TA-20 "Thermo-Array" of 8 thermoelectric junctions is capable of pumping as much as 30 watts (100 BTU per hr.) or typically attaining a temperature differential in excess of 85°C. The TA-20 is for installation in a manufacturer's thermal design for cooling or heat pumping purposes and can be used in parallel to increase capacity. The TA-20's dimensions are: height, $\frac{5}{16}$"; width, $1\frac{31}{32}$"; depth $1\frac{9}{16}$". Prices are $20.00 (1 to 29); $19.00 (30 to 99); $18.00 (100 to 299); $17.25 (300 to 999); $13.20 (1000 to 2999); $10.00 (3000 or more).



The TA-20M "Thermo-Module" of 8 thermoelectric junctions can pump as much as 30 watts or typically attain a temperature differential in excess of 80°C. It can be used in series or parallel to increase capacity and comes complete with electrically insulated aluminum mounting plates $\frac{1}{16}$" thick. The TA-20M is for use as a complete thermoelectric unit for cooling or heat pumping purposes. Special mounting plates can be provided with volume orders. The TA-20M measures $\frac{7}{16}$" high, including plates; $2\frac{3}{8}$" wide, including "ears" on bottom mounting plate; and $1\frac{1}{2}$" deep. The TA-20M is priced as follows: $25.00 (1 to 29); $24.00 (30 to 99); $23.00 (100 to 299); $22.50 (300 to 999); $17.25 (1000 to 2999); $11.00 (3000 or more).

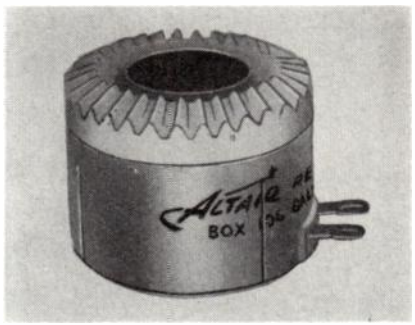


Like the TA-12, the TA-20 and TA-20M operate in a temperature range of −100°C to +125°C. The latter two devices come equipped with a $5\frac{1}{2}$" lead complete with solderless terminals.

Additional information can be obtained from Sales Department, Ohio Semiconductors.

Magnetic Clutch

New, miniature magnetic clutch, Model MC541, a product of **Altair Research & Manufacturing Co.**, Box 106, Baldwin Park, Calif., fitted with steel housing and a nylon gear mounted on ball bearings, has been specifically designed for incorporation in multiple clutch-potentiometer assemblies. If desired, a coupling is available which converts clutch to an in-line drive.

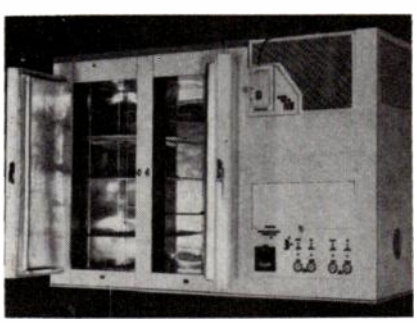


Engineered to meet critical space requirements, the MC541 clutch is one inch in diameter and one inch long. It has a minimum torque of 6 ounce/inch and exerts no drag on potentiometer when de-energized. Available with coil voltages of 6 to 110 volts dc.

The price is $15.00 per unit, FOB factory, small quantities are available for immediate delivery.

Environmental Chamber

A new environmental chamber from **Solar Systems, Inc.**, 11936 Valerio St., North Hollywood, Calif., producing temperatures down to −125° F., is designed to meet all applicable Mil Specs for the testing and storing of wide variety of electronic, mechanical, solid state, electromechanical components and cryogenic fluids. Unit shown is upright Model 4033-100, available in exteriors of baked enamel or stainless steel, and measuring 130″ long, 72″ high, and 36″ deep.



This series of chambers, all of which feature self-contained mechanical refrigeration, fully automatic defrost, adjustable shelves, gradient of 15°F., and stainless steel interiors, are available in capacities of 40, 80, 120, and 240 cubic feet.

Front panel instrumentation consists of indicating lights, pressure and temperature gauges, and temperature control knob. The semi-hermetically sealed compressor is air cooled and available in either 220 or 440 V. on a control circuit of 115 V.

(Continued on page 90A)

FAST Delivery WRIGHT Servos

24 HOURS
400 Cycle Servo Motors
Motor Tachometers
Inertia Damped Motors

10 DAYS
MOUNTED WITH PRECISION CLASS III GEARHEADS

This NEW service on quantities up to 10 standard pieces per item. No specification modifications. Full production quality. Meet all MIL specs. Gearheads not sold separately.

SERVO MOTORS	FRAME SIZE	VOLTS Ø1	VOLTS Ø2	STALL TORQUE	SPEED RPM	STALL PWR/Ø	ACCEL. RAD/SEC²
20D633-2C	8	26	26	.22 OZ. IN.	6200	2.5W	106,000
20D633-4C	8	26	36 CT	.22 OZ. IN.	6200	2.5W	106,000
20D632-4C	8	26	36 CT	.35 OZ. IN.	6200	3.1W	99,800
20D627-2C	8	115	115/57	.33 OZ. IN.	6200	3.5W	98,000
20D603-2C	11	115	115/57	.60 OZ. IN.	6200	3.5W	43,300
20D603-4C	11	115	36 CT	.60 OZ. IN.	6200	3.5W	43,300
20D590-2A	15	115	115/57	1.3 OZ. IN.	4800	6.2W	27,800
20D612-2C	18	115	115/57	2.3 OZ. IN.	4800	9.1W	31,000

MOTOR TACHS	FRAME SIZE	VOLTS Ø1	VOLTS Ø2	TACH. VOLTS	STALL TORQUE	SPEED RPM	V/1000 RPM	TOTAL NULL	ACCEL. RAD/SEC²
20D628-2C	8	115	115/57	26	.33 OZ. IN.	6200	.20	.019V	75,800
20D631-4C	8	26	36 CT	26	.35 OZ. IN.	6200	.20	.019V	80,500
20D634-2C	8	26	26	26	.22 OZ. IN.	6200	.20	.019V	75,000
20D634-4C	8	26	36 CT	26	.22 OZ. IN.	6200	.20	.019V	75,000
20D604-2F	11	115	115/57	115	.60 OZ. IN.	6200	.500	.019V	32,600
20D604-4F	11	115	36 CT	115	.60 OZ. IN.	6200	.500	.019V	32,600
20D593-2A	15	115	115/57	115	1.3 OZ. IN.	4800	3.1	.019V	17,500
20D614-2C	18	115	115/57	115	2.3 OZ. IN.	4800	3.1	.019V	25,900

INERTIA DAMPED	FRAME SIZE	VOLTS Ø1	VOLTS Ø2	STALL TORQUE	STALL POWER	SPEED RPM	DAMPING
20D618-2B	8	115	40/20	.30 OZ. IN.	3.5W	6200	40 DYNE CM
20D605-2D	11	115	115/57	.60 OZ. IN.	3.5W	6200	100 DYNE CM
20D613-2C	18	115	115/57	2.3 OZ. IN.	9.1W	4800	750 DYNE CM

Gear Backlash and Torque

Size 8: 30′ max. backlash: 1 in. oz. rev. load. 5 inch ounces operating load torque.

Size 11: 30′ max. backlash: 4 in. oz. rev. load. 25 inch ounces operating load torque.

Size 15: 30′ max. backlash: 5 in. oz. rev. load. 35 inch ounces operating load torque.

Size 18: 30′ max. backlash: 10 in. oz. rev. load. 60 inch ounces operating load torque.

DIVISION OF
WRIGHT SPERRY RAND
DURHAM, NORTH CAROLINA TEL. 682-8161

"Servos For New Horizons"

Stainless Steel ATTENUATORS by WEINSCHEL



Model 210
Range: 1 to 20 DB
Connectors: Type N

The ruggedness and longer life of stainless steel connectors and metal parts make these attenuators exceptional—and only Weinschel makes them. The Model 210 has these additional

Exclusive Weinschel Features:

- Weinschel film resistors withstand shock and vibration and give maximum stability under peak pulse power and under extreme temperature and humidity cycling.
- Certificate of Calibration showing insertion loss test data with guaranteed accuracy explicitly stated.
- Critical dimension of inner contact depth held to ± 0.005 inches, exceeding all government specifications.

Write for Weinschel Engineering Bulletin 17 for full information and prices on the Model 210 and similar attenuators with other connectors. For special models to meet other requirements, contact our Application Engineering Department.

WEINSCHEL ENGINEERING

KENSINGTON, MARYLAND
Phone: LOckwood 4-0121
TWX Kensington, Maryland 446

IRE People

Budd-Stanley Co., Inc., Syosset, N. Y., has announced the appointment of **Tore N. Anderson** (S'47–A'49–SM'55) as Vice President for Engineering. Budd-Stanley is a major producer of precision microwave components and instruments. He will devote his efforts to microwave technology, directing new microwave product development and guiding the establishment of a new Budd-Stanley Research and Development Group.



T. N. ANDERSON

Prior to his Budd-Stanley appointment, he was President of FXR, Inc., and a Director of Micromega Corporation, an FXR subsidiary. His previous experience includes 12 years at Airtron, Inc., where he served as Chief Engineer, Director of Engineering and Vice-President, and was responsible for development work in coaxial and waveguide transmission line components. For six years he was chairman of the EIA Committee on Wave Guide Connector Standardization and is presently the U. S. delegate to the International Electrotechnical Commission for International Waveguide and Flange Standardization.

Mr. Anderson holds a group of patents covering the design and applications of microwave components. In 1959 he was appointed consultant to the Defense Department's Advisory Group on Electronic Parts. A native of Sweden, he graduated from Oregon State College and the Cooper Union with degrees in Electrical Engineering. He is a life member of Tau Beta Pi and an associate editor of the *Microwave Journal*. During World War II he served with the Engineering Research and Development Agency, Fort Belvoir, Virginia as project officer on underwater mine detectors, and as Chief of the Laboratory and Test Section in mine detector development.

❖

Dynamic scientist and industrialist **Isaac L. Auerbach** (S'46–M'49–SM'52–F'58), recently received a Citation of Honor from the Drexel Institute of Technology. As a prominent alumnus, he was honored for his unusual achievements in the new and burgeoning field of information sciences and technology.

Mr. Auerbach is the first President of the International Federation of Information Processing (IFIP), a world organization representing technical societies of nineteen nations, of which he was founder and propelling force. As a result of his contribution to the international aspects of a technical field only fifteen years old, there is to be an IFIP Congress 62, held in Munich, Germany, from August 27 to September 1, 1962. The Congress is intended to offer computer scientists and other specialists in the information processing field a unique opportunity for interchange of ideas and technical developments. In connection with his work with IFIP, Mr. Auerbach has organized the Technical Committees on Terminology and Symbols and Programming Languages.

President of the Auerbach Corporation of Philadelphia and New York, he was founder of the Philadelphia Chapter of the Technion Society, which supports the Israeli Institute of Technology in Haifa, Israel. He is current Co-Chairman of the Electronics Division of the Trade Council of the Allied Jewish Appeal. He was one of the country's pioneers and leaders in computer technique development and its application to digital systems. His technological breakthrough in magnetics led to digital communication systems of unparalleled reliability. He developed and designed the first application of magnetic cores to high-speed, coincident-selection memories. His name is closely associated with such early computers as BINAC, UNIVAC, SAGE system radar data processor, and with the ATLAS missile guidance computer. From 1957 to 1960 he was scientific advisor to UNESCO on automation and information processing. In 1959, he was the recipient of the Grand Medal of the City of Paris, awarded for his outstanding service in helping to organize the UNESCO-sponsored First International Conference on Information Processing.

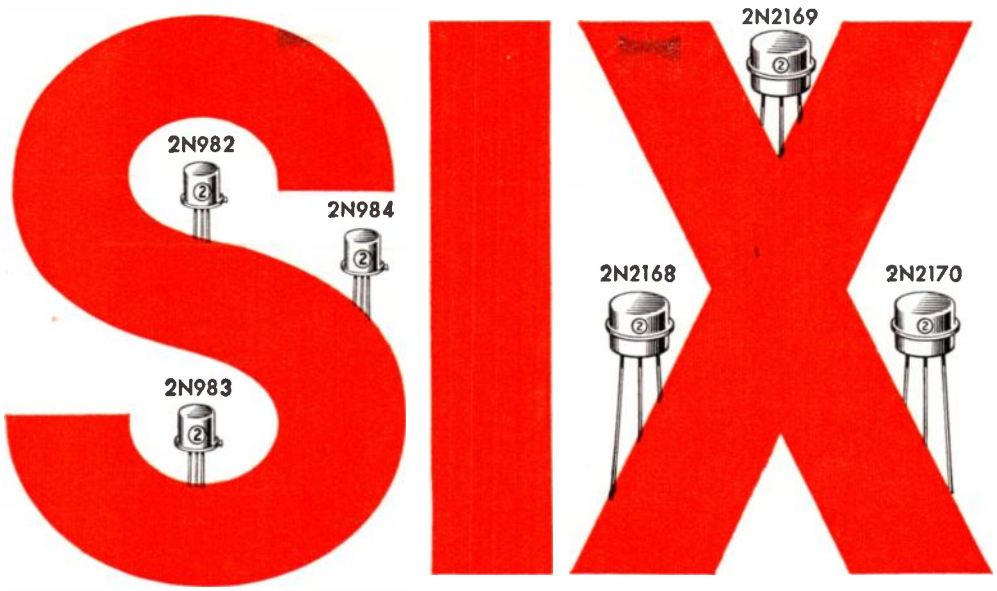
Mr. Auerbach received the B.S. degree in Electrical Engineering from Drexel Institute of Technology, and the M.S. degree in Applied Physics from Harvard University. He has published extensively in technical journals and holds numerous patents. He received the IRE Fellow Award for his contributions to the development and application of computer techniques, and the 1961 Achievement Award for his personal contributions, stimulation, and leadership in the international exchange of information in the electronic computer field.

❖

PRD Electronics, Inc., Brooklyn, N. Y., subsidiary of Harris-Intertype Corporation, microwave and electronic test equipment manufacturer, has announced the appointment of **Norman Ellen** (SM'57) to the position of Senior Project Engineer, Products and Components Division. He has returned to PRD after a nine year absence. In 1952 he left PRD to join General Precision Labora-

(Continued on page 42A)

New from Sprague!



ALL-NEW MADT® TRANSISTORS

Wide Range of Parameters *Plus* Choice of Package Tailored to Meet Your Actual Circuit Requirements

Sprague now offers a complete new line of ultra-high-speed switching transistors. For greater design flexibility, these Micro-Alloy Diffused-Base Transistors provide interchangeability of packages (identical specifications are available in either the TO-9 or TO-18 case).

THIS NEW FAMILY OF SWITCHES OFFERS:

- High BV_{CEO} Ratings
- Low Saturation Voltage
- Low Storage Time
- Low Output Capacitance
- Excellent Frequency Response
- Close Parameter Control in Your Choice of Package

HERE ARE SOME KEY PARAMETERS:

Type No.	JEDEC CASE	Max. I_{CBO} (μa)	Min. BV_{CES} (volts)	Min. BV_{CEO} (volts)	Min. h_{FE}	Typ. f_T (mc)
2N982	TO-18	3	20	15	50	450
2N2168	TO-9					
2N983	TO-18	3	15	15	40	450
2N2169	TO-9					
2N984	TO-18	5	15	10	20	350
2N2170	TO-9					

For application engineering assistance, write Product Marketing Section, Transistor Division, Sprague Electric Co., Concord, N.H.

For complete engineering data, write Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

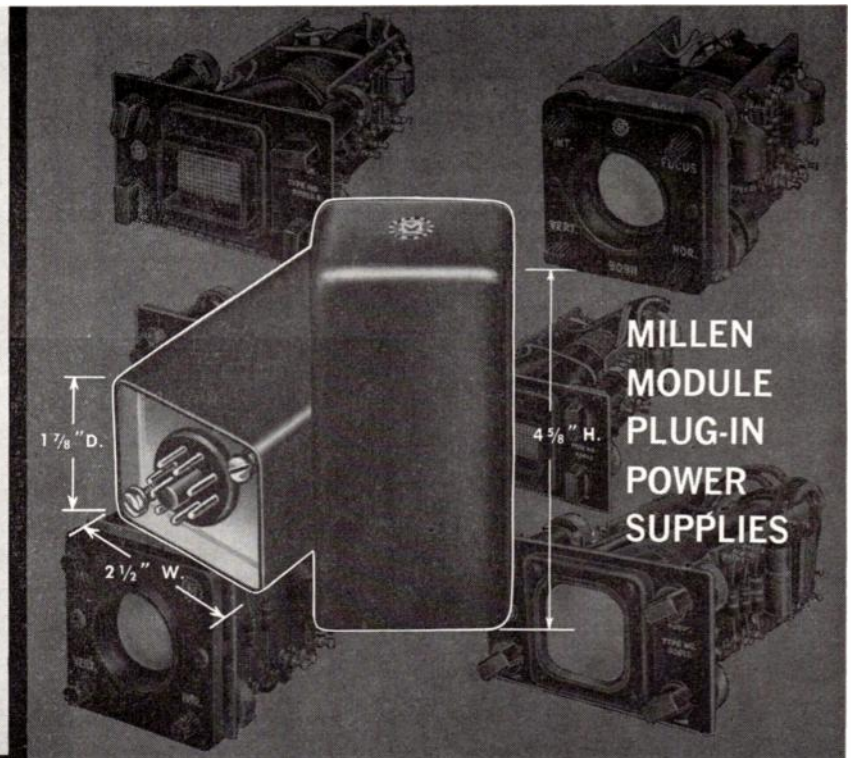
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CAPACITORS
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Designed for M Application



MILLEN MODULE PLUG-IN POWER SUPPLIES

MODULE POWER SUPPLIES

Plug-in high voltage and low voltage oscilloscope power supplies are provided in economical compact packages. No. 90202 supplies 750 v.d.c. at 3 ma. to 1020 v.d.c. at 1.1 ma and 6.3 v.a.c. No. 90203 supplies plate and heater power for four amplifiers.

JAMES MILLEN MFG. CO., INC.
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6 colorful, fully illustrated pages on new, expanded applications for Phelps Dodge Styroflex®, Spirafil and Foamflex Air Dielectric coaxial cable . . .

- Delay Lines
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- Special Bent Shapes
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New Haven 8, Conn.
Fabricators-Distributors of
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IRE People

(Continued from page 40A)

tory, Inc., Pleasantville, N. Y., as Associate Section Head of the firm's Microwave Components Section. At GPL, he was responsible for the design and fabrication of microwave components and systems.

Prior to joining PRD in 1950, he was associated with The Channel Master Corporation, Ellenville, N. Y., in the designing and developing of VHF and UHF antennas and multicouplers. From 1948 to 1949 he held the position of radio engineer responsible for the installation, field modification, and flight testing of radio navigational aids in low frequency and VHF ranges, with the Civil Aeronautics Administration. He has also been associated with the Premier Microwave Corp., Port Chester, N. Y., and Ess-Gee, Inc., Elmsford, N. Y.

In his present capacity as senior microwave project engineer, Mr. Ellen will be responsible for the design and development of special order requirements. His educational background includes a B.E.E. degree from the City College of New York (1948) and graduate study in Electronics at the Polytechnic Institute of Brooklyn.

❖

Use your IRE DIRECTORY! It's valuable!

Dr. Carl H. Becker (M'57) has been named to the position of Director of Research of Precision Instrument Company, San Carlos, Calif. He was born in Frankfurt/Main, Germany, and received his scientific education there in the fields of engineering physics, atomic physics, molecular physics, electrodynamics, physical chemistry and thermodynamics, and was a former student of Arnold Somerfeld in theoretical physics. He graduated at the Technical University of Munich, Germany, with the Diploma of Engineering Physics in 1928 and the Doctorate of Engineering Physics in 1930. His professional background in Europe was concerned mainly with research and development in the field of photographic and magnetic recording, as well as electro-acoustics. His scientific contributions have resulted in more than twenty German patents and new principles of multiple-channel recording.

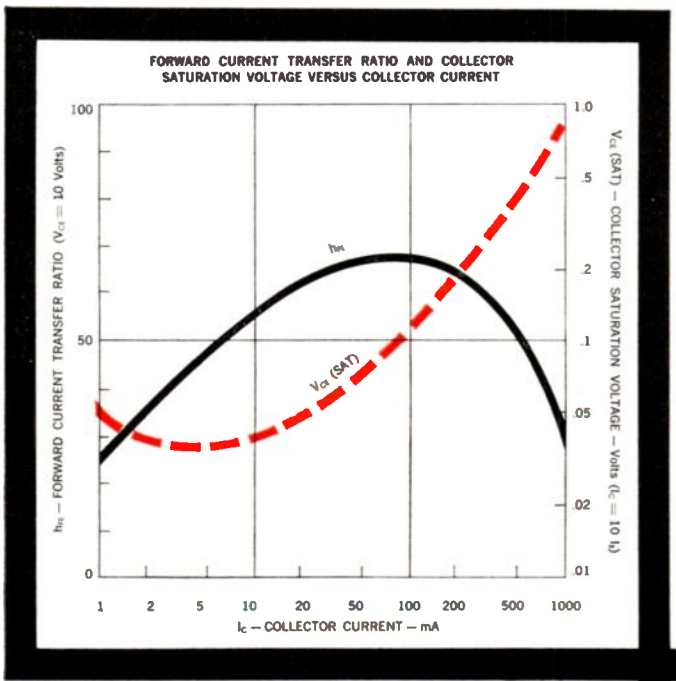


C. H. Becker

From 1930 to 1943 he was associated with UFA, at that time the largest European motion picture company, in charge of development and research in the photographic and magnetic recording field. He owned and managed his own research company in electro-acoustics from 1943 to

(Continued on page 44A)

2N2297 SILICON PLANAR EPITAXIAL



1 VOLT MAX V_{CE} (SAT) AT 1 AMP

ELECTRICAL CHARACTERISTICS & RATING AT 25°C UNLESS OTHERWISE NOTED

Storage Temp.		−65 to +300°C
Operating Temp.		200°C Maximum
Power Dissipation		1 watt @ 25°C ambient temp. 5 watts @ 25°C case temp.
BV_{cbo}	@ 0.1 mA	80 volts MIN
BV_{ebo}	@ 0.1 mA	7 volts MIN
LV_{ceo}	@ 30 mA	35 volts MIN
h_{FE}	@ I_c = 150mA Vce = 10 volts	40 to 120
h_{FE}	@ I_c = 1.0A V_{CE} = 10 volts	15 MIN
V_{CE} (SAT)	@ I_c = 1 ampere I_b = 0.1 ampere	1 volt MAX
V_{CE} (SAT)	@ I_c = 150mA I_b = 15mA	.2 volts MAX
V_{BE} (SAT)	@ I_c = 1 ampere I_B = 0.1 ampere	2 volts MAX
C_{ob}	@ V_{cb} = 10 volts	12 pf MAX
Gain Bandwidth:	@ V_{ce} = 10 volts I_c = 50 mA f = 20 mc	60 mc MIN
I_{cbo}	@ 60 volts & 25°C 150°C	10 nA MAX 10 μA MAX

For more detail see data sheet. Available from distributors.

TYPICAL APPLICATIONS

MEMORY DRIVERS

T_{ON} < 70 n sec

T_{OFF} < 150 n sec

I_C 1 AMP @ 50 V.

RF AMPLIFIERS

P_O 1.3 watts

@ 27 mc V_{CC} = 12V

CONVERTERS

FAIRCHILD

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545 WHISMAN ROAD, MOUNTAIN VIEW, CALIF. • YORKSHIRE 8-8161 • TWX: MN VW CAL 853

A DIVISION OF FAIRCHILD CAMERA AND INSTRUMENT CORPORATION

IRE People

(Continued from page 42A)

1952. After emigrating to the United States in 1952, his professional activities concentrated on the introduction of molecular physics in various electronic fields. Some of these activities are characterized by the discovery of high-frequency magnetization and the establishment of new methods of geophysical exploration. Recently he has been engaged in quantum electronics in the solid state, which yielded the phonon-maser. He has operated as an independent consulting physicist, and as Consulting Professor at Stanford University, where he taught a course in molecular engineering and quantum electronics. In 1957 he was Director of Radiation Physics for the Trionics Corporation in Madison, Wis., and he served as Advisory Engineer to Westinghouse Electric Corporation from 1958 to 1961. At the 1961 meeting of WESCON in San Francisco, he delivered a paper outlining for the first time a method of amplification that uses no electric currents or electrical effects in the amplifications mechanism, but relies on the nonlinear acoustic phenomena in a quartz crystal.

Dr. Becker is a member of the American Physical Society.

❖

The appointment of **Beardsley Graham** (S'39–A'41–SM'47) of Palo Alto, an executive of the Missiles and Space Company of Lockheed Aircraft Corporation, to the presidency of Spindletop Research Inc., Lexington, Ky., was announced recently. He will head the formation of a research center at Lexington to bring scientific effort to bear on the development of Kentucky's industrial potential and greater utilization of the state's labor and natural resources.



B. GRAHAM

Associated with Lockheed and its Missiles and Space subsidiary at Sunnyvale and Palo Alto for the past five years, he was Manager of Satellite Research Planning. Previously he was an Assistant Director of Stanford Research Institute, Menlo Park, and Manager of SRI's Mountain States Division at Phoenix, Ariz. Earlier in his career he held a succession of responsible research and development positions with Bendix Aviation Corporation of Detroit, Lewyt Corporation of Brooklyn, National Broadcasting Company in Hollywood, and New York and Pacific Telephone and Telegraph Company in San Francisco.

A native of Berkeley, Calif., Mr. Graham is a graduate of the University of California, where he also had graduate studies. He had additional studies at Columbia University.

❖

Milton Halpern (A'43–M'58) has been promoted to chief engineer, special projects, of Instruments for Industry, Inc., of Hicksville, N. Y. He had been a supervising engineer with the company, an independent developer of electronic systems and precision components. He will be responsible for all special engineering programs. Before joining IFI in 1957, he was project engineer with Kollsman Instruments Corporation. Earlier, he was assistant project engineer with Sperry Gyroscope Corporation.



M. HALPERN

Mr. Halpern holds a B.S. and a M.S. degree from City College of New York. He is a member of Tau Beta Pi and Eta Kappa Nu.

❖

Metcom, Inc., Salem, Mass., manufacturers of microwave tubes and devices, announces the appointment of **Wellesley J. Dodds** (SM'45–F'61) as Chief Engineer of beam tubes. He will be responsible for gas plasma amplifiers and traveling-wave devices. He comes to Metcom from Bomac Laboratories where he was Vice President of Engineering. He has also been associated with Varian Associates and RCA in various engineering capacities.



W. J. DODDS

Mr. Dodds has spent almost twenty years working on microwave devices and holds several patents in this field. He is a Fellow of the IRE, and a member of Sigma Xi, Sigma Pi Sigma and Pi Mu Epsilon.

❖

First Lieutenant David C. Ayers, Jr. (S'54–A'54–M'60) has been assigned as a product engineer at the U. S. Army Signal Material Support Agency, Fort Monmouth, N. J. He will review and analyze signal equipment to be procured, and recommend desirable design changes to improve equipment producibility by assembly-line methods. Before his recall with the 301st Signal Company, he headed a group of engineers at Collins Radio Co., Cedar

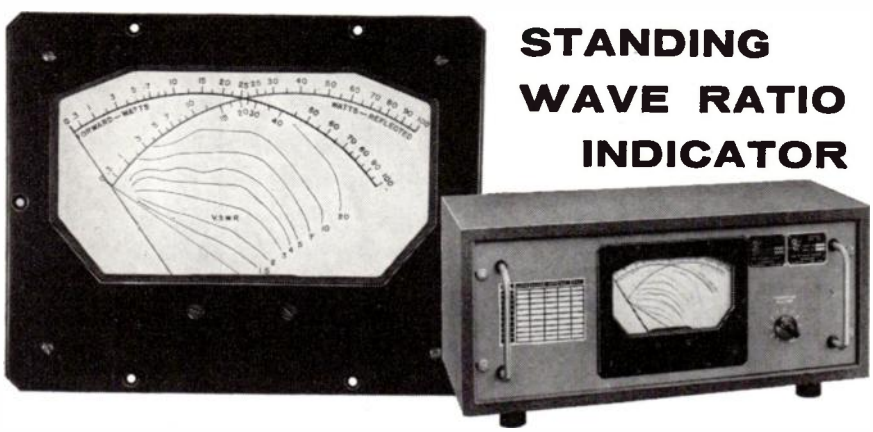


D. C. AYERS JR.

(Continued on page 46A)

SWR-3000

STANDING WAVE RATIO INDICATOR



The TMC Model SWR-3000 is a multi-purpose instrument which will instantaneously provide visual indications of Forward Power, Reflected Power, and Voltage Standing Wave Ratio.

The SWR-3000 may be used in any 50 or 70 ohm unbalanced transmission system covering 2-30 MCS, with average powers up to 3000 watts.

The SWR-3000 is used as an operational and maintenance tool at transmitter stations, in electronic plants for production testing of transmitters, and laboratories for RF transmission system measurements.

For additional information about the SWR-3000 and other test equipment, please contact TMC Test Equipment Division, Mamaroneck, New York.

The Test Equipment Division Of

THE TECHNICAL MATERIEL CORPORATION

World Wide Suppliers of Electronic Communications Equipment

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NICKEL CADMIUM RECHARGEABLE BATTERIES

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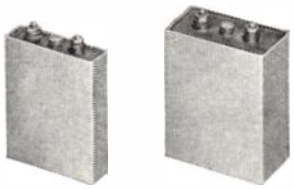
3 TYPES HERMETICALLY SEALED CELLS



BUTTON 20-500 mAh

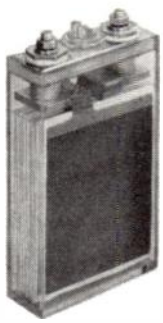


CYLINDRICAL 100 mAh-4.0 Ah



RECTANGULAR 1.5-2.3 Ah

2 TYPES VENTED CELLS



SINTERED PLATE 500 mAh-240 Ah (Miniaturized and Large Industrial Types)



POCKET PLATE 10 Ah-2048 Ah

NICAD Batteries, in miniaturized Sealed and Vented Cells, are constantly in demand where there is a vital need for highly reliable and durable power sources. Suitable for the most stringent military requirements, these versatile cells are also adaptable to a wide variety of critical commercial, consumer and industrial applications.

Rechargeable, miniaturized, lightweight, NICAD Sealed Cells require no maintenance and operate in any position. High current discharge capabilities and extended shelf life make these cells ideal for battery-powered equipment that must operate instantly and efficiently.

NICAD sintered plate Vented Cells are designed to give exceptionally long life under cycle and float service. They can be stored in any state of charge for long periods without attention or fear of deterioration. They will operate in any position on discharge.

Pocket plate types are ideally suited for standby power and control, signal and alarm systems, emergency lighting, stationary engine cranking, cycling service, switchgear tripping and closing, locomotive starting. They seldom require maintenance, have an exceptionally long life and will maintain sustained voltage at heavy discharge rates. There's no sulphation, no corrosive fumes.

For further information on both Vented and Hermetically Sealed Cells, additional usage and application, write . . .



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NICAD is a trademark of Gould-National Batteries, Inc.



CONSUMER PRODUCTS
Cordless Appliances
Dictating, Recording Equipment
Shavers
Radios
Hearing Aids
Electronic Medical Aids





ELECTRONICS
Instruments
Receivers/Transmitters
Computers
Telemetering
Portable Communications Equipment



MISSILES AND ROCKETS
Control
Guidance
Telemetering



AIRCRAFT
Engine Starting
Emergency Lighting
Controls and Alarms
Rescue Gear



COMMUNICATIONS
Telephone
Telemetering
Microwave

TYPICAL USES



POWER PLANTS
Switchgear Operation
Supervisory Control
Emergency Lighting and Power
Microwave



RAILROADS
Switchgear Operation
Microwave
Engine Starting
Lanterns



INDUSTRIAL
Switchgear Operation
Emergency Lighting
Engine Starting
Flashlights and Lanterns
Cordless Power Tools



MARINE
Engine Starting
Radio and Alarms
Life Boats

IRE People

(Continued from page 44A)

Rapids, Iowa, designing single-side-band transmitters for military and commercial use.

Lt. Ayers, an electrical engineer, received his commission in the Signal Corps upon graduation from Kansas State University in 1954, and served the next two years as Signal Equipment Maintenance and Repair Officer in the 1st Armored Division, Fort Hood, Tex. He operated his own television service company in Hutchinson, Kansas, from 1956 to 1958.

❖

The election of **Dr. Elmer W. Engstrom** (A'25–M'38–F'40) as President of the Radio Corporation of America was announced recently. He has been Senior Executive Vice President of RCA since 1955. In this capacity, he will have supervision of all company operations and will report to the Chairman of the Board.

In 1923, he received a B.S. degree in electrical engineering from the University of Minnesota, and in 1949, the honorary degree of Doctor of Science was conferred on him by New York University. In 1960, he was awarded an honorary Doctor of Laws degree from Findlay College, Findlay, Ohio, and in 1961, an honorary Doctor of Laws degree from Rider College, Trenton, N. J.

After graduation from the University of Minnesota, he was associated with the General Electric Company, and was assigned to engineering development work in the Radio Engineering Department. When the radio engineering and manufacturing activities were transferred to RCA in 1930, he continued as Division Engineer in charge of photophone development and design. He then took over the engineering responsibilities for RCA's broadcast receivers and continued in this field until he began the organization of a research department, the first in the fields of apparatus and systems. Later, he was responsible for radio tube research and coordinated this with the apparatus and systems work. In 1942, he became Director of General Research, and in 1943, Director of Research of RCA Laboratories. In 1945, he was elected Vice President in Charge of Research of the RCA Laboratories Division; in 1951, Vice President in Charge of RCA Laboratories Division; in 1954, Executive Vice President of that division and of Research and Engineering. In 1955, he was appointed Senior Executive Vice President.

Among Dr. Engstrom's many awards are: the silver plaquette of the Royal Swedish Academy of Engineering Sciences, the Progress Medal Award of the Society of Motion Picture and Television Engineers, the John Ericsson Medal of the American Society of Swedish Engineers, the Industrial Research Institute Medal, the Christopher Columbus International Prize in Communications, and the Medal for the Advancement of Research of the American Society of Metals. He is a member of many organizations, including the executive committee of the Defense Science Board, the Advisory Committee of the Research Division of the College of Engineering of New York University, Sigma Xi, the Research and Development Committee of the National Security Industrial Association, the Advisory Council to the Department of Electrical Engineering of Princeton University, and the Committee on Research of the National Association of Manufacturers. He is also a fellow of the IRE and the AIEE, and Vice Chairman of the Hoover Medal Board of Award for 1962.

❖

Alfred G. Holtum, Jr. (A'47–M'53–SM'56) has recently been appointed Chief of the newly formed government research and development group at Andrew Corporation. The new engineering group will concentrate its activities on development of new antenna and transmission line system concepts. He was formerly Chief Engineer of the company's California operations, specializing in telemetry antenna systems. He previously held engineering posts at Fort Huachuca, Arizona Electronic Proving Grounds, and at the Fort Monmouth Signal Corps Laboratories where he was actively engaged in the design and evaluation of ground and airborne antennas.



A. G. Holtum, Jr.

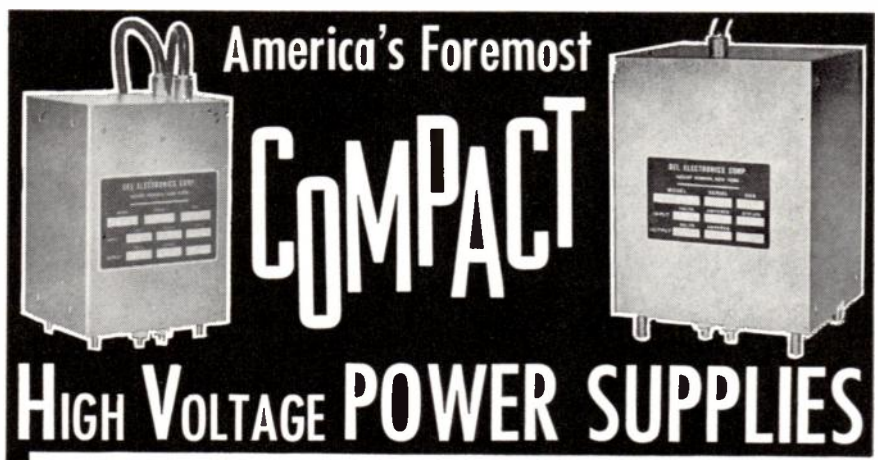
Mr. Holtum received his B.A. degree from New York University and his M.S.E.E. degree from Rutgers University.

❖

J. R. Johler (A'47–M'55–SM'55) has been named Chief of the new Applied Electromagnetic Theory Section, National Bureau of Standards, U. S. Department of Commerce. The section will be located in the Radio Systems Division at the NBS Boulder Laboratories. He will have responsibility for developing techniques for the solution of electromagnetic theory problems in terrestrial radio wave propagation by numerical analysis, analytic and asymptotic series techniques, and will direct the work to electronic computer solutions.

Since 1953, he has worked principally in the field of radio navigation systems evaluation, in low and very-low frequency radio wave propagation, and sferics (radio noise caused by lightning flashes). From 1953 to 1955, he was responsible for the theoretical prediction of the performance of certain pulse-type radio navigation systems. His early work at the Bureau laboratories dealt with instrumentation and measurement of radio noise of natural origin, especially in the very-high frequency band; with radio astronomy and measurement of cosmic radio noise; and

(Continued on page 48A)



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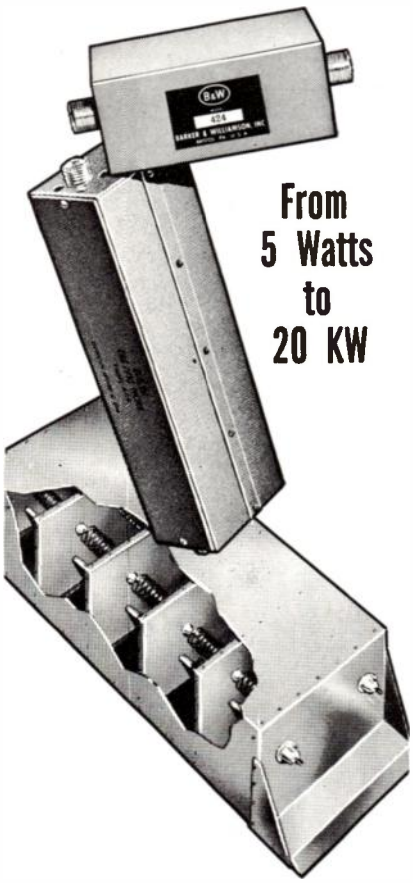
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IRE People

(Continued from page 46A)

with particular emphasis on the design and development of specialized equipment for measuring atmospheric radio noise. Before joining NBS in 1946, he had worked on the computational and engineering details of exterior ballistics at Aberdeen Proving Ground, Aberdeen, Md.

Mr. Johler received a B.A. degree from American University in 1941, and a B.S. degree in engineering from George Washington University in 1950. He has also done graduate work at the University of Colorado and the NBS Graduate School.

He is a member of the American Association for the Advancement of Science, American Geophysical Union, and International Scientific Radio Union (URSI). He has also contributed to international conferences of the Advisory Group for Aeronautical Research and Development (AGARD) of NATO.

❖

Appointment of **Paul G. Hansel** (S'39–A'42–SM'50–F'57) as Vice President of Engineering has been announced by Electronic Communications, Inc. A recognized authority on radio navigation, direction finding and meteorological electronics, he brings to ECI more than 20 years experience involving technical and administrative responsibility for the conception, planning, research, design and advanced development of major systems and product programs. He both invented and directed the development of Doppler direction finders and Doppler omnirange equipments and other devices, all of which are now coming into worldwide use for air navigation and traffic control. He holds a total of 27 U. S. and a number of British, French, Japanese, German and Swedish patents, and he is the author of many technical papers and articles.

His most recent position has been that of Chief Radio Engineer for Servo Corporation of America, a company with which he had been associated for 14 years. From 1941 through late 1947, he was a member of the Signal Corps Engineering Laboratories where he directed pioneering research and development efforts on UHF and microwave receivers, countermeasure devices and direction finding systems. In addition, while with this organization, he invented the basic operating principles embodied in all major ground-based direction finding systems employed by the U. S. Army.

A graduate of the University of Kansas, Mr. Hansel received his bachelor of science degree in engineering physics. He is a member of AIEE, and in 1957 was awarded the grade of Fellow in the IRE for his "... contributions in fields of radio navigation and direction finding."

❖

Marlin Kroger (S'49–A'50–M'55–SM'57) a 10-year Motorola veteran, has been appointed Associate Director of Research and Development, Command and Control Systems for Motorola's Military Electronics Division. He will be headquartered at the company's Systems Research Laboratory in Riverside, Calif. and will report directly to the Director of R & D for the Military Electronics Division. His appointment was announced following his completion of an 18-month assignment with the Institute for Defense Analyses (IDA), a non-profit corporation in Washington, D. C. managed by nine universities. His work during this assignment was in the Research and Engineering Support Division of IDA which provides technical assistance to the Director of Defense Research and Engineering.


M. KROGER

In his new capacity his efforts will be mainly concentrated on identification and recommendations for action in the command and control areas, program planning, direction and consulting assistance throughout the Military Electronics Division. He will cooperate closely with all Motorola command and control projects, especially those performed in the Systems Research Laboratory. In addition to these responsibilities, he will continue as a consultant on a part-time basis to the Office of the Secretary of Defense and to IDA.

Mr. Kroger is a member of Sigma Xi, Pi Mu Epsilon, Eta Kappa Nu, and Sigma Tau.

❖

The Research and Development Division of Fairchild Semiconductor has announced the appointment of **Dr. Peter Lauritzen** (S'58) and **Pierre Lamond** (M'58) to the laboratory's technical staff.

Dr. Lauritzen began his career at the California Institute of Technology where he received a B.S.E.E. in 1956. In 1958 he was awarded a M.S.E.E. from Stanford University, and in the early part of this year, his Ph.D. in Electrical engineering from Stanford. He recently returned from Zurich, where he studied at the Federal Institute of Technology under a fellowship. He has done engineering work for Minneapolis-Honeywell Regulator Company.

Mr. Lamond, a native of Paris, France, is a graduate of the University of Toulouse where he received a degree in physics in 1955. In 1961 he received a M.S. degree in communications from Northeastern University. He recently had his paper on the "Spectral Response of Solar Cells" presented at the Paris Conference on Semiconductors. He was manager of device development for Gransitron, Inc., for three years and will be working in a similar area for Fairchild.

❖

(Continued on page 50A)

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

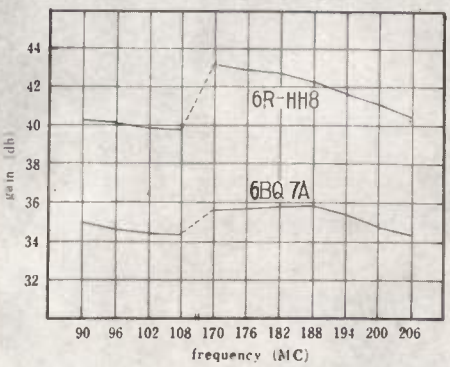
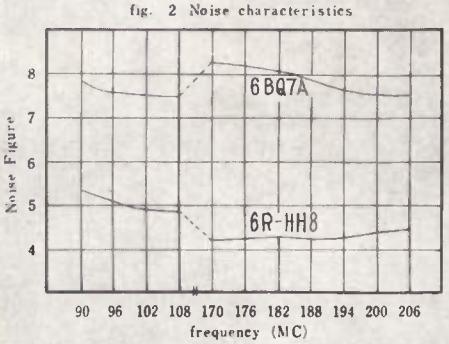
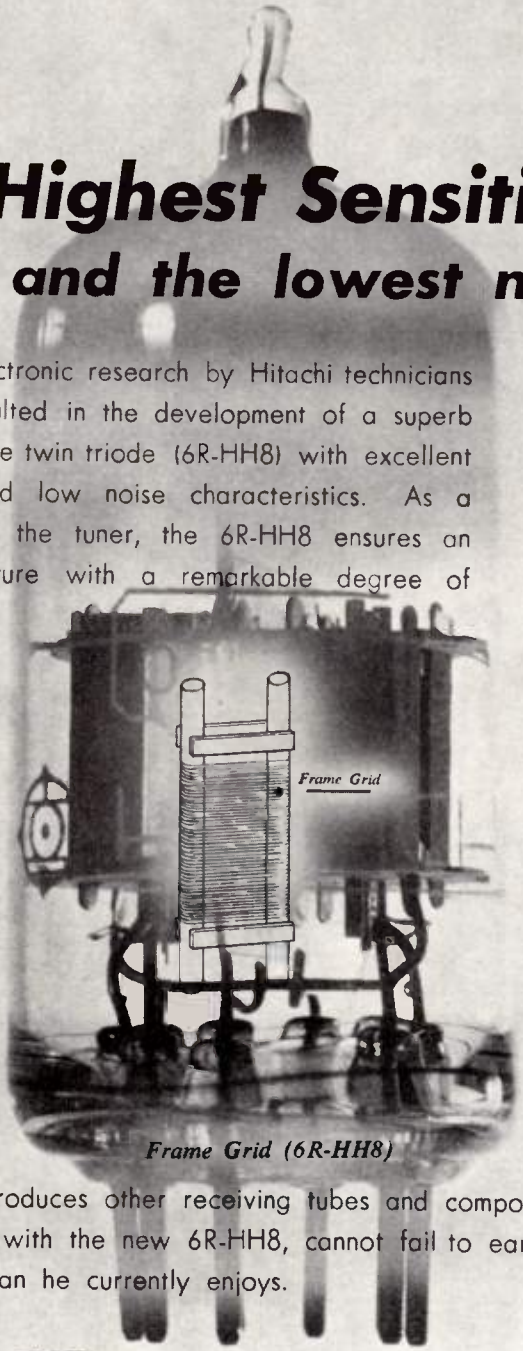


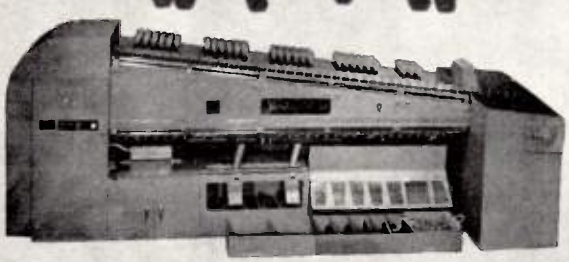
fig. 2 Noise characteristics





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.



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IRE People

(Continued from page 48A)

Dr. Bruno F. Ludovici (S'53–M'59) has joined the Hoffman Science Center in Santa Barbara, Calif., as a senior scientist. His work will involve specialized communications systems research in fields such as parametric amplifiers, masers and lasers. He recently completed his Ph.D. in electrical engineering at Stanford University, where he was a research assistant. He was born in Graz, Austria, and studied at the Technische Hoschschule in Graz and Vienna.



B. F. Ludovici

Before coming to Hoffman, Dr. Ludovici was a radio engineer with Chisholm Industries, Vancouver, B.C., and then completed his master's degree at the University of Florida. He has written seven technical papers, his specialties being plasma physics, parametric amplifiers, and microwave techniques. He is a member of the American Physical Society.

❖

The appointment of **Drs. Lee A. MacKenzie** (S'54) and **Alexander S. Gilmour, Jr.** (M'60) to the School of Electrical Engineering at Cornell University has been announced.

Dr. MacKenzie was born in Brockton, Mass., on April 18, 1934. He received the Bachelor of Electrical Engineering degree in 1957, the Master of Science degree in 1958, and the Doctoral degree in 1960, all from Cornell University. During his graduate study at Cornell, he held two General Electric Fellowships and during the summers he was a research assistant on klystron and microwave noise studies at Cornell. His Ph.D. thesis, "A Study of Noise in Self-Excited Oscillators," is a generalized study of noise in microwave oscillators. In 1960, he joined the Electrical Engineering faculty at Cornell University, and immediately took a leave of absence from 1960 to 1961. During this year he was employed at Varian Associates in Palo Alto, California, where he was concerned with experimental studies of high-perveance, hollow beam interaction. In the fall of 1961 he returned to Cornell. He is a member of Eta Kappa Nu, Tau Beta Pi, Phi Kappa Phi, and Sigma Xi.

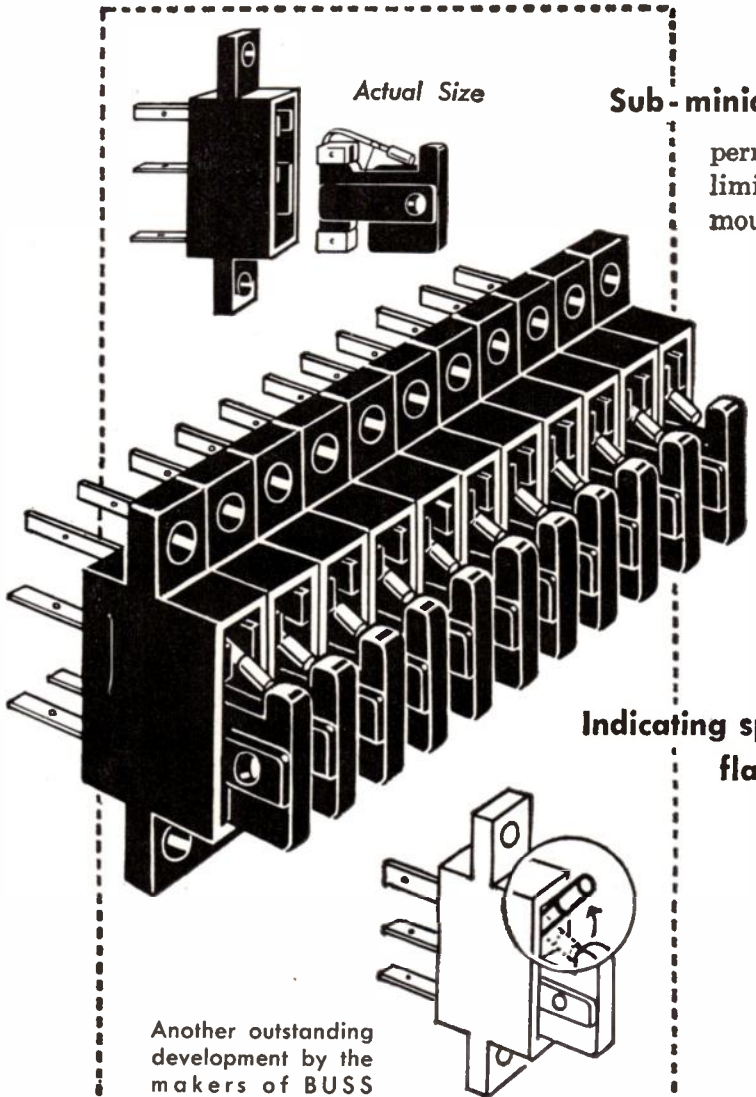
Dr. Gilmour was born in Deposit, New York on July 20, 1931. In September, 1950 he enrolled at Cornell University in the School of Electrical Engineering and received the degree of Bachelor of Electrical Engineering in June, 1955. After serving in the Navy for a period of two years, he returned to Cornell and enrolled in the Graduate School in September, 1957. He received the degree of Master of Science in Electrical Engineering in June,

(Continued on page 52A)

NEW! BUSS . . .

Signal Indicating • Alarm Activating

GMT Fuse & HLT Fuseholder



Sub-miniature design

permits multiple mounting of fuses in limited space. Fuseholders can be mounted on ¼ inch horizontal centers.

Fuse and holder combination readily adaptable for use in equipment operating at 300 volts or less, such as: communication equipment, business machines, computors, control equipment or other multiple circuit apparatus where space is at a premium.

Indicating spring flashes color-coded flag when fuse opens

to give quick, positive identification of faulty circuit.

Indicator spring also makes contact with an alarm circuit so, it can be used to flash a light—or sound audible signal on fuse panel or at a remote location.

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you'll find the type and size fuse to fit your every need . . . plus a companion line of clips, blocks and holders.



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semiconductor products news

Matches Are Made In Heaven

...they say of affairs of the heart. But when it comes to matched diode pairs and quads, they'd better be made in the laboratory and on the production line. G-E is involved in extensive work of this kind. We take choice units from the SD-150 Planar Epitaxial Passivated (PEP) signal diode line and come up with pairs and quads so closely matched that each diode in a pair (MP-1) or quad (MQ-1) is matched to every other diode to within 10 mv over a range of current from 0.1 ma to 10 ma, and to within 20 mv over a range of current from 10 ma to 50 ma at ambient temperatures from −55°C to +125°C. Co-existence at room temperature is typically matched to within 5 mv from 0.1 ma to 10 ma and to within 10 mv from 10 ma to 50 ma.

Incidentally, most ordinary matched diode assemblies are matched only over a small range of forward current, 2 to 1 or 5 to 1. G.E.'s range of forward current is 500 to 1, making it possible to use one type for a wide variety of application, such as: gates, ring modulators, bridge modulators, clamps, discriminators, choppers, and sampling circuits. MP-1 and MQ-1 are intended for high voltage, high current applications. MP-2 and MQ-2 are recommended for low voltage, low current applications where cost is a primary concern and wider voltage match is permissible. Got the sudden feeling that you need to know more? Write to us at Section 23B119.

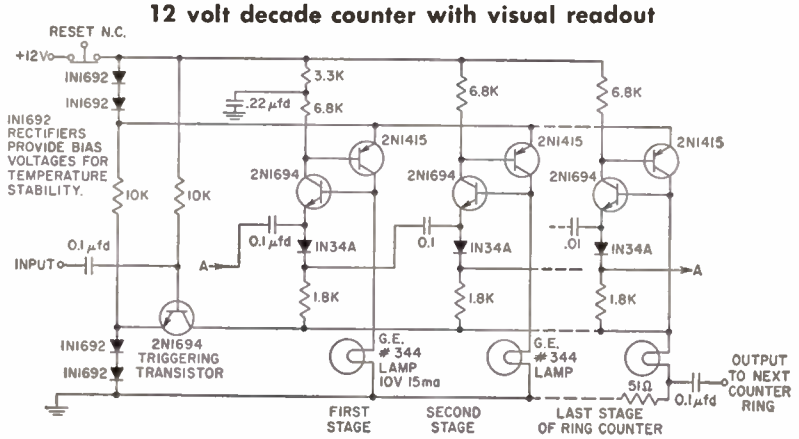
is an anatomical delight, and to solve this problem he has come up with a beaut.

Operating from only 12 volts, it uses only six components per stage and combines the counter and indicator functions in a low battery drain configuration. The 0.22 ufd capacitor makes sure that the first stage turns on after the reset button is released. None of the stages draws current except when a lamp is on. And as many stages as desired can be included in a ring.

There is an added anatomical advantage in the circuit under discussion. The G-E 2N1694 shown as the triggering transistor represent a real "best buy" for this application since the voltage, I_{CO} and h_{FE} specifications assure trouble-free performance at a truly modest price.

New controlled conductance PEP signal diodes in glass subminiature and microminiature packages are here. Guaranteed higher forward voltage drop at 100 and 200 μamps combined with guaranteed lower forward voltage drop at 1-20 ma with the forward voltage drop specified at six current levels make this important news even more important. See your District Sales Manager right away.

12 volt decade counter with visual readout



The "PEP" line of planar epitaxial passivated transistors continues to grow... and the 2N914 NPN Silicon switch now offers you unprecedented circuit design flexibility with low V_{CE} (sat) and total switching speed of 80 nanoseconds. Ask your District Sales Manager.

Any questions? Write us at Section 23B119 General Electric Company, Semiconductor Products Department, Electronics Park, Syracuse, New York. In Canada: Canadian General Electric, 189 Dufferin St., Toronto, Ont. Export: International General Electric, 159 Madison Avenue, New York 16, New York.

The Anatomy of A Circuit

Looking for a decade counter or shift register with visual readout that doesn't need some sort of buffer stage between the counter and the indicator? Isn't everybody? Been frustrated by the high voltage required by neon type lamps and the high current and power required by incandescent lamps? Isn't everybody? Everybody but our boy Dick Stasior, application engineering hero distinguished by several oak leaf clusters on his transistor manual, that is. To him a circuit

GENERAL ELECTRIC

IRE People

(Continued from page 50A)

1959, and completed the requirements for the Ph.D. degree in September, 1961. His thesis, "The Velocity Distribution in a Velocity-Modulated Electron Beam from a Shielded Pierce Gun," was an analytical and experimental treatment of the velocities in modulated Brillouin beams. During his graduate study, he served as a research assistant and received the John McMullen Graduate Scholarship, the Sperry Gyroscope Fellowship, and the Republic Aviation Fellowship. He is a member of Sigma Xi.

❖

H. Tyler Marcy (M'46) has been promoted to Vice President of the General Products Division of International Business Machines Corporation, White Plains, N. Y. In his new position, he will be responsible for product development for the division—one of two divisions in the company handling product development and manufacturing of commercial data processing equipment. He was formerly manager of the IBM development laboratory in Poughkeepsie, N. Y., a part of the Data Systems division. He joined IBM in 1951 as a technical specialist at the Airborne Computer Laboratory the company then operated in Vestal, N. Y.



H. T. MARCY

A graduate of the Massachusetts Institute of Technology, Mr. Marcy holds a master's degree in electrical engineering. Before joining IBM, he was associated with M. W. Kellogg in Jersey City, N. J.

❖

Joseph T. McNaney (SM'53), an engineering consultant and former technical director of Stromberg-Carlson in San Diego, was selected to receive the E. H. Rietzke Achievement Award of 1961 for his invention of the Charactron tube. This award is sponsored by the Capitol Radio Engineering Institute in Washington, D. C.



J. T. MCNANEY

The Charactron tube invention, means of forming the cross section of an electron beam in the shape of message characters, was patented in 1942 primarily for use in a high speed teletype communications system. In 1953 it was selected by the Lincoln Laboratories of the Massachusetts Institute of Tech-

(Continued on page 54A)

Sanborn® 7-Channel FM tape system for $6800* complete



uses interchangeable FM and direct record/reproduce electronics entirely contained in 7″ x 19″ panel space

COMPARE PERFORMANCE, PRICE PER CHANNEL

Here is the ideal combination of high performance *and economy* in a 7-channel, 4-speed system that meets IRIG Telemetry Standards. *Versatility* is another advantage. The Model 2000 system uses interchangeable Sanborn FM or direct record/reproduce electronics — all solid-state, in 7″ of panel space — and you can have any combination of direct and FM channels simply by changing circuit cards. Recording capability may be extended beyond the system's minimum input levels through the use of Sanborn "850" and other compatible amplifiers.

The Model 2000 Magnetic Data Recorder has four speeds and uses standard ½-inch tape on 10½-inch reels. All controls are on the front, and several convenience features are included: an integral FM Alignment Meter that eliminates the need for electronic counters, an automatic squelch, a tape footage counter, and provision for using one channel for flutter compensation.

Complete details are available from Sanborn Sales-Engineering Representatives in principal cities throughout the U. S., Canada and foreign countries.

**Price FOB Waltham, Mass., in Continental U. S. A.; subject to change without notice. State and local taxes must be added where applicable.*

(Specifications subject to change without notice)

SPECIFICATIONS

Input ± 2.5 V into 10,000 ohms, single ended, adjustable.

Output ± 2.5 V into 1,000 ohms or more, single ended; level, position adjustable.

Bandwidths (Max)

Speed	*FM*	*Direct*
3¾″/sec	0-625 cps	50-6,250 cps
7½″/sec	0-1,250 cps	50-12,500 cps
15″/sec	0-2,500 cps	50-25,000 cps
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(100% modulation on FM = ± 40% carrier deviation)

Linearity Max dc nonlinearity: 0.5%

Drift ± 0.5% of full scale for 10 V power line change, 10°C ambient temperature change, or for 24 hours at constant power line voltage and ambient temperature.

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Direct: 40 db at all speeds.

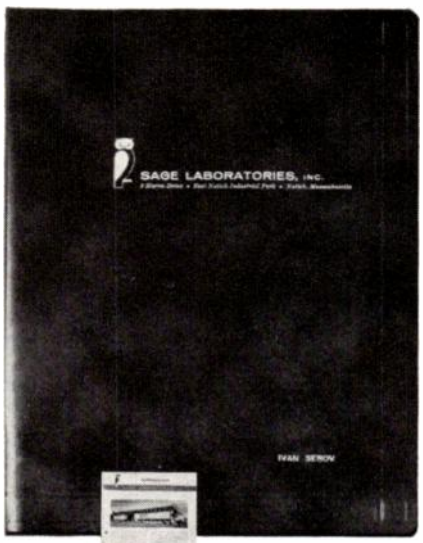
FM: 40 db RMS at 30″/sec and 15″/sec; 35 db RMS at 7½″/sec; 33 db RMS at 3¾″/sec.



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IRE People

(Continued from page 52A)

nology for use in the SAGE System. Since his original invention he has recorded at least 70 additional improvements to the basic idea in the form of U. S. patents.

Mr. McNaney received his education in electrical engineering with the Capitol Radio Engineering Institute, and is a duly registered Professional Electrical Engineer in California.

❖

Dr. Gerhard K. Megla (SM'60) has joined the Hoffman Science Center in Santa Barbara, Calif., as a senior scientist. He will work in association with the Director of Engineering for Hoffman's Military Products Division. He came to this company from the Martin Co., Orlando, Fla., where he was a senior designer. His previous experience includes teaching and research work in Germany, specializing in electromagnetic wave theory, wave propagation, antennas and radio communications systems.



G. K. Megla

A graduate of the University of Dresden, Germany, Dr. Megla has written three books, delivered 17 papers in his field, and has been granted 7 patents.

❖

Appointment of **John D. Moynihan** (S'43–A'49–SM'55) as product specialist for high reliability capacitors at the Sprague Electric Co. was announced recently. He has rejoined Sprague Electric after a 5-year period of service as chief engineer and later sales representative for the Ferroxcube Corp. of America at Saugerties, N. Y. He had previously been carrier current development engineer with the former Sprague Moduline Department. He has been associated also with Westinghouse Electric Corp. in design engineering capacities at its carrier and relay department in Newark, N. J. and in the central station department in Pittsburgh, Pa.



J. D. Moynihan

Mr. Moynihan holds a degree of B.S.E.E. from the University of Oklahoma and a M.S. degree from the Stevens Institute of Technology. He also engaged in post-graduate work at the University of Pittsburgh. He is a member of the American Institute of Electrical Engineers.

❖

Thomas H. Miller (SM'59) has joined Frederick Research Corporation in a senior capacity as Director of Engineering. For the past five years he has been with Vitro Laboratories as Chief of the Radio Frequency Interference Section. Prior to that he was Assistant Chief Engineer of International Electronics Engineering, Inc. During his five years at I.E.E. he became widely known for his work and contributions in the field of RFI. He is particularly known for his work on the Navy's Interference Specification MIL-I-16910 in 1952.



T. H. Miller

Mr. Miller received his B.S.E.E. degree from Johns Hopkins University in 1949.

❖

The appointment of **Dr. Leon S. Nergaard** (A'29–M'38–SM'43–F'52) as Director of the Microwave Research Laboratory at RCA's David Sarnoff Research Center, Princeton, N. J., has been announced. He has been an Associate Laboratory Director since early 1960, and was the 1960 recipient of RCA's David Sarnoff Outstanding Achievement Award in Science for "his basic and practical contributions to the science of microwave electron tubes and thermionic cathodes."

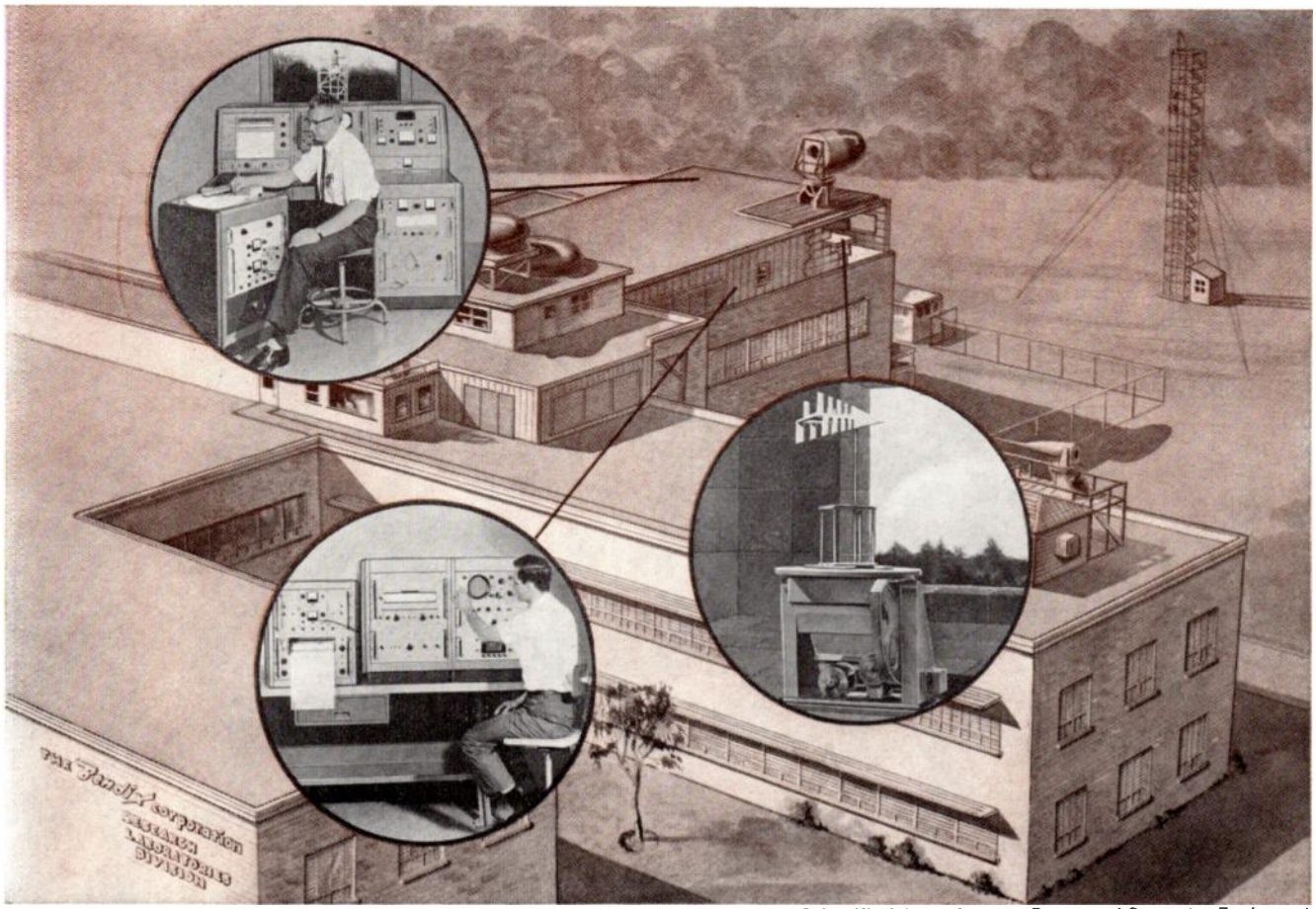


L. S. Nergaard

He has been with the Radio Corporation of America since 1933. Upon his graduation from the University of Minnesota in 1927, he joined the General Electric Company as a research engineer in the field of vacuum tubes. He received the M.S. degree from Union College, Schenectady, N. Y., in 1930, and returned to the University of Minnesota as a teaching assistant in physics while completing his studies for a doctorate, which was awarded to him in 1935. As a research engineer with RCA at Harrison, N. J., during the 1930's, he was responsible for such basic advances as the application of ultra-high-frequency techniques to power tube design, the development of the wideband television transmitter tube, and the joint development, with Andrew V. Haeff, of the inductive output tube which laid the foundation for the later development of klystrons and many other microwave tubes. In 1942, he was transferred to the newly-established RCA Laboratories at Princeton, N. J., where he was a leader in microwave radar tube development during World War II. In the post-war era, he conducted basic studies of oxide cathodes and achieved results which led directly to major improvements in the life and reliability of virtually all types of electron tubes in use today. One of his most recent

(Continued on page 56A)

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Bendix uses Scientific-Atlanta equipment, obtains optimum designs in minimum time

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IRE People

(Continued from page 54A)

contributions has been the successful redesign of the high-power circuit for transmitters used in the Nation's Ballistic Missile Early Warning System.

Dr. Nergaard holds numerous patents and is the author of many basic papers relating largely to tube theory and design. He is a Fellow of both the American Physical Society and the IRE, and a member of the American Association for the Advancement of Science.

❖

Thomas J. Norman (A'55–M'55–SM'59) has been named chief engineer at National Data Processing Company, a department of Remington Rand Univac in Dallas. He will be responsible for the coordination of engineering activities in the design and development of the firm's MICR bank automation systems and optical character recognition equipment. He joined NDP in September, 1959, as project manager on the NDP Magnetic Ink Character Reader. Recently he has been project manager of the optical character recognition card punch equipment. Prior to his association with NDP he was a senior aero-physics engineer with General Dynamics where he worked on special-purpose, air-borne computers.



T. J. Norman

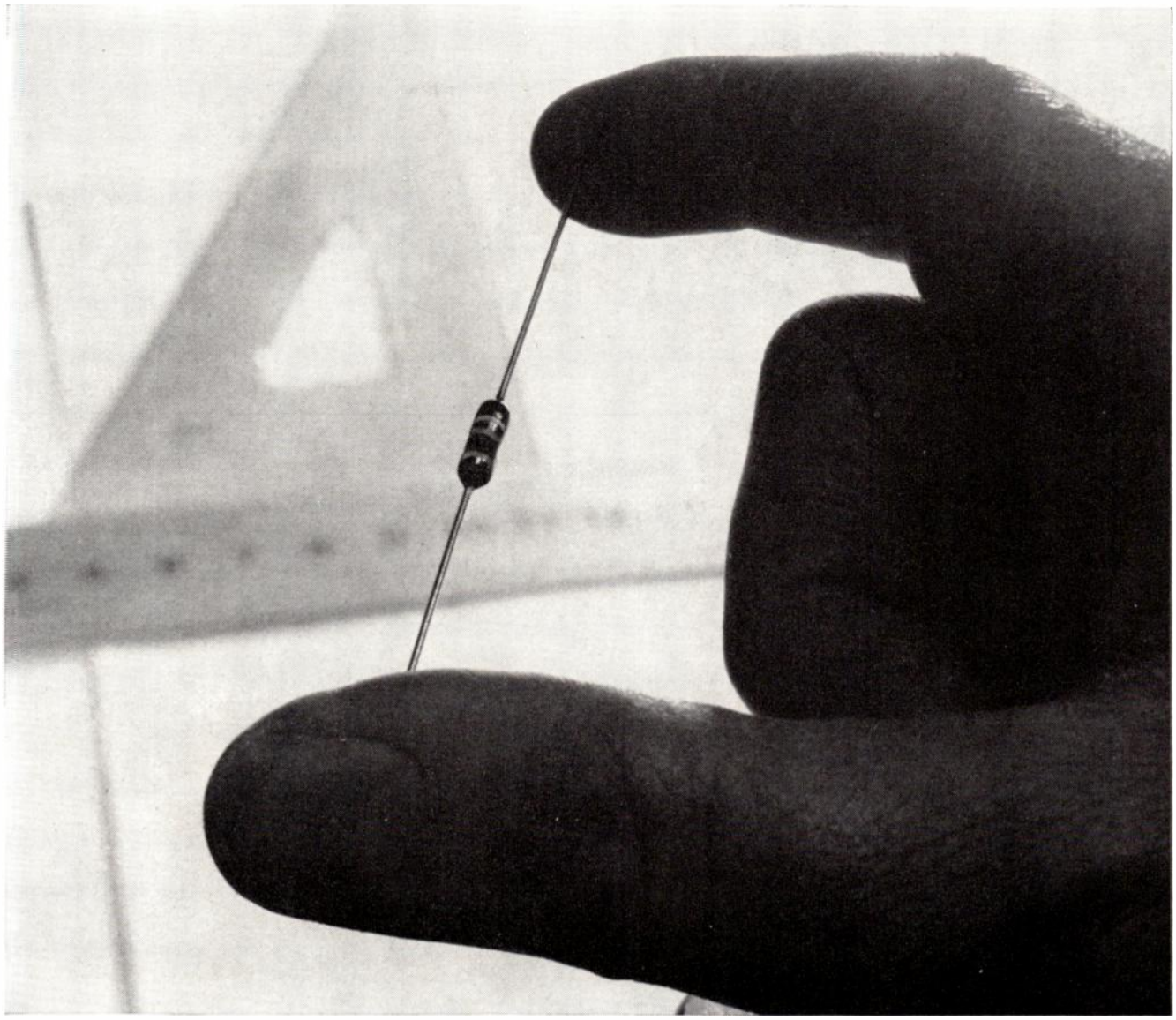
Mr. Norman was graduated from the University of Texas in 1949 with a B.S. in electronics engineering. He is a member of the Texas Society of Professional Engineers, Tau Beta Pi and Eta Kappa Nu.

❖

A distinguished engineering educator and electrical engineering researcher has been appointed to the faculty of the newly opened School of Engineering at the University of California, Santa Barbara. **Dr. Philip F. Ordung** (S'40–A'43–M'48–SM'49–F'59) Professor of Electrical Engineering at Yale University will become professor of engineering on the Santa Barbara campus effective July 1, 1962. He has been at Yale since his graduate studies, earning both the M.S. degree and Ph.D. degree in engineering. Joining the faculty, he received his full professorship in 1957.

His professional activities have included a notable record of engineering writings including 19 published articles, co-authorship of two textbooks, "Microwave Theory and Techniques" and "Microwave Principles." He has written numerous technical research reports and is a regular participant in professional meetings in the electrical and radio engineering fields. In the research field he also managed several large research contracts including

(Continued on page 58A)



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Type	Wattage	Resistance (ohms)		Nominal Dimensions	Design Tolerance Full power, 70°C.
		min.	max.		
C07, Mil Style RL07	¼	51	150K	.250"x.090"	5% plus purchase tolerance
C20, Mil Style RL20	½	51	150K	.375"x.138"	
C32, Mil Style RL32	1	51	470K	.562"x.190"	
C42S, Mil Style RL42	2	10	1.3 meg	.688"x.318"	

CORNING
Electronic Components

IRE People

(Continued from page 56A)

those with the Army Signal Corps, the Navy and General Electric Company. As a research specialist, however, he retains an active interest in teaching, both on the undergraduate and graduate level. As a result of his outstanding achievements, Dr. Ordung was elected a fellow of the IRE. He also served as a fellow of the Timothy Dwight College at Yale.

❖

Ralph J. Peduto (S'58–M'58) has been named Product Manager for oscillators and frequency control systems at the Electronics Division of the Bulova Watch Company, Inc. Prior to rejoining the Electronics Division, located in Woodside, N. Y., he had been associated with Aeroflex, Inc. as a project engineer. Previously he had two years service with Bulova as a project engineer in the filter department.



R. J. PEDUTO

An electrical engineering graduate of City College of New York, Mr. Peduto earned a master's degree in electrical engineering from New York University. Presently he is on the staff of New York Institute of Technology, teaching electronics. He is a member of the IRE Audio Engineering Society, and Eta Kappa Nu.

❖

John M. Phelps (SM'59) Captain, USN (Ret.), who retired from the Navy after 27 years of active commissioned service, has joined Granger Associates, Inc., Palo Alto electronic systems firm, as Assistant to the President. His last active duty assignment was as Director of Research and Development, Defense Communications Agency. In his new position, he will assist in planning and organizational development activities. In his DCA post in 1960 and 1961, he was responsible for the conduct of the agency's R/D activity and for reviewing and coordinating R/D programs of the Army, Navy and Air Force.



J. M. PHELPS

Previously, from 1956 to 1960, he served as Commanding Officer and Director of the Navy Electronics Laboratory at San Diego, Calif., where he supervised programs involving radio, radar, sonar, human factors, weapons effects, tactical data systems, oceanography and supporting functions. Between 1946 and 1956 he was associated with the Bureau of Ships as Head of the Components Section and the Systems Branch, and as Director of the Electronic Design Division. He also served in naval shipyards as Electronics Officer, Industrial Relations Officer and Repair and Shipbuilding Superintendent.

Captain Phelps was graduated from the U. S. Naval Academy with the B.S. degree in 1934, and from the Naval Postgraduate School course in Electronics Engineering in 1942. He also completed the Advanced Radar Course at Massachusetts Institute of Technology in 1942. He is a member of the Armed Forces Communications and Electronics Association and a member of the U. S. Naval Institute.

❖

The appointment of **William J. Ronis** (M'59) as Contract Relations Manager of the Microwave and Antenna Department of TRG, Inc. has been announced. Experienced in the field of microwave antenna engineering and sales, he will direct all military and commercial customer activities. TRG specializes in antenna research and production, and in the research and production of millimeter microwave components. His most recent affiliation was with George Gostenhofer Associates as a microwave sales engineer. Prior to that he was a staff member at the Mitre Corporation and Gabriel Electronics, where his activities were concerned with antenna and microwave component development in a broad frequency spectrum.



W. J. RONIS

Mr. Ronis received his Bachelor of Science Degree in Electrical Engineering at Northeastern University in 1956. In addition he has taken a number of graduate courses in Mathematics, Physics and Electrical Engineering at Northeastern.

❖

Wheeler Laboratories, Inc., a subsidiary of Hazeltine Corporation, has announced the election of **Ned Spencer** (A'47–M'54–SM'59) of Port Washington, N. Y., as a Vice President.

A graduate of Massachusetts Institute of Technology, he joined Wheeler Laboratories in 1948 and has been in charge of the Antenna Development Laboratory at Smithtown, N. Y. since 1947. While at Wheeler, he has specialized in microwave and antenna development. He has recently been associated with major projects in the Nike-Zeus program and with communication satellite



N. SPENCER

(Continued on page 60A)



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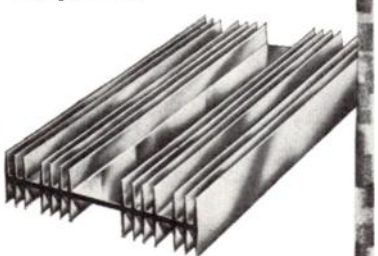
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(Continued from page 58A)

antennas. In 1960, he was appointed assistant chief engineer. During the past four years, Mr. Spencer has presented lectures on antenna fundamentals at many colleges.

❖

Karl E. Sterne (S'52–A'55–M'60) has been named to the position of Sales Manager of Precision Instrument Company. He comes to Precision from the Autonetics Division of North American Aviation, where he was Marketing Supervisor for military requirements from 1960 to 1961. He was Computer Sales Manager for The Berkeley Division of Beckman Instruments from 1956 and 1960, and also worked at the Lawrence Radiation Laboratory in Berkeley, from 1954 to 1956 as a Design Engineer for the development of nuclear particle accelerators.



K. E. STERNE

As Sales Manager, Mr. Sterne will be in charge of the domestic sales operation. He holds a B.S. degree in physics from the University of Michigan, granted in 1948, and he received a M.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1954. He is a member of the American Marketing Association.

❖

Hazeltine Technical Development Center, Inc. has named **William S. Sturgeon** (M'48–SM'57) as its Chief Engineer. Formerly assistant manager of the Greenlawn (New York) Laboratory of Hazeltine Electronics Division, he has more than 20 years' experience in the electronics field. Before joining Hazeltine in 1946, he served as a Lieutenant Commander in the U. S. Navy in charge of design and development of electronic identification equipment. His diversified career includes design, development and installation responsibilities for undersea, ship-board, ground, airborne and missile-borne electronics systems.



W. S. STURGEON

Mr. Sturgeon, who holds a B.S.E.E. Degree from New York University, is an associate member of the American Institute of Electrical Engineers and a member of the Advisory Commission on Electronic Technology of the New York City Community College.

❖

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Louis Weinberg (M'52–SM'56–F'60) has been named Vice President, Information Processing, of Conductron Corporation, Ann Arbor, Mich., according to a recent announcement. Internationally recognized for his contributions to the field of network theory, his published work covers such varied subjects as noise through nonlinear devices, partial differential equations, matrix theory, and combinatorial topology. He is most widely known for his fundamental work in network synthesis. He has presented an invited series of lectures at many universities, both in this country and abroad, among them being lectures at the University of Manchester, England, in 1955, and the University of Wisconsin in 1956, and in 1960 at Chalmers University of Technology, Gothenburg, Sweden, and at the Instituto Elettrotecnico Nazionale, Turin, Italy.



L. WEINBERG

In 1941, he graduated with a B.A. degree from Brooklyn College as the first-ranking scholar in his class. He then received an M.S. in Engineering Sciences and Applied Physics from Harvard University in 1947, and the Sc.D. degree in Electrical Engineering in 1951 from the Massachusetts Institute of Technology. He was an instructor in Electrical Engineering at the Massachusetts Institute of Technology from 1947 to 1951. During this period he also performed research on statistical communications and network theory at the Research Laboratory of Electronics. Prior to this in 1942–1943 he was a radio engineer with the Office of the Chief Signal Officer. He left this position to join the Air Force where he remained until 1946, serving as a radar and electronics officer and as Technical Advisor to the American Embassy at The Hague.

In 1951, he joined the Hughes Aircraft Company where he worked in such fields as fighter-bomber tactics, noise theory, transistors, logical design of digital computers, and network theory. During the last five years of his employment there, he was in the Hughes Research Laboratories as Head of the Communications and Networks Research Section. In this capacity, he supervised research on multipath communication, feedback and control theory, parametric amplifiers, linear programming, plasma physics, and applied network theory. He also held part-time academic positions while at Hughes. For two years, 1952–1954, he was a lecturer in Engineering at the University of California at Los Angeles where he introduced the first graduate course in network synthesis; he was a visiting Associate Professor of Electrical Engineering at the University of Southern California in 1955–1956, where he helped start the doctorate program in Electrical Engineering. He also held the post of visiting Professor of Electrical Engineering at the California Institute of

(Continued on page 62A)

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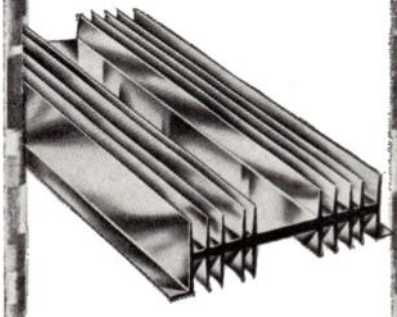
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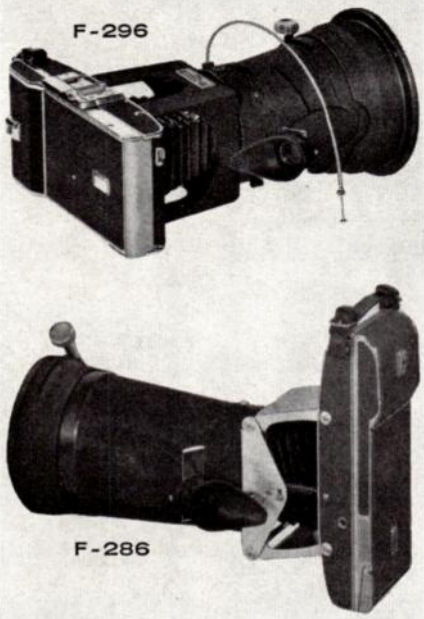


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(Continued from page 61A)

Technology for the two years 1958–1959.

Dr. Weinberg has received a number of honors, among them being the New York State Scholarship (1937–1941) and the Dean Bildersee Scholarship in 1941. He is a member of Commission VI of URSI and represented the United States in solid-state circuit theory at the General Assembly of URSI that took place in London in September 1960. At this General Assembly he was elected International Vice President of URSI Commission VI. He is a member of Sigma Xi and Proplyaea, the New York Academy of Sciences, and is a Fellow of the IRE and a Fellow of AAAS. He was Vice President (in 1959) and President (in 1960) of the Hughes Branch of the Research Society of America (industrial affiliate of Sigma Xi).

❖

Frank H. Williams (S'47–M'48–SM'54) has been named Chief Engineer and Executive Vice President of Wheeler Laboratories, Inc., a subsidiary of Hazeltine Corporation. Since 1955, he has been assistant chief engineer in charge of the company's microwave laboratory in Great Neck, N. Y. After joining Wheeler Laboratories in 1949, he served as project supervisor on numerous study, design and development projects, including work on scanned antenna arrays, microwave filters and multiplexers and an experimental, high resolution, airborne radar. He also worked on various microwave components for the NIKE Ajax and NIKE Hercules tracking radars. Recently, he has been project supervisor for a study program connected with precision microwave guidance systems for advanced Air Force missile systems.


F. H. Williams

Following his graduation from Polytechnic Institute of Brooklyn in 1947, Mr. Williams became a Hazeltine engineer and participated in the development of DME beacon plumbing, FM studio-transmitter link equipment and an airborne intercept analogue computer. He also taught at Polytechnic Institute for several years after receiving his B. E. E. degree. He is a licensed professional engineer in the State of New York, and a member of Eta Kappa Nu. During World War II, he served in the U. S. Navy as an instructor at the Radar School at the Naval Research Laboratory.

❖

Harold V. T'Kach (A'51–M'53) has resigned as President of Mobile Radio Engineering, Inc., Minneapolis, to open a private consulting service for upper-midwest government and industrial communications systems. Since its inception in 1948, T'Kach's firm specialized in the installation and maintenance of two-way communications systems for police, fire, other public safety, and industrial use. His new advisory service can be used by governmental departments and private industry throughout the area, which require engineering assistance when two-way communications systems are installed initially or need modernization. His service will include surveys as to communications needs, preparation of product specifications and supervision of installation.


H. V. T'Kach

Mr. T'Kach is a member of the Minnesota Society of Professional Engineers and the National Society of Professional Engineers.

❖

Merrimac Research & Development, Inc., announced the appointment of **Frank J. Weber** (M'55) as a development engineer. Formerly an ITT engineer, he will be working in the area of parametric devices.


F. J. Weber

He was born in New York City in 1931, and received his first electronics training at the U. S. Marine Corps Signal School in San Diego, Calif. He received the B. S. degree in physics from Purdue University in 1954, and the M.S. degree in electrical engineering from Newark College of Engineering in 1958. Following his graduation from Purdue, he went to work for ITT's Federal Laboratories on electronic countermeasures projects involving a strong proportion of microwave circuits. Some of these were the AN/APD-4 and other reconnaissance systems; false target repeaters for active countermeasures; and r-f sections of the AN/ASQ-18 system. In 1959, he transferred to ITT's space and missile systems lab, where he designed and developed microwave circuits for the LOBAR system. Another missile-system project with which he was involved was a feasibility study of a precision missile-tracking system using parametric techniques. Mr. Weber will pursue this relatively recent electronics specialty at Merrimac.

❖

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Current Rating	30 Amps	60 Amps
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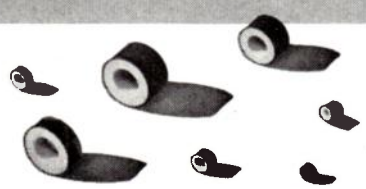
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Professional Group Meetings

Aerospace and Navigational Electronics

Boston—November 28

Guided Tour of the Boston Air Route Traffic Control Center, W. Henderson, FAA, Tower Bldg., Logan Airport, Boston.

Antennas and Propagation

Boston—November 29

"Radar Exploration of Venus," Dr. G. Pettingill, Lincoln, Millstone-Hill Site.

"Dipole Antennas in Conducting Media," Dr. K. Iizuka, Harvard Univ.

Dayton—October 4

"1000-foot Spherical Reflector Antenna," E. L. Bouche, Air Force Cambridge Research Labs.

San Francisco—November 8

"Some Aspects of Sandwich Wire Antenna Design," G. J. Koloboff, Dalmo Victor Co., Belmont, Calif.

Washington D. C.—November 21

"Space Rendezvous," H. Reuter and P. Hacker, Westinghouse Electric Corp., Baltimore.

Antennas and Propagation Microwave Theory and Techniques

Columbus—November 30

"The Generation of Optical Harmonics by a Ruby Laser," Prof. C. W. Peters, Univ. of Michigan, Ann Arbor.

Orange Belt—September 28

"Recent Developments in Micro Ferrite Devices," A. Clavin, Rantec, Calabas, Calif.

Philadelphia—September 18

"Plasma Amplifiers," L. W. Roberts, Metcom Inc., Salem, Mass.

Social hour sponsored by J. H. Electronics preceded the dinner and meeting.

Audio

Boston—October 26

"FM Stereo Reception," D. Von Recklinghausen, H. H. Scott, Inc.

Chicago—November 14

"Seeing with Sound Waves," H. Karplus, Armour Res. Foundation, Chicago, Ill.

Dayton—October 4

"Audio Applications," E. Meagher, Amperex Corp.

San Diego—November 16

"A Discussion of Some Novel Audio Amplifier Circuits," L. Dalzell, Pacific Telephone Co.

Automatic Control

Boston—November 30

"MH-1, A Computer-Operated Mechanical Hand," H. Ernst, MIT.

Boston—October 24

"Dynamic Behavior of R&D Organizations and Management by Feedback Control Techniques," R. B. Wilcox, C. M. King and W. H. Phoenix; RCA, Burlington, Mass.

Panel discussion after three speakers gave papers.

Chicago—December 5

"A Servo Controlled Aircraft Arresting Gear," B. Johnson, American Machine and Foundry, Niles, Ill.

Los Angeles—November 14

"Lyapunov's Second Method in the Analysis and Design of Nonlinear Control Systems," Dr. M. Margolis, Aerospace Corp., El Segundo, Calif.

Pittsburgh—November 27

"Systems Engineering from an Industrial Viewpoint," H. Chestnut, General Electric Co., Schenectady.

Bio-Medical Electronics

Boston—November 8

"Potential Physical Hazards of Microwave Radiation," R. L. Carpenter, Ph.D., Tufts University, Medford, Mass.

Boston—October 17

"Physical Aid for Life Support," Dr. A. T. Kornfield, Biosearch Co., Boston.

Columbus—November 9

"Blood Pressure Measurements," Dr. G. H. Hoover, Ohio State University, Columbus.

"Cardiac Catheterization," Dr. J. Tomashefski, Ohio Tuberculosis Hospital, Columbus.

Los Angeles—November 16

"The Instrumentation of Muscle Activity," J. H. Platt, M.D., S. Elazar, Meditron Co., and Robert B. Pearson, M.D., Loma-Linda School of Medicine.

(Continued on page 66A)



8036 SMALL, RUGGED CERAMIC HYDROGEN THYRATRON SAVES VALUABLE SPACE

Tung-Sol leads the way with a ceramic Hydrogen Thyratron that fills an important design need. An electrical equivalent of the popular Tung-Sol 5949A — only one third tube volume is required by this new member of the family.

Tung-Sol ceramic Hydrogen Thyratron 8036 has rugged environmental ratings. It is designed for flange mounting with flexible connectors to achieve a solid mounting with loss-free terminations. Grid connection is made to the flange through the grid ring clamp.

For full technical data, consult your Tung-Sol representative or write: Tung-Sol Electric Inc., Newark 4, N.J. TWX:NK193.



HYDROGEN THYRATRON 6587A

6587A, a glass thyratron, is a direct plug-in replacement for Tung-Sol 5C22. Valuable inches of overall height are saved by means of the ring-disk type of construction, which also provides the advantages of external (cool) anode and lower lead inductance. It is rated for higher voltages with higher currents than prototype tubes. Grid connection can be made through the grid ring or through the tube base pin. An internally-connected hydrogen reservoir promotes long life.

	8036	6587A	5C22
Overall height, Max	3.75"	6.	8.75
Peak forward voltage	25. KV	18.	16.
Peak current	500. Amps	365.	325.
Peak Pulse Power (Delivered to the load)	6.25Mw	3.25	2.6

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Professional Group Meetings

(Continued from page 64A)

Philadelphia—December 7

"Automatic Monitoring of Hospital Patients" (Symposium), Dr. Thorner, Univ of Pa.; Mr. Kaslow, Dekker Corp.; Dr. Winters, Temple U.; and Mr. Singler, AE.

San Francisco—November 21

"Programmed Learning: Some Experimental Findings," Dr. D. McConnell, Britannica Films Inc., Menlo Park.

Broadcasting

Omaha-Lincoln—November 11

"Open House, KTIV and KVTV," Chief guides: A. Smith, KTIV; M. Shadbolt, KVTV; B. Engelhardt, L. Scott.

Circuit Theory

Philadelphia—November 15

"Element Value Solutions for Networks with Limited Access," Dr. R. S. Berkowitz, Moore School of Electrical Engineering, Philadelphia.

Philadelphia—October 18

"The Concept of Distribution in the Study of Linear Systems," Dr. A. Papoulis, Polytechnic Institute of Brooklyn.

Communication Systems

Los Angeles—October 31

"Error Free Data Transmission Using the Bose-Chaudhuri Code," C. Lindholm, Rand Corp., Santa Monica, Calif.

Northern New Jersey—November 15

"NORAD—Communications and Electronics," D. P. Gibbs, NORAD, Colorado Springs, Colo.

Northern New Jersey—September 7

"Design—Problem and Techniques: The Future of Worldwide Communications Systems," Dr. H. Busignies, ITT Corp., N. Y.

This being the first meeting of the new chapter, Mr. G. Karger, PG Coordinator, delivered a brief address on behalf of the Section.

Oklahoma City—November 27

"Oklahoma Civil Defense Communications Network," J. Canada, Oklahoma State Dept. of Civil Defense.

Rome-Utica—November 21

Meeting devoted to planning Chapter Activities for 1961–1962 season.

Toronto—December 11

"Communications, Navigation and Flight Control System for a Modern Transport Aircraft," J. MacQuarrie, Collins Radio Co., Canada.

Toronto—November 27

"Visual Communications," A. C. Stewart, Automatic Electric Sales, Ltd., Canada.

Washington—November 27

"Microwatt—Microelectronics," Dr. W. W. Gaertner, CBS Labs., Stamford, Conn.

Communications Systems Vehicular Communications

Omaha-Lincoln—November 2

"Data phone and Other Advanced Telephone Equipment."

Tour: Northwestern Bell Telephone Co.

Component Parts

Dayton—October 4

"The Ruby Laser," M. R. Curre, Hughes Research Labs., Los Angeles.

Philadelphia—November 14

"Thin Films in Digital Computers," G. Fedde, Remington Rand Univac, Blue Bell, Pa.

Short tour of Univac Computation Center and display of early computers prior to and following meeting. Tours conducted by R. J. Klein, T. Anglim and E. Minnett of Univac.

(Continued on page 68A)

Sub-Miniature Indicator Lights

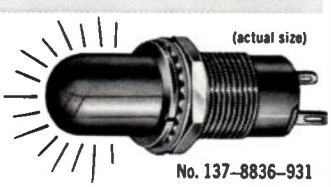
Conform to applicable Military Specifications.

Mount from FRONT of Panel in 15/32" Clearance Hole

NEON



T-2



(actual size)

No. 137-8836-931

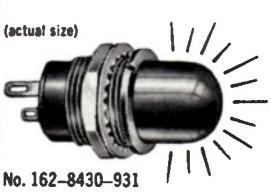
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(actual size)

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Assemblies conform to MS25256

Accommodate T-1-3/4 Incandescent bulb with midget flanged base, in voltages ranging from 1.3 to 28 (the 6 V. and 28 V. conform to MS25237).

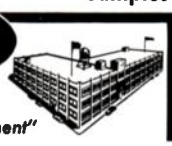
For complete data request Brochure L-156E.



T-1¾

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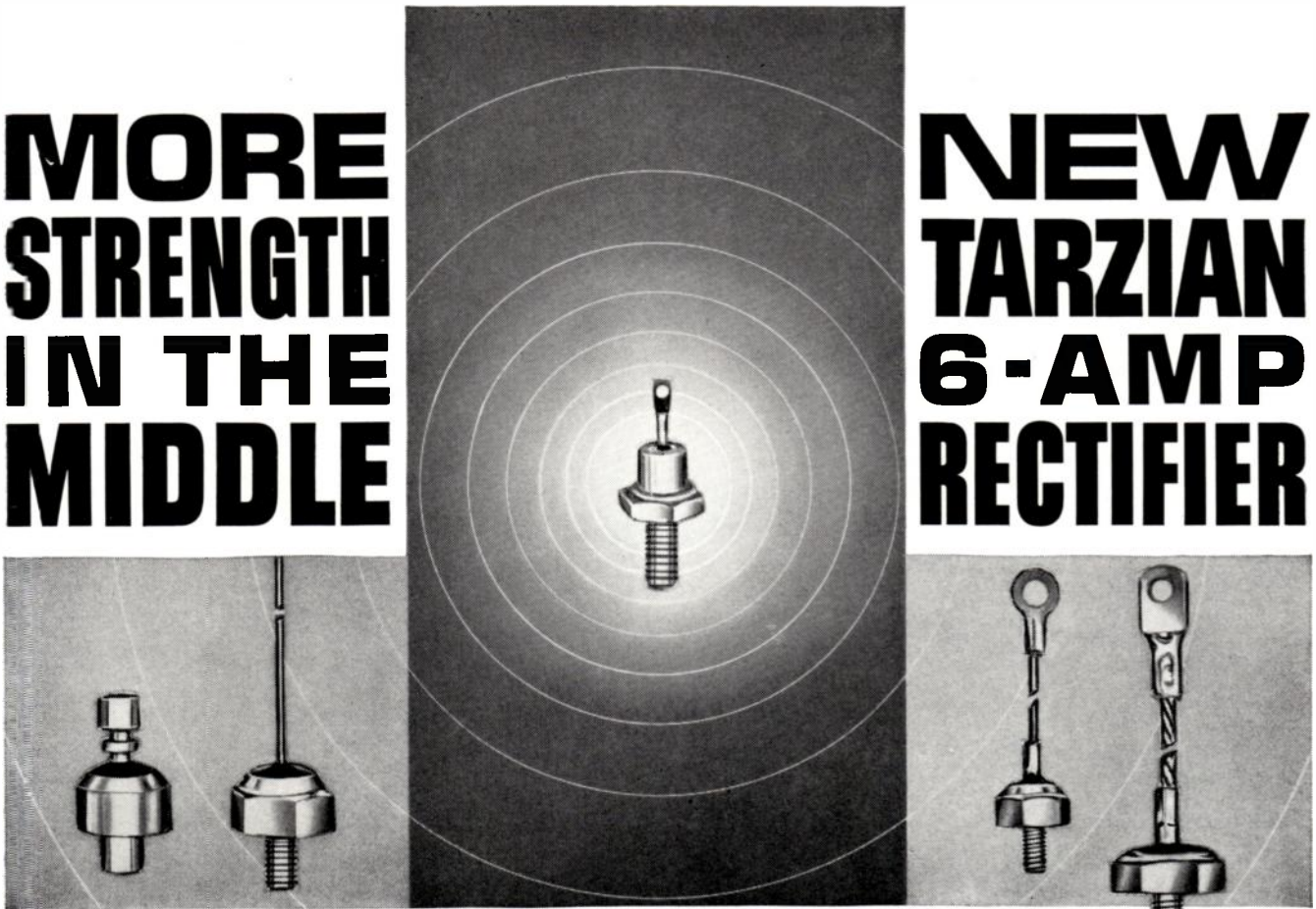
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Professional Group Meetings

(Continued from page 66A)

Philadelphia—October 17

"User Approaches to High Reliability Specifications for Component Parts."
Moderator: W. T. Sumerlin, Philco Corp., Philadelphia.
Panel of Users: M. F. Chamow, GE, King of Prussia, Pa.; A. J. Schwartz, RCA Camden; W. Garner, Westinghouse, Baltimore.
Panel of Manufacturers: L. Jacobson, International Resistance Co, Philadelphia, Pa. D. Van Winkle, Continental Devices Corp., Hawthorne, Calif.; P. Cook, Corning Glass Works, Bradford, Pa.

Electron Devices

Boston—November 9

"State of the Art for Gallium Arsenide Devices," Dr. J. Hilibrand, RCA Semiconductor and Materials Div., Somerville, N. J.

San Francisco—November 8

"Lasers," Dr. Schawlow, Stanford University.

Syracuse—November 21

"Superpower Tubes," Dr. E. D. McArthur, General Electric Co., Schenectady.

Electron Devices
Microwave Theory and Techniques

Boston—November 20

"State of the Art of Solid State Microwave Transmitters," W. Maciag, Sylvania Electric Products Inc., Buffalo, N. Y.

Electronic Computers

Dayton—June 8

"Microelectronic Techniques," C. Carr, General Electric.

Omaha-Lincoln—November 11

Tour: Sage Computer Center, Sioux City, AFB, Iowa, IBM Tour leaders: J. F. Edgerton and K. Love; Military Tour and Discussion Leaders: Maj. R. E. Jennings, Maj. K. L. Sharkey, Capt. E. H. Blum.

Philadelphia—November 14

"Thin Films in Digital Computers," G. Fedde, Remington Rand Univac, Blue Bell, Pa.
Plant Tour.

Pittsburgh—October 25

"Programming the Carnegie Tech Management Game," Dr. K. J. Cohen, T. C. Gilmore and P. R. Winters; Carnegie Tech.

San Francisco—October 24

"Executives 100 Feet Tall," J. P. Eckert, Jr., Sperry Rand Corp.

Syracuse—December 5

"Fluid Amplifiers," B. M. Horton, Diamond Ordnance Fuze Labs., Washington, D. C.

Electronic Computers
Military Electronics

Omaha-Lincoln—November 29

"Critical Path Scheduling (CPM, PERT, PEP, LESS) With Demonstration on IBM 1620 Computer," G. W. Wilson, HD and R, Omaha.

Engineering Mangement

Baltimore—November 16

"The Market for Electronic Stocks," J. McEnroe, Merrill, Lynch, Pierce, Fenner and Smith, Baltimore.

Boston—December 7

"The PERT System and its Applications," J. Pearlman, Control Systems, Cambridge; R. W. Miller, Raytheon, Waltham; and Lt. Col. (USAF) W. W. Diamond, LG Hanscom Field, Bedford.

Dayton—October 11

"Electronics, Man and Space," Maj. Gen. W. A. Davis, Commander ASD.

(Continued on page 70A)

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The new Pin-Trim delivers: (1) more capacitance per cubic centimeter than any other conventional piston trimmer; (2) 75 per cent less weight and 50 per cent less volume than JFD's own miniature trimmers; (3) greater sensitivity; (4) finer adjustment.

If you are looking for maximum compactness between stacked circuit boards, or less stray capacitance in a given area, check the JFD Pin-Trim specifications for your subminiature trimmer applications.

- Overall diameter: 1/8 inch. Overall length above panel: 3/8 inch to 1 inch.
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- 10^4 megohms insulation resistance.
- Q factor of 500 (measured as per JFD #5178).
- 0.5 inch ounce tuning torque.
- Meet or exceed applicable performance requirements of MIL-C-14409A.

For further data, call your local JFD Field office or your JFD franchised Industrial Distributor.

JFD Adjustment Tool No. 5284 (illustrated) available at 85¢.

Model*	Capacitance Range MMF Measured Per JFD #5177 Min.	Capacitance Range MMF Measured Per JFD #5177 Max.	D.C. Working Volts	Dielectric Strength Measured For 5 Seconds at 50% R.H. at Max. Rated Cap.	Insulation Resistance Measured After One Minute at 500V. D.C. and 50% R.H.	Q Factor Measured Per JFD #5178	Unit Weight Grams	Dimen.** Max. ±1/32
PT901	0.5	2.0	500	1000	10^4 Megohms	500	0.62	3/8"
PT902	0.5	3.0	500	1000	10^4 Megohms	500	0.64	1/2"
PT903	0.5	5.0	500	1000	10^4 Megohms	500	0.79	3/4"
PT904	0.5	7.0	500	1000	10^4 Megohms	500	0.94	1"

*These units are also available in the same capacitance values for printed circuit boards in models TT911, TT912, TT913 and TT914.
**Length front of panel.

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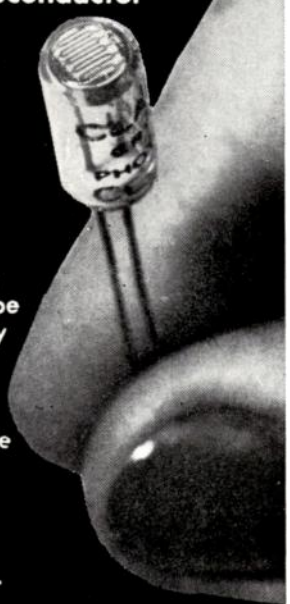
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Professional Group Meetings

(Continued from page 68A)

Dayton—September 6

"Skybolt Guidance Sub-system Management," R. F. Miller, Nortronics, Hawthorne, Calif.

Seattle—September 28

"Theoretical Control Model Pa Program Organization of Massive Eng. Activity" F. Little, Boeing Co., Seattle.

Engineering Writing and Speech

Northern New Jersey—November 16

"Writing from a Novelist's Point of View," Mrs. L. Chinn, Borger, Tex.

Northern New Jersey—October 19

"Improving the Communications Process in our Technical Society," M. Lent, Kearfott Div. of General Precision, Little Falls, N. Y.

San Francisco—November 21

"Book Review of Style Manual for Technical Writers," V. Whipple, Lockheed Missile and Space Co., Sunnyvale, Calif.

Information Theory

Baltimore—December 6

"Communication Systems with Feedback," Dr. L. S. Schwartz, NYU.

Baltimore—November 1

"What Is Information Theory?" Dr. W. C. Gore, Johns Hopkins University.

San Francisco—November 30

"A Generalization of Woodward's Theorem on FM Spectra," Dr. N. M. Blachman, Sylvania Electronic Defense Lab., Mountain View, Calif.

San Francisco—November 9

"Some Recent Results in the Study of Feedback Communication Systems," L. S. Schwartz, New York University.

Instrumentation

Los Angeles—October 17

"Radiological Defense Problems and Protection from Nuclear Fallout," J. W. Davis, State of California Disaster Office, Los Angeles.

Films: "Mission Fallout" and "Bomb Burns and Blast."

Washington D. C.—December 4

"Mohole Measurement Problems," R. Taggart, Robt. Taggart Inc., Falls Church, Va.

Microwave Theory and Techniques

Albuquerque-Los Alamos—November 9

"Some Complex Problems in Microwave Power Measurements," Dr. G. V. Sorger, Weinschel Engineering.

Los Angeles—October 5

"High Power Filter Design," Dr. E. M. T. Jones, Stanford Research Inst., Menlo Park.

Northern New Jersey—October 18

Dr. George C. Southworth, BTL Retired, described the evolution of microwave theory, components and applications. He spoke of his personal work and that of others. Beginning with his own work with water filled pipes, the description progressed, step-by-step, to what we know today as the state of the art.

Orlando—November 29

"Low Noise Amplifications," Dr. G. P. Rodrique, Sperry Microwave Electronics Co., Clearwater, Fla.

Orlando—October 18

"Directive Antennas for Space Application," K. S. Kelleher, Aero Geo Astro Corp.

Washington, D. C.—November 14

"High Power Microwave Switching," G. Klein, Westinghouse, Baltimore, Md.

Panel to stimulate discussion and bring out points of interest: Dr. R. Young, DOFL; Dr. W. Gabriel, AGA; and R. Garver, DOFL.

Washington, D. C.—September 12

"The Future of Microwaves," T. S. Saad, Sage Labs., Inc., Natick, Mass.

3 man panel to stimulate discussion and bring out points of interest: Dr. H. Sommer, W. P. Hooper, Norman T. Fairhurst.

Military Electronics

Dayton—November 1

"History and Development of the Space Detection and Tracking System," Col. W. E. Kelley, Dir. 496L Space Track Program, ESD, AFSC.

Dayton—October 11

"Electronics, Man in Space," Maj. Gen. W. A. Davis, Commander, Aeronautical Systems Division.

Los Angeles—November 29

"Solid State Electronic Systems," Dr. W. Welch, Motorola, Phoenix, Ariz.

San Francisco—October 3

"The Signal Corps Program in Tactical Automatic Data Processing," Maj. Gen. F. F. Uhrane, USA Commanding Gen., Fort Huachuca, Ariz.

(Continued on page 76A)

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The *all mesh* grid and filament structure of STC power triodes for RF heating has given this range of valves an excellent reputation for industrial performance due to the high efficiency and considerable safety factor on grid dissipation. Boilers, water jackets etc. can be supplied, together with terminal connecting rings which have been standardized for size throughout the range.

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Forced air-cooled triode type 3J/187E



All ceramic water-cooled triode type 3RC/223E

Type	Cooling	V_f	μ	gm	V_a dc kV	P_a kW	Output kW
3J/187E	Forced Air	5	12	22	6	3	7
*3JC/223E	Forced Air	5	12	60	6	10	24
*3QC/223E	Water	8	12	60	6	12	24
*3RC/223E	Water—integral jacket						
*3ZC/223E	Vapour						

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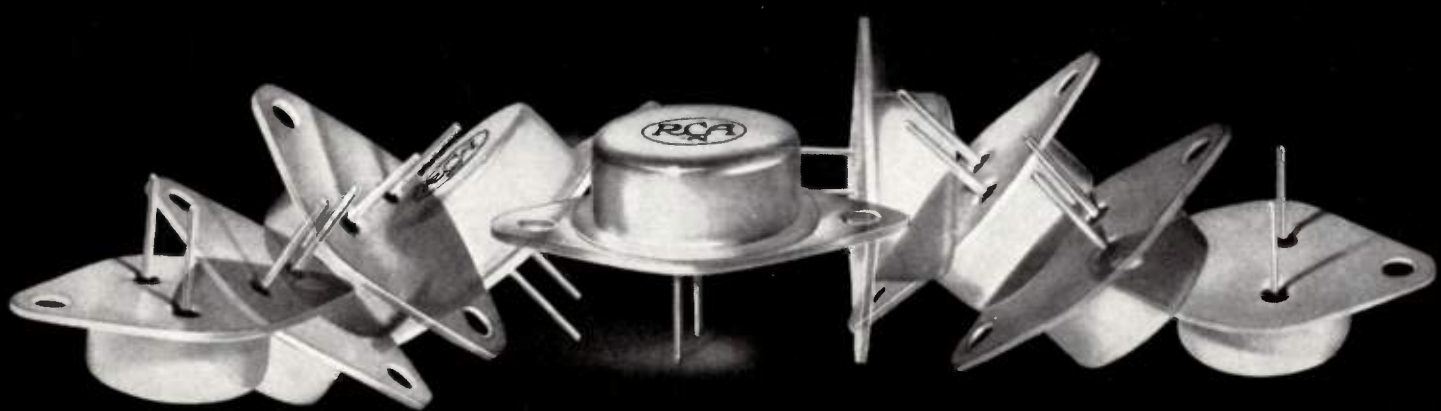
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Base Current	3	3 amp.
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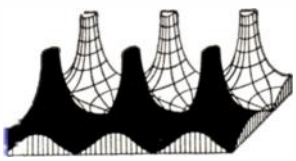
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Proceedings of the IRE

Poles and Zeros



The Two Cultures. We Radio Engineers, along with other engineers and scientists in general are finding ourselves drawn, with increasing frequency, into discussions of the interrelations of science and the humanities.

It is interesting and informative to extract from literature the views of several well-known humanists. Aldous Huxley in his famous "Point Counterpoint,"* has a leading character say " . . . all these famous theories about the cosmos and their practical applications—they've got nothing whatever to do with the only truth that matters. And the non-human truth isn't merely irrelevant; it's dangerous. It distracts people's attention from the important human truth. It makes them falsify their experience in order that lived reality may fit in with abstract theory." It would be interesting to really understand, completely, what the venerable Mr. Huxley means by that. Pulitzer Prize Winner William Saroyan, in his new impressionistic play, "High Time Along the Wabash," has been interpreted to say that the dedicated scientist has to be "hog-tied" to be brought down from his ivory tower into real-world contact with man's every day problems. The British scientist-humanist C. P. Snow, in his "Scientific Progress and the Two Cultures"† makes an urgent plea to all leaders of our society: "Closing the gap between our cultures is a necessity in the most abstract intellectual sense, as well as in the most practical. When those two senses have grown apart, then no society is going to be able to think with wisdom. For the sake of the intellectual life, for the sake of the western society living precariously rich among the poor, or the sake of the poor who needn't be poor if there is intelligence in the world, it is obligatory if there is intelligence in the world, it is obligatory for us and the Americans and the whole West to look at our education through fresh eyes."

In an effort to close the gap, the ECPD (accrediting organization for engineering colleges) sets the minimal humanities content of engineering curricula at 21 semester hours and urges that the schools require at least 24.

In a recent article (Look, Dec. 19, 1961) author Snow writes of the decline of British prestige, "We produced an extraordinary kind of 'gentlemanly' education, supposed to be suitable for proconsuls administering the empire and from this we have never quite recovered."

Dean Gordon Brown of M.I.T. once quipped, "We're making definite progress in humanizing the scientist. Shouldn't we, at the same time, take steps to 'simonize' the humanist?"

A related statement from the editorial page of The New York Times is that quoted in "Poles and Zeros" by past-editor Fink in December 1957, " . . . The scientist, even when doing his best to be simple, speaks another language . . . . Perhaps we had all better learn that language."

In any case I am sure we all agree that we, as engineers, must concern ourselves with value judgements including those of social responsibility. And there is, as Snow points out, a desperate need for intimate understanding and mutual cooperation between "the two cultures." The editor invites your comments, thoughts, and observations on this important problem.

To the 49th and 50th. At the annual meeting, the Board of Directors of the National Electronics Conference voted unanimously to honor the University of Alaska and the University of Hawaii with gifts to their libraries of a set of *Proceedings* of the NEC. Our hats are off to the NEC for this welcoming gesture to our new sister states.

You Can Help! How many people do you know who qualify for IRE membership but just don't happen to be members? You will find it interesting to list your technical acquaintances and check them against the new IRE Directory which hit your desk around December 1st. Help IRE by suggesting IRE membership to those worthy men who do not check out.

Present membership stands at over 96,000—we are virtually certain to pass the 100,000 mark during '62 (What's your guess? Mark it in your appointment book and watch "Poles and Zeros" for announcement). Your help in recruiting new members will be rewarded throughout your career by increases in services to members that are made possible by increasing membership. No other professional society provides so much to its membership for so little.

A membership blank is included in the Directory following page 20.

Another Bonus to Members. IRE celebrates its Golden Anniversary this year. We quote Past President Berkner, "You will be interested to know that the May 1962 issue of PROCEEDINGS OF THE IRE will be a very special anniversary number distributed on publication only to members and subscribers, with no copies sold to anyone, including members, after publication. This issue will be a real 'collector's item,' and should be an inducement for non-members to take immediate action to fill out and submit an application to insure receipt of this limited special edition."

Blast Off! The name has been changed—the former Central Florida Section is now the Canaveral Section. Here's hoping the active membership skyrockets!—T. F. J.

* Harper and Brothers, New York, N. Y.
† Cambridge University Press, New York, N. Y.



T. A. Hunter

Vice President, 1962

Ted A. Hunter (A'29–M'45–F'53) was born in Dike, Iowa, on December 5, 1900. He received the B.S. degree in electrical engineering (1922), the M.S. degree in physics (1923), and the M.S. degree in electrical engineering (1931) from Iowa University, Iowa City. He has taught physics at the University of Pittsburgh, Pa., and at the Rose Polytechnic Institute, Terre Haute, Ind. At present, he is Professor of Research in the Department of Psychology at Iowa University.

His industrial experience includes being Head of the Sound Division of Crosley Radio Corporation, and a transmission line inspector for the Bell Telephone Company. He was also in charge of the oscillator and amateur division of the Collins Radio Company, Cedar Rapids, Iowa, where he did extensive research and development work on linear permeability tuned oscillators. These oscillators were used in the Collins Radio 32V and 75A series of transmitters and receivers, as well as other frequency generating systems developed by the company. At present he is President of his own firm, the Hunter Manufacturing Company in Iowa City. This firm designs and builds laboratory equipment for use in research laboratories in the fields of psychology and medicine.

Mr. Hunter has served on many IRE committees, including Nominations, Policy Development, Sections, Regional, Professional Groups, and Papers Procurement. He has also been a member of the Board of Editors since 1954. Along with many others, he helped to promote the Regional Director system of representation for the IRE, and was the first Regional Director of Region 5. On two other occasions, he held a one-year term of office as an appointed Director to the Board of Directors. Active in initiating the Professional Group publications system, he was Editor of the IRE TRANSACTIONS ON EDUCATION for the first two years. During his term of office as Regional Director, he began the IRE STUDENT QUARTERLY, and has been Editor since its start in 1954. More recently, he was instrumental in the establishment of the position of Student Affairs Secretary of the IRE. In conjunction with the Cedar Rapids Section, he proposed a brochure on electronics as a career, to be made available to pre-college students. This brochure was proved by a sample run and is now being redesigned and reprinted by EIA, soon to be ready for distribution.

Mr. Hunter has been co-organizer of the IRE Cedar Rapids Section, the Cedar Rapids Engineers Club, and the Collins Radio Technical Association. He has served as Chairman and Program Chairman of the Cedar Rapids Section, and as Chairman of the first Cedar Rapids IRE Conference. He is a member of Eta Kappa Nu, Sigma Xi, and Tau Alpha Pi.

Scanning the Issue

Open Letter from the President, III (Berkner, p. 132)—In December the PROCEEDINGS published a letter from President Berkner to IRE Section chairmen, dated October 20, acquainting them with the action of the IRE and AIEE Boards of Directors to consider consolidation of the two societies, and asking them, and through them the membership, for comments which would be helpful to the joint committee which had been established to study the proposal. The procedures adopted by the joint committee for carrying out their study were outlined in a subsequent letter published here last month. In a third and final letter, President Berkner now discusses eight broad questions which were raised by IRE Sections and members in response to his initial letter. His comments provide information that will be of immediate interest and importance to all IRE members.

Some Potentialities of Optical Masers (Oliver, p. 135)—Barely three years have passed since it was pointed out that the principles governing microwave masers could be extended into the optical region to produce coherent light. The great interest which this suggestion aroused in the scientific community was evidenced by the fact that it took only 19 months to make this prediction come true. The announcement in mid-1960 that an optical maser had been successfully operated was a major milestone, signifying that the spectral bonds confining the communications art had been broken, that radio techniques could now be extended into the far broader reaches of the optical portion of the spectrum, that by adding the attributes of coherence to the dimensional advantages of micron waves it would now be possible to produce much higher concentrations of radiant energy in tiny regions of space. The general reader who has not had an opportunity to familiarize himself with the principles and possible applications of optical masers will find this paper to be an excellent introduction to a timely subject. It is interesting to note that a significant step toward increasing the power output of optical masers is reported on page 213 of this same issue.

Moisture Exclusion from Encapsulation of Long-Life Transistors (Cooper, p. 141)—When the transistor first burst upon the scene it was hailed as a revolutionary device largely because of its small size and weight, low power consumption, and long life. The latter attribute arose from the fact that the transistor was the only active component which had no inherent wearout mechanism. In practice, the transistor is not the methuselah it might otherwise be because it must live in an unfavorable environment, one which exposes its surface to a contaminating atmosphere and also subjects it to severe temperature changes. In both cases the most corruptive influence is moisture. In order to reduce this source of trouble to acceptable proportions, extensive efforts have been made to develop satisfactory methods for encapsulating transistors in a dry, moisture-proof surrounding. Today, these methods are more important than ever because of the development of new devices in the areas of microelectronics, integrated circuits, and molecular electronics. The survey of encapsulation techniques presented here will therefore be of timely interest.

Microwave-Carrier Modulation-Demodulation Amplifiers and Logic Circuits (Eckhardt and Sterzer, p. 148)—The central role which data handling systems play in modern electronics has placed a high premium on speed of information flow. Speed in a digital system is dependent on the pulse rate of the components of the system and the amount of time delay these components introduce. The maximum pulse repetition rates reported for various types of digital circuits so far range from 100 to 700 Mc, with corresponding delay times of from 1.5 to 10 nanoseconds. This paper describes solid-state microwave components capable of handling pulse repetition rates of the order of 1000 Mc with time delays of only a fraction of a nanosecond. As a demonstration of how parametric amplifiers can be used to obtain large gain-bandwidth products, this development will be of wide interest to microwave, and circuit analysts as well as to computer engineers.

An Analysis of a Class of Variable-Frequency Sinusoidal Oscillators Using a Single Active Element (Caputo and Highleyman, p. 162)—Oscillators whose frequency can be varied electronically have a wide variety of uses in the present day art. The type of oscillator described in this paper should therefore be of interest to many readers because it has the virtues of being simple (as demonstrated by the practical examples given), of requiring only one rather than two active elements, and of producing a sinusoidal output rather than a nonsinusoidal wave that would require further filtering.

Relativistic Beam-Wave Interactions (Rowe, p. 170)—The never-ending climb toward higher and higher power outputs has resulted in the use of very high voltages in present day RF generators and amplifiers. One interesting consequence of this is that we have reached a point where, in some cases, the voltages are so high that the electron velocity of a stream is a significant fraction of the velocity of light, thus causing the relativistic mass of the electrons to change. What effect would this change have on the operation of a klystron or a traveling-wave tube? The author finds that high electron velocity results in high interaction efficiency, which strongly suggests that such interaction would be very useful in millimeter wave generation and amplification.

Truncation Error of Sampling-Theorem Expansions (Helms and Thomas, p. 179)—This brief paper is concerned with the representation of a continuous band-limited signal by a discrete set of samples and with the error which occurs in practical situations in attempting to reconstruct the original signal from a limited number of such samples. The results presented here will be very useful to information theorists, numerical analysts, communication system designers, and control system specialists.

Transferred Electron Amplifiers and Oscillators (Hilsum, p. 185)—A new, and as yet unobserved, mechanism for producing negative resistance effects in certain semiconductors is proposed, which may make it possible to develop solid-state devices for use at frequencies up to 10^{12} cps. The proposed mechanism calls for transferring electrons between two closely separated regions of the conduction band by means of a high electric field so that they go from a state of higher mobility to one of lower mobility. The conductivity of the material might thereby be decreased as the field is increased, giving rise to a differential negative resistance. The prediction is made that the necessary conditions for observing this phenomenon may be found in gallium antimonide and in semi-insulating gallium arsenide—a prediction that will be of substantial interest to anyone concerned with new device concepts.

Current-Limited Rectifiers (Montgomery, p. 190)—The inherent property of certain rectifiers also to act as current limiters has been enhanced in the several circuits presented in this brief paper. The result is six rectifier circuits that are useful for short-circuit proof, regulated power supplies.

Beta Cutoff Frequencies of Junction Transistors (Lindmayer and Wrigley, p. 194)—This paper presents a simple way to calculate an important transistor parameter by a method which can be applied to all transistors. The topic is one which has not received the attention it deserves, at least in the large-circulation scientific literature, and is one which will be of general interest to all who are concerned with the design and use of transistors at high frequencies.

Scanning the Transactions appears on page 238.

Open Letter from the President—III

The following is the third of a series of monthly letters from President Lloyd V. Berkner concerning the possible consolidation of IRE and AIEE and the progress of studies which began last October.—The Editor

December 28, 1961

Dear Mr. Section Chairman:

This is my last general letter to the Sections concerning IRE affairs, relating principally to progress toward formal consideration of IRE-AIEE consolidation. My term as President will conclude at the year-end, and the affairs of the Institute will then continue in the able hands of Patrick E. Haggerty.

By mid-December, more than 75 of our 112 Sections, and dozens of individual members, had responded to my letter of October 20 on the proposed consolidation. The comments contained in these responses have been invaluable to your Officers and Directors in expressing the attitudes of the membership, and in inviting attention to many aspects of the Institute's affairs that could be improved in the consolidation. Some of these ideas are so useful that they will be considered by the Board for incorporation into our IRE Bylaws during the coming months.

The meeting of the Joint Committee on Consolidation of IRE and AIEE was held in New York on December 13 to study a first draft of a proposed constitution and bylaws for a consolidated society. These discussions were effective and no serious differences of view were encountered. Several points which required further study were defined and referred to appropriate groups or committees for advice. The discussions have proceeded on a note of high professional dedication and in an air of cordiality.

In light of this progress, the four-man executive committee of the Joint Committee (Haggerty, Chase, Pratt, and Teare) was instructed to proceed with proposals for interim measures that would be needed for the transition from the two separate institutes to the consolidated society. They will meet in New York on January 7 to prepare these proposals in light of the reports of the several two-man committees named in my last letter. At this time I would estimate that the time schedule specified in the resolution of our Board could be met.

It seems appropriate that I comment now on several general points raised in some of your letters in light of the discussions so far. I might add that these comments have been reviewed by President Chase of AIEE.

1) *Procedure.* Some members have felt that they could not comment on the proposal until the final constitution and bylaws were available for study. Certainly final commitment by each member will not be asked until specific constitution and bylaws of a consolidated society are placed before the memberships by the two Boards for final vote. In the meantime, however, your Officers and Directors felt that their intent to enter into serious negotiations, and the frame of reference within which consolidation might be considered during such negotiations, should be made known to the membership without delay so that your attitudes might be taken into account. The general frame of reference for negotiations was spelled out in my letter of October 20, and the procedures in my letter of November 28, 1961.

2) *Publications.* Two sharp points of view have been expressed.

a) Nothing should be done to dilute the high technical effectiveness of the PROCEEDINGS OF THE IRE in any way since it is the foundation of our professional proficiency.

b) A tutorial journal of broad technical coverage with nonspecialized mathematical treatment is needed to keep members in touch with the full range of electrical and electronic developments.

Both points of view are being considered. We hope that one of the great benefits of consolidation will be to make both types of publication available to all members without increase of dues.

3) *Professional Groups.* Several questions have been

raised concerning the future of the Professional Groups. Both institutes recognize the professional technical group as a major development in professional engineering organization. The professional technical group represents a quasi-independent society of technical specialists organized under its own administrative committee with its own publication (the quarterly TRANSACTIONS produced with the aid of the parent society), its own membership (as a segment of the parent society), its own fees (with added subsidy of the parent society), its own national meetings, and its own local professional technical chapters (organized under the umbrella of the Local Section of the parent society). The professional technical group represents the opportunity for specialists to develop their subject under the most favorable conditions and atmosphere, while at the same time maintaining broad contact with the full content of the profession. Thus the parent society offers opportunity for both breadth and depth to the profession. In consolidation, the total number of specialties would be increased, enlarging the number of professional groups, with each group organized around a center of interest and enthusiasm. With enlarged membership of groups, made possible by reduction of duplication between societies, funds available to each group should be substantially increased. Under consolidation, the enlarged Headquarters staff of the consolidated society could render even greater executive-secretarial services to each group. In recognition of the importance of the technical work of the professional technical groups, your Board will consider at the January meeting the appointment of the Chairman of the Professional Groups Committee as a Vice President of the Institute, to serve on our Executive Committee. Bylaws to effectuate this proposal have already been circulated.

4) *Non-national character.* Questions as to the international character of the new society have been raised. The new society, like the IRE, would be non-national in character (not national or polynational). I remind you that AIEE influence has spread substantially in sponsorship of professional activities in Mexico and Canada, and it has substantial membership in Europe. The word "American" will not be used in the title or constitution to define the activities of the consolidated society. Careful attention is being given to sectional and regional (or district) boundaries which in some cases may cross international boundaries to fuse natural geographic professional interests. At least two geographic regions, principally involving areas not in the U. S., are being considered on the basis of strong existing IRE and AIEE membership and professional activity.

5) *Conventions.* Concern has been expressed about the future of the IRE International Convention and Radio Engineering Show. A special study of convention activity is being undertaken so that the needs of the consolidated society could be met, at the same time avoiding the proliferation that duplication now encourages. Certainly the International Convention and Radio Engineering Show would be retained.

6) *Vitality.* There has been some fear expressed that the vitality of the IRE might be lost through lack of competition between the two institutes, or through consolidation of IRE with the older society.

Careful attention to this problem shows:

a) The AIEE is a vigorous society, of ever-broadening interests, which has continued to attract young members, and is steadily evolving toward greater strength.

b) The vigor of IRE arises out of its subject matter and organization, and not out of competition. "IRE was at the right place at the right time."

c) The subject materials of IRE and AIEE are migrating onto the same ground. Thus power, switching, transmission, wire communications are employing, to an ever increasing extent, the techniques familiar in IRE, and introducing new techniques useful to the field of IRE.

d) Competition between the societies dissipates our professional resources which could better be used for subject penetration in depth, wastes time in duplicate meetings, confuses efforts toward standardization, divides the same subject between two independent groups and publications, compounds student uncertainty on graduation as to which group to join, and is creating an ever more unsolvable administrative problem for both societies.

Therefore your Board believes vitality can be gained, not lost, through a suitable consolidation of the two institutes.

7) *Size.* A few have expressed concern that the very size of the consolidated society might react to submerge the interests of the individual member. Without proper measures in organization, this danger certainly exists. At the outset, one must accept the fact that our profession is growing rapidly. Our IRE membership exceeded 96,000 at year-end, and will quite certainly exceed 100,000 by the 50th Anniversary in May. Thus, size is not a matter of choice since it is a function of the power of our profession in today's society. If IRE is to maintain its leadership, it must respond to the needs of the whole profession. Consequently, it must be organized to serve the individual member in spite of size. Consolidation simply advances the date of reaching a given size for the resultant society, a date that would soon be reached by IRE in any event. IRE has already demonstrated that the problem of size can be met by simple, decentralized organization, through the professional technical groups, and through well planned sectional activity. The rapid development of the local professional group chapters under the auspices of the Sections represents an ever-enlarged opportunity for the individual member in IRE affairs. We would anticipate that in the consolidated society the number of Sections would increase by 30 to 40, and the professional group chapters by perhaps as much as 1000.

8) *Organization.* In response to questions on organiza-

tion, I believe there is full agreement on the basic organizational principles:

a) A small Board, no more than twenty-five or thirty, with both regional and broad technical interests represented.

b) Clear separation of policy and administration, with policy residing in the Board and full administrative responsibility residing in a strong Headquarters.

c) A small Executive Committee of the Board that meets very frequently, with each member an expert on, and in close touch with, some aspect of the administrative work of the Headquarters, representing a continual performance audit of the society. We might recall that individual members of IRE's Executive Committee now audit: Membership (Secretary); Finance (Treasurer); Professional Groups (Vice President suggested); World Affairs; Sectional Affairs; Student Affairs; Publications (Editor); Standards; etc. I have no doubt that a consolidated society would follow similar procedures.

In developing the subject of consolidation over the past year, my own attitudes have evolved considerably. As one delves into the intricacies of the problems involved, what seemed earlier merely desirable now seems to me essential. The mechanical problems that earlier seemed insuperable now seem relatively easier to solve.

In your individual responses, you have helped the Officers and Directors mightily in their perspective and evaluation of the problems of consolidation. Every suggestion has been tabulated, evaluated and discussed in the development of our thinking.

As nearly as I can judge, almost the whole membership represented by the responses (certainly more than 85 per cent) would favor consolidation in a framework represented in my earlier letters. Therefore, I am optimistic that this may be brought about.

May I express to you personally my thanks for your vigorous activity in bringing qualified members into IRE. The rapid rise in membership (1000 a month) is a measure of your response. It is a recognition that to retain leadership, IRE must represent the whole profession. From the several sectional plans that have been outlined to me, I know that your continuing efforts will be most effective. Equal attention to raising membership grades of properly qualified members, as described in my letter of November 20, will similarly strengthen our professional organization.

I know I speak the feeling of the membership when I express our gratitude to the administrative and editorial staff of our Headquarters under the leadership of George Bailey. To them, IRE is a dedication to an ideal from which incalculable strength derives for our profession. It is a pleasure to leave my chair with the warm friendships among the staff and the membership that my years with IRE have rewarded me.

In turning the reins of the Institute over to Pat Haggerty, I do so with confidence in his deep appreciation of IRE's ideals and objectives, and his intimate knowledge of its mechanisms. May I express to you, and to your officers and members, my own thanks for the high honor and the many courtesies you have rendered to Mrs. Berkner and me, for which we could never give adequate return.

Sincerely yours,



L. V. BERKNER
President

Some Potentialities of Optical Masers*

B. M. OLIVER†, FELLOW, IRE

Summary—**This paper, originally presented at the 1961 WESCON Convention, is intended as an introduction to the principles and possible applications of the optical maser. Very little prior knowledge on the part of the reader is assumed. The intent is to acquaint him with these exciting new devices with the hope that he, in turn, will discover applications not foreseen by the author. The methods of generating coherent radiation, of focussing it, and of collimating it into light beams are described. The use of lasers for communication is explored, and certain medical and other applications are suggested.**

Introduction

NOT LONG AFTER the development of the microwave maser by Townes, Bloembergen and others, it became apparent that the same principles could be applied to the generation of radiation of far higher frequency. In a historic paper in the *Physical Review*[1] Schawlow and Townes predicted the possibility of extending maser principles into the optical region, described general types of structures which would be required, and speculated upon the performance possible. Very few developments in recent years have excited the imaginations of so many scientists and engineers as has the optical maser, or "laser." The reason, of course, is that this device offers for the first time a means for producing and amplifying coherent light and therefore opens up the optical spectrum for exploitation by all of the techniques currently used in the radio spectrum.

In July, 1960, Maiman [1] of Hughes Laboratories announced the first successful production of coherent light on a pulse basis using optically pumped ruby. Early in 1961 Javan, Bennett and Herriott of Bell Laboratories announced the successful CW operation of a gaseous optical maser [2]. In addition, there has been a great deal of activity in this field at many other laboratories, and the effort is increasing.

The basic principles of the optical maser are the same as those of the microwave maser. Fig. 1 illustrates in schematic form the essential ingredients. First of all, there must be a resonant cavity. In an optical maser this is formed by two precisely oriented mirrors, one of which is slightly transparent. It can be shown that resonant modes will exist between these mirrors at frequencies for which the spacing is an integral number of half wavelengths. In this space between the mirrors is placed an active medium which may be either a gas or a crystal doped by certain atoms (such as chromium in the case of ruby). The medium must possess two atomic states separated in energy by an amount corresponding to the frequency desired, and it must be possible to over-

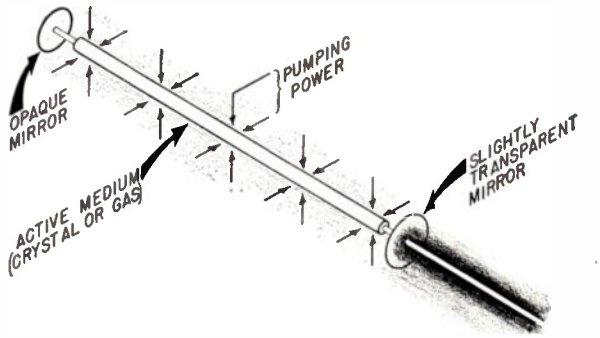


Fig. 1—Elements of an optical maser.

populate the upper of these states with respect to the lower. This is done by "pumping" the atoms from a ground state to a higher energy state either electrically or optically. From this higher energy state the atoms usually decay nonradiatively to the upper of the two energy states involved in the desired transition. From this upper state some atoms will decay spontaneously to the lower state and emit light just as occurs in any neon sign. The light caused by such spontaneous emission is incoherent and is radiated in all directions at random. However, in the presence of the resonant cavity some of this spontaneous emission will excite one of the resonant modes of the cavity, and the field associated with the resonance will induce emission in the medium. This induced emission is phase coherent with the field which induces it and as a result, if the interaction is strong enough, a coherent electromagnetic wave will build up corresponding to one of the modes of the resonant cavity. Some of this energy will leak through the partially transparent mirror forming one end of the cavity and emerge as a sharply defined beam of coherent light. The significant thing is that this beam of light is a plane coherent electromagnetic wave just as would be produced by a radio transmitter, but of vastly higher frequency.

There are two aspects to wave coherence: spatial and temporal. A wave is *spatially* coherent if there exist surfaces over which the wave amplitude as a function of time is highly correlated. If the coherence is complete, the correlation will be unity, and the voltage at the one point will be proportional to the voltage at other points on the surface. As an example of spatial coherence, consider the amplitude at any two points on an equiphase front in the light from a distant star. The amplitude at two points is the same function of time, and similarly at a given time, the amplitude along different rays is the same function of distance.

A wave exhibits *time* coherence to the degree that there is correlation between the amplitude of the wave

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[1] A. L. Schawlow and C. H. Townes, "Infrared and optical masers," *Phys. Rev.*, vol. 112, pp. 1940–1949; December 15, 1958.

at a given point at one time and at some later time. A single frequency represents the extreme of time coherence; so does any combination of single frequencies, harmonically related. If the line components of the spectrum are broadened due to random modulation, the time coherence is lessened and in the extreme, when the spectrum of the wave consists of a smooth distribution of frequencies, as is true of black body radiation, for example, time coherence virtually disappears. Spectral purity (*i.e.*, the degree to which the spectrum approaches a line spectrum) is thus a measure of time coherence.

The following analogy may be helpful in visualizing the difference between coherent and incoherent radiation. Incoherent radiation may be thought of as the three dimensional analog of the pattern of waves on the surface of a swimming pool in its usual state just after the swimmers have left. Waves of all different wavelengths are racing every which way at random, and there is little correlation between the time functions representing the amplitudes at two widely separated points. If the pool were surrounded by quiet water at the same level and the walls suddenly removed, the agitation would spread out in all directions in a way which resembles the radiation of incoherent light. By contrast, if the surface of an otherwise quiet swimming pool were set in motion by the up and down oscillation of a float extending clear across one end, a series of plane waves would be produced. These would exhibit high spatial coherence, because the wave amplitudes as a function of time at different points would be highly correlated. And if the motion of the float were periodic, the waves would exhibit time coherence as well. If the walls were again removed the wave pattern would propagate out in a beam normal to the float, and if the wavelength were short compared with the length of the float, the beam would exhibit very little spread with distance.

Ruby lasers operate on a pulse basis and give quite high peak power output. Characteristically, pulses on the order of 10 kw output power and pulse durations of the order of 1 msec are obtained. The total energy per pulse is therefore on the order of 10 joules. Although the spectral line is clearly narrowed by the maser action, the time coherence is still relatively poor in the ruby laser. The ruby rods tend to oscillate in several modes simultaneously and to produce a series of short spikes instead of a single pulse. The line width of each mode is on the order of 10 Mc as a result of the spike amplitude modulation. The spatial coherence also is far from perfect and is thought to be limited by optical imperfections in the ruby crystals themselves. Hopefully, these are defects which will be eliminated with further research, and in our discussion to follow we will assume ideal coherent operation to be possible. The CW gaseous masers operate at relatively low power levels—on the order of 20 milliwatts—and exhibit excellent coherence. Line widths of approximately 1 cps have been reported, and the spatial coherence seems to agree with what would be expected theoretically. There is every reason to believe that the power output of the gaseous laser can be increased, and these devices look very promising for communication applications.

Coherent Optics

The availability of coherent light greatly increases the scope of things that can be done with optical systems. In particular, the spreading of beams of light becomes limited only by diffraction. In an ordinary searchlight the beam spread is principally due to the finite size of the source of light. A point of the source lying on the optical axis produces a beam parallel to the optical axis, while various points of the source lying off the optical axis produce beams at various angles with respect to the axis. The totality of all these beams thus spreads at a rate determined by the greatest extension of the source and by the focal length of the objective. By contrast, plane waves radiated by an optical maser spread in the same fashion as would the beam from an antenna having the same size measured in wavelengths. Having coherent light is tantamount to having a point source.

When the light from a star is imaged by a small telescope objective under ideal seeing conditions, a diffraction pattern called an Airy disk is formed. It consists of a central patch of light surrounded by a series of rings, the intensity as a function of radius, r, being given by

$$I(r) = I_0\left[\frac{2J_1\left(\frac{\pi d}{\lambda f}r\right)}{\left(\frac{\pi d}{\lambda f}r\right)}\right]^2, \tag{1}$$

where

d = diameter of the objective,
f = focal length of the objective,
λ = wavelength,

and J_1 is the first order Bessel function. This same diffraction pattern is formed when the light from an optical maser is brought to focus with an ideal lens, assuming the beam illuminates the lens uniformly as shown in Fig. 2. In general, the diffraction pattern is the (two-dimensional) Fourier transform of the aperture illumination. In this case the intensity at the center of the spot is

$$I_0 = \frac{A}{\lambda^2 f^2}P \tag{2}$$

where A = area of objective, and P = power in light beam.[2] Taking a value of 10 kw for the power output of the ruby laser at a wavelength of 0.7 μ, we find the power density at the center of the image to be about 10^{16} watts per square meter. This is a power density far in excess of anything normally obtained in the laboratory. As a comparison, the power density at the surface of the sun is less than 10^8 watts per square meter. Thus, the ruby laser is theoretically capable of producing a power

[2] This relation can be derived directly from the radio transmission expression, (5), by setting $D=f$, $A_T=A$, $P_T=P$ and $I_0=P_R/A_R$. Again it assumes a uniformly illuminated aperture.

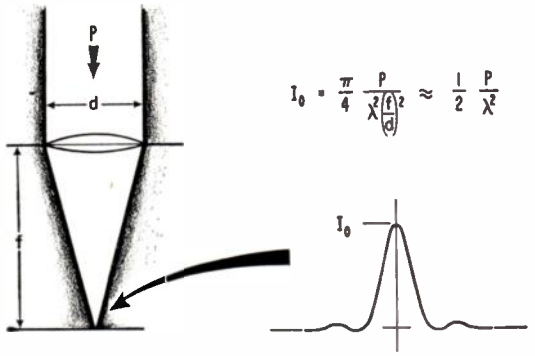


Fig. 2—Focusing of coherent light.

density one hundred million times that of the surface of the sun! This high-power density is accompanied by a correspondingly high-electrical field strength given by

$$E = \sqrt{\eta I_0} \tag{3}$$

where

$\eta = 120\pi$ = impedance of free space.

For the above case we find

$E = \sqrt{120\pi \times 10^{16}}$
$= 2 \times 10^9$ volts per meter
$= 2$ million volts per millimeter.

At such fields it should be possible to produce many effects heretofore unobservable. Such possibilities as the alteration of the construction of molecules, the disruption of chemical bonds in small regions inside homogeneous substances, etc., suggest themselves.

If, after having been brought to a focus by an initial lens, the light from the diffraction pattern is allowed to propagate further, it can illuminate a much larger lens which can in turn recollimate the light as shown in Fig. 3. Such an arrangement will be recognized as the simple astronomical telescope used in reverse. Just as the resolving power of a telescope is increased in proportion to the diameter of its objective, so the beam spread of the emerging beam from this "optical antenna" is inversely proportional to the diameter of its objective. In fact, the width of the major lobe from the peak to the first null is found from (1) by setting $J_1 = 0$, and is the usual formula for the resolving power of a telescope

$$\frac{\pi d}{\lambda}\frac{r}{f} = \frac{\pi d}{\lambda}\theta = 3.8317 \cdots$$

$$\theta = 1.22\frac{\lambda}{d}. \tag{4}$$

As a result of the decreased beam spread, distant points will be illuminated more intensely, and the antenna will exhibit a power gain equal to the square of the normal magnification of the device as a telescope. Used as a searchlight, a laser followed by a telescope can produce a remarkably small spot of light at great distances. For example, a 12-inch diameter telescope on the earth

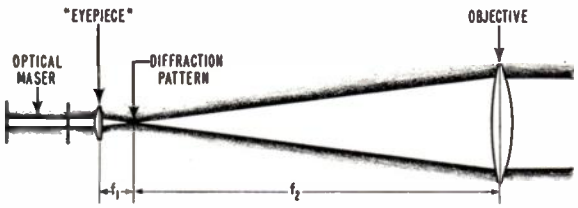


Fig. 3—Optical "antenna" with power gain of $(f_2/f_1)^2$.

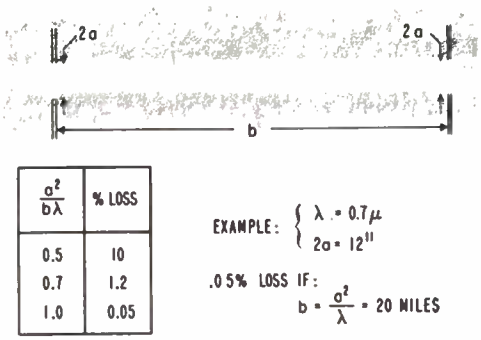


Fig. 4—Transmission between apertures.

would produce a central spot of light only 8800 feet in diameter on the moon. The illuminated patch, of course, corresponds to the figure of confusion of the same objective when used as a telescope.

The extremely small beam spread possible at optical frequencies with coherent light suggests that it should be possible to transmit power over considerable distances with relatively little loss. This turns out to be the case. For a given maximum size of antenna (considering both the receiving and transmitting antenna) the least power loss will occur if the beam intensity as a function of radius off axis is properly shaped. This shape is approximately Gaussian and is, in fact, the distribution which arises naturally when the resonator of the laser consists of two confocal concave mirrors as described by Fox and Li [3]. The beam from a laser employing such a resonator one meter in length is only about a millimeter in diameter: a fine thread of light. However, its radial intensity distribution will be preserved after passing through an antenna of the type shown in Fig. 3, and so it is simple to create a beam of large cross section having this Gaussian distribution. If the diameter of the transmitting aperture and receiving aperture are both $2a$ and if the distance between them is b, as shown in Fig. 4, then, with such a beam, the power loss will be as given by the table and example in that figure. A loss of one twentieth of one per cent of the power in a twenty mile hop is less than occurs with the average transmission line. However, this low loss can only be achieved if scattering and refraction in the medium are absent, and hence can be realized only in free space or if a controlled atmosphere is provided in a pipe connecting the two apertures. The possibility of transmitting power by optical means from earth to a satellite or from one space vehicle to another is a very real one.

Communication by Coherent Light

The use of coherent light for communication purposes is an obvious application, and in this section we will consider the potentialities of lasers in this service. The factors which must be considered are the bandwidths afforded by the new type of channel, the transmission loss, the inherent noise, and the cost. Let us look at these in turn.

In the red end of the spectrum the frequency of light is approximately 4×10^{14} cps. A 1 per cent band in this portion of the spectrum has a width of 4 million Mc, enough for a billion simultaneous telephone conversations. Thus, it should be possible in theory to transmit all of the conversations going on anywhere in the world simultaneously over a single thread of light one millimeter in diameter. Apparently we have bandwidth to burn. This is especially true when one considers the fact that because of the extremely directional beams that can be produced, many simultaneous channels can exist in the same frequency band without mutual interference. The problem is not one of lack of spectrum space, but of how to make use of it: of how to modulate the optical maser or its output so as to fill up the spectrum. Work in this direction is going on at the present time, and beams from lasers have been modulated at frequencies up to X band, so that channels 10,000 Mc wide have been achieved but at the cost of large modulating powers. There seems little doubt that ways will be found to produce extremely wide-band modulation and thus utilize the bandwidth capabilities of optical channels, but even on a less ambitious bandwidth basis; these channels are still attractive, as we shall see.

The transmission loss in an optical channel is small because of the very directional beams easily achieved, as illustrated by the example of power transmission. The basic formula of transmission loss in the optical region is the same as in the radio region, namely,[3]

$$\frac{P_R}{P_T} = \frac{A_T A_R}{\lambda^2 D^2} \tag{5}$$

where

P_T = transmitted power,
P_R = received power,
A_T = area of the transmitting antenna,
A_R = area of the receiving antenna,
λ = the wavelength,
D = the distance between the antennas.

This formula assumes uniform illumination of the transmitter aperture, which indeed gives the least power loss on axis. The λ^2 in the denominator shows that the power received is proportional to the square of frequency and this reflects the increasing concentration of energy by the transmitting antenna as frequency is increased. It is important to remember that with fixed antenna sizes any improvement in transmission with increased frequency comes about from this cause alone.

There are practical limits to the directivity of beams which can be achieved even in the optical region. In order for the theoretical beam spread to be attained, an objective mirror must be well within a quarter wavelength of the true figure over its entire surface. While the surfaces of a lens are less critical, there are more of them. As the size is increased this accuracy requirement becomes increasingly difficult to meet. The 200-inch telescope, for example, is far from accurate enough to realize its full resolving power.

Even if the optical surfaces were perfect, the turbulence of the atmosphere would limit the usable beam sharpness. It is very seldom that seeing conditions are good enough to permit the resolution of points separated by less than one half second of arc, so that beams sharper than this would not be suitable for communicating from the earth's surface to space. Along the earth's surface the situation is even worse because of the greater length of high density airpath, so that unless a controlled atmosphere or vacuum is provided, beams of angular spread of at least a few seconds must be used.

If we now further assume that atmopsheric refraction effects are absent, as in an evacuated pipe or free space, there still remains the problem of pointing an exceedingly sharp beam in the right direction. Vibration of mounting structures, bending of supports by thermal expansion and, in the case of free bodies, changes in the moments of inertia due to motions of the parts will all introduce pointing errors difficult to reduce below one second of arc.

In addition to turbulence troubles in the medium, a great deal of power can be lost over a long path through molecular scattering and scattering due to suspended particles. Again the use of vacuum or perhaps a filtered helium atmosphere would reduce this loss. Further, any terrestrial path must be bent to conform to accessible routes and to the earth's surface. This can be done through mirrors or prisms or lenses disposed along the path at frequent intervals. It is difficult to make such a beam bender with less than 1 per cent loss, so that an attenuation on the order of one or more nepers per 100 bends is to be expected.

Quantum mechanical analyses [5] have shown that even an ideal amplifier has a noise power spectral density referred to the input which is given by

$$\psi(\nu) = \frac{h\nu}{\exp\left(\frac{h\nu}{kT}\right) - 1} + h\nu, \tag{6}$$

where T is the absolute temperature of the source (resistance for circuits, weighted average of the field of view for antennas). The first term in this equation represents thermal noise and at low frequencies (for which $h\nu \ll kT$) this term approaches the familiar kT watts per

[3] Eqs. (1), (2) and (5) may all be derived from Huyghens' principle. Eq. (5) also follows from the fact that the aperture of an isotropic antenna is $\lambda^2/4\pi$, as may be proved thermodynamically.

cps while the second term is negligible by comparison. At high frequencies (such that $h\nu \gg kT$) thermal noise disappears, but by then the second term alone produces more noise than the total at low frequencies. Thus, at high frequencies more signal power must be received to achieve a given SNR. This is true regardless of whether coherent or incoherent detection is used, and basically stems from the fact that the energy per quantum increases with increase in frequency. At optical frequencies the SNR is given by

$$\frac{S}{N} = \frac{P_R}{h\nu B} = \frac{W_R}{\frac{h\nu}{2}} = 2\bar{n}, \tag{7}$$

where

W_R = average energy received per pulse or per Nyquist interval, $1/2B$.

$\bar{n}$ = average number of quanta per pulse or per Nyquist interval.

Thus, the number of quanta received per pulse (or per Nyquist interval) is directly related to the SNR.

The linear increase of noise power density with frequency detracts from the square law increase of received signal power with frequency (with fixed antenna size) so that only a first power improvement in SNR with frequency is obtained.

It appears that optical masers and associated equipment will not be significantly more expensive than other terminal equipment for present-day communication channels. The big question mark in the application of optical masers to telephone communication would seem to be in the expense of the construction and maintenance of the light pipes required. Not enough is known about this at the present time to make even a rough guess as to the cost per mile. Suitable light pipes will probably not be cheap compared with coaxial cable for example, but if very wide-band modulation means can be devised, their cost can probably be cheap per channel of communication.

A Terrestrial Example

Using the equations presented in the last section, it is possible to compute the performance of any given channel aside from the unknown of losses in the medium. As a particular example, let us assume that we have a gaseous CW maser with a power output of 80 mw at one micron wavelength. We wish to use transmitting and receiving apertures no bigger than two inches in diameter; we want a 4-Mc channel and a SNR of 40 db at the receiver. It then turns out that the maximum distance over which this can be done is 4000 miles. This figure does not at first seem very surprising since coaxial circuits exist over this distance today, but what is impressive is that this distance is achieved without repeaters. Losses in the medium and in beam benders would reduce this distance perhaps by a factor of 1/10.

Space Communication

It is interesting to compute the communication capabilities of optical masers over long paths such as are involved in space communication. This subject has been investigated by Schwarz and Townes [6] using somewhat different assumptions from those which follow. Here let us assume a pulse code channel so that we can use devices such as the ruby maser with its higher power output. Using the preceding transmission formulas the curves of Fig. 5 were computed. The ordinate in this chart is the geometric mean of the diameter of the objective at the transmitter and receiver, *i.e.*, $\sqrt{d_T d_R}$. The curves show the value of this mean objective diameter required as a function of distance between transmitter and receiver. The parameter n_1 attached to each curve is the expected number of photons received per joule of radiated energy. Thus if a pulse of energy W joules is radiated, a number of photons $\bar{n} = n_1 W$ will be received on the average. With an ideal receiver (7) shows the SNR in db would be $3 + 10 \log \bar{n}$. Allowing for less than ideal performance in the receiver we may take $(S/N) \approx 10 \log \bar{n}$.

As a first example, let us assume a ruby maser with 10 joules output per pulse. We see that the mean objective diameter required for a 60-db circuit between earth and moon is only 0.3 inch. In other words, the ruby rods themselves without any associated optical system would be sufficient for communication. As a second example, we see that a 3-inch objective is sufficient to provide a 40-db SNR channel between earth and Mars, even when the latter is at superior conjunction (*i.e.*, on the opposite side of the sun from the earth). Finally, we note that 10-inch telescopes would suffice to provide a 30-db SNR clear across the solar system. From these figures it would appear that optical masers are very attractive for interplanetary communication.

What happens if we attempt to use optical masers for interstellar communication? The situation we find here is depicted in Fig. 6. We note that n_1 now ranges from unity to 10^{-6}, so that for points on the lowest curve a million joule pulse is required to receive one photon on the average. Above a 10-inch antenna diameter the curves are shown dotted because it is very doubtful that the sharpness of the beams involved for larger diameters is practical for the reasons stated earlier. It appeared quite practical with lasers of present performance to communicate across the solar system. However, we see from Fig. 6 that the solar system is only a millilight year in diameter. By contrast, the nearest star, Alpha Centauri, is four light years distant. This star is a binary and so probably has no planetary system. In fact, we have to go out to ten light years before we find solar type stars such as Epsilon Eridani, and so it is not until this distance is reached that there is more than an infinitesimal likelihood of encountering other intelligent life. It is obvious from the curves that a tremendous increase in optical maser power output must occur

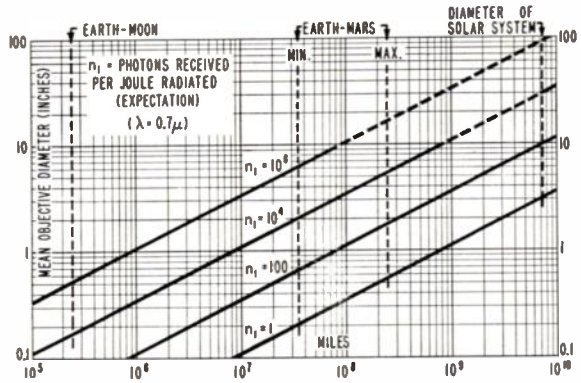


Fig. 5—Coherent optical transmission over interplanetary distance.

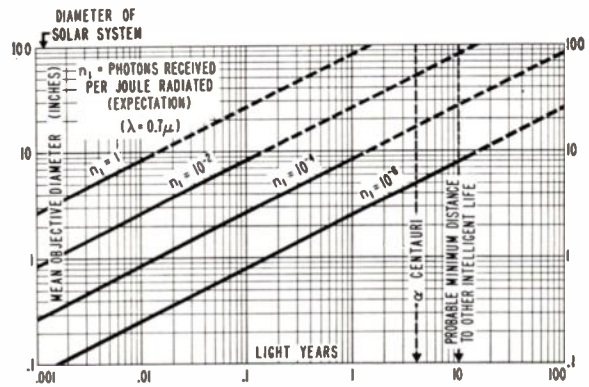


Fig. 6—Coherent optical transmission over interstellar distances.

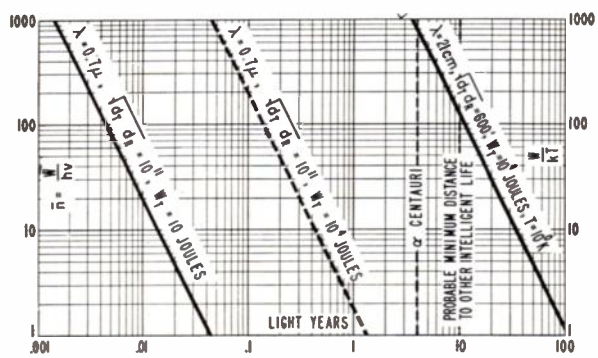


Fig. 7—DX performance—optical maser vs microwave system.

before interstellar communication by this means is feasible. Fig. 7 shows a comparison between the performance of optical masers and a practical, high-power 1420-Mc system at interstellar distances. The ordinate here is the expected number of photons per pulse in the case of the optical system and the quantity W_R/kT which has the same significance for the radio system. The curve at the far left is for the present ruby maser and 10-inch antennas, the middle curve shows the effect of increasing the power per pulse to 10^4 joules, and we see that even here the expectation is far less than one photon per pulse at 10 light years. By contrast, a radio system near the hydrogen line frequency ($\lambda = 21$ cm) using 600-foot antennas such as the one currently under construction at Sugar Grove, a transmitter power of 10^4 joules per pulse (for example, 10 Mw for 1 msec) and a receiver noise temperature of 10°K would have the performance shown by the curve at the far right and would be able to achieve a SNR of over 20 db at 10 light years. The radio system has about 417 times the beam spread of the optical system in Fig. 7. The reason for its superior performance in spite of this poorer directivity is two-fold. First, the assumed receiving antenna has 720 times the diameter and therefore picks up $(720/417)^2 = 3$ times the power. Second, the noise power per cycle is some 2000 times less at the radio frequency and 10°K than at the optical frequency. Thus at a given range the SNR for the radio system is 6000 times that of the optical system. For a given SNR the range is therefore $\sqrt{6000} = 77.5$ times as great.

Conclusions

Optical masers appear to offer exciting possibilities in several areas. First, they make it possible to achieve tremendous power densities in tiny regions of space. They are thus a research tool for investigating the effects of such densities on physical and chemical processes. There are also undoubtedly hundreds of industrial processes where extremely high energy densities could be used. Examples suggest themselves such as microengraving to produce complex semiconductor devices, or ultraminiature circuit components of various sorts. For example, grids for klystron tubes might be formed by evaporating a regular pattern of holes in a thin metal film.

Pulses of parallel light from optical masers may be imaged by the lens of the eye onto the retina where they will produce scar tissue as a result of steam formation and consequent tissue damage. This is attractive as a nonsurgical means of preventing retinal detachment. It also presents the less attractive possibility of blinding enemy troops. There has been much speculation that the development of the optical maser might make possible the proverbial death ray. It is clear that enormous increases in the amount of radiated power would be required before such a ray would be effective, except possibly for retinal damage and incendiary effects produced by focussing the light.

Optical masers can produce far tighter beams of radiation than have heretofore been possible. Such beams will allow efficient communication over great distances. Small antennas suffice even for interplanetary distances, but for interstellar distances the larger antennas required are impractical and optical systems are not competitive with decimeter and microwave radio systems. Highly directive beams make possible optical radar and detection schemes of heretofore unachievable resolving power. For certain space applications where the receiver may be in an inaccessible location such as a satellite, coherent optical beams may be economical for the transmission of power.

Optical masers make available tremendous communication bandwidths both because of the enormous number of cycles in a small percentage of the visible spectrum, and because of the space diversity resulting from the tight beams.

Finally, it should be added that optical masers are so new and so revolutionary that other uses will undoubtedly develop which cannot be foreseen at the present time. Some of these uses may far exceed in importance anything which has been discussed.

BIBLIOGRAPHY

[1] T. H. Maiman, "Stimulated optical radiation in ruby masers and superconductors," *Nature*, vol. 187, pp. 493–494; August, 1960.
[2] A. Javan, et al., "Research breakthroughs in optical masers and superconductors," *Bell Labs. Record*, vol. 39, pp. 83–86; March, 1961.
[3] A. G. Fox and T. Li, "Resonant modes in a maser interferometer," *Bell Sys. Tech. J.*, vol. 40, pp. 453–488; March, 1961.
[4] S. Silver, "Microwave Antenna Theory and Design," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 12, chs., 4, 5 and 6; 1949.
[5] M. W. P. Strandberg, "Inherent noise of quantum-mechanical amplifiers," *Phys. Rev.*, vol. 106, pp. 617–620; May 15, 1957.
[6] R. N. Schwartz and C. H. Townes, "Interstellar and interplanetary communications by optical masers," *Nature*, vol. 190, pp. 205–208; April 15, 1961.

Moisture Exclusion from Encapsulation of Long-Life Transistors*

R. I. B. COOPER†

Summary—**The requirement for long-life transistors places stringent demands on sealing processes. There are empirical reasons for believing that water content is a critical factor and should be preserved at a very low value.**

The first consequent problem is to design the sealing operation so that the encapsulate is sufficiently dry at the beginning of life. A practicable solution demands the use of moisture getters and the properties and usage of these are compared.

The second problem is to reduce later leakage of water vapor into the can to a low level. The radioactive method of leak detection is shown to be the most satisfactory to use for controlling the quality of the high-grade packages required.

INTRODUCTION

PARTICULAR attention has been focused on the reliability of the transistor as it is the only active component which appears to have no intrinsic wear-out mechanism. Its rival as an amplifier and high-speed switch, the thermionic vacuum tube, eventually fails in emission however well sealed it may have been. Its rival as a low-speed switch, the electromagnetic relay, eventually suffers from contact erosion. An ideal transistor, *i.e.*, one whose operation involves no surface processes, should in principle last almost indefinitely if used within its specified ratings.

The only limitation to the life of such a transistor would be the process of bulk diffusion which would gradually soften the boundaries between the active regions. With the possible exceptions of copper or lithium, whose rate of diffusion in the operational temperature range of transistors may be open to doubt, it can safely be said that the diffusion of all other atomic species in germanium and silicon would be too slow to require consideration in a realistic assessment of life. Progressive dissolution of germanium by indium, a separate phenomenon to be distinguished from diffusion, has been known to terminate the life of some devices manufactured in a special way.[1]

In a practical transistor, surface processes inevitably play some part and, consequently, slow changes in the surface structure due to reaction with the environment progressively alter its characteristics. Eventually, the circuit in which it is being used may fail to operate. The purposes of this article are 1), to argue that the water is the most potent agent for such changes, at least in the case of germanium, and 2), to describe the technology employed to reduce this source of trouble to acceptable proportions.

Although the article concerns itself exclusively with the technology of moisture exclusion from transistors, it cannot be doubted that at least some other components, for example, capacitors and quartz crystals, may eventually require similarly rigorous sealing processes. At present, reliable figures for the limits of acceptable moisture content of these components are not available, nor is any reliably accredited correlation of ageing with moisture ingress forthcoming. This is due partly to the difficulty of ensuring the exclusion of other

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[1] J. Roschen and C. G. Thornton, "Solid-state dissolution of germanium by indium in semiconductor devices," *J. Appl. Phys.*, vol. 29, p. 923; June, 1958.

impurities from affecting the system and partly to the great expense of carrying out comparative life-testing on intrinsically reliable components. Simple *a priori* calculations suggest, however, that quantities of water of the same order, *viz.*, 1 or 2 μg, may be as important as in the case of transistors.

The Importance of Moisture Exclusion

At the outset, three different effects due to the presence of moisture have to be clearly distinguished. The first is the dependence of the characteristics of the device on water vapor content at a fixed temperature. Within quite wide limits, however, it is reasonable to assume that the design of the device could be adjusted to take account of this dependence. In some cases this dependence can be predicted and is often slight; for example, in the case of some *n-p-n* devices. In any case it is a matter readily settled by experiment. Such considerations are thus not definitive of the permissible moisture content.

The second effect is the change in the device performance due to adsorption and desorption of water when the ambient temperature is changed. For example, a device may be stored at −55°C and then suddenly switched on to full dissipation as a result of which its temperature rises in a few seconds to 150°C. Under these conditions, which could occur in a missile application, it must continue to work. It will be seen later that this is, in fact, the consideration which is most amenable to calculating a defined limit to the permissible moisture content.

The third effect is the possible catalytic action of water in promoting surface change as a life-terminating process. Most truly electrical failures of transistors on life-test suffer from faults (*e.g.*, softening of reverse characteristics), which can be attributed to surface deterioration. It should be stressed that it has not invariably been proved that the faults were so caused, and also that all life-test failures have, of necessity, been observed after short times (*i.e.*, a year or two). There is no means of being certain that what has been observed is, in fact, an accelerated version of what will occur to the majority of devices eventually. However, it is clear that attention must first of all be paid to inhibiting surface changes, although once this has been done still other deteriorating processes may afterwards come to light.

It does not seem to be important in this context to distinguish between surface chemical and physical change; any rearrangement of, addition to, or removal from, the surface atomic structure down to a depth not exceeding a Debye length (*i.e.*, a few hundred Angstrom or so), will be called a surface change. Changes envisaged are reaction of the semiconductor with water or oxygen including adsorption or desorption, migration of ionic impurities, and rearrangement of the crystal structure.

The simplest initial approach to reliability is to attempt to restrict as far as possible the rates of any such processes after encapsulation and testing of the device. One step which is usually taken is to bake the device for some hours at an elevated temperature before encapsulation in order to accelerate the processes as much as possible towards equilibrium before the vital selective stage of testing. It is by no means demonstrable that practical baking times and temperatures are adequate to enable equilibrium to be reached. It seems likely that baking may in fact accelerate an approach to a high temperature equilibrium from which slow recovery occurs subsequently at room or operating temperatures. The literature on the behavior of germanium surfaces in this respect is well known.[2] Nevertheless, the weight of evidence seems to be in favor of the point of view that baked germanium devices are nearer to their final characteristics than unbaked ones. The problem is apparently not so severe in the case of silicon devices.

The catalytic effect of water on many chemical reactions has been remarked upon often enough for discussion to be given in many standard works (see, for example, Parkes and Mellor[3]). While the facts in some cases seem to be in dispute still, an example where agreement seems to be general is the rusting of iron.[4] Here it seems that at least a monolayer of water must be present on the surface, probably to allow free migration of ions. Whether similar considerations apply in the case of oxidation of germanium is not known, but as iron and germanium are of similar reactivity it seems quite probable that exclusion of water would inhibit this reaction also. However, such considerations are not yet capable of quantitative expression.

Choice of Encapsulate

The choice of water-excluding ambient lies first of all between solid, liquid and gas. At first sight, the most attractive solution would be to "pot" the transistor, provided that a potting compound of suitably low melting point, viscosity and thermal expansion properties can be found. Considerable effort has been put into this approach, notably with low-melting glasses of the As—T—S and As—T—I systems (see, for example, Kalnin and Pisarchik,[5] and Hare[6]). The mechanical problems have not yet been overcome so that work cannot be begun on an assessment of the permeability or otherwise of the glass "jacket."

In the early days of transistor manufacture the practice of filling the package with a liquid medium (*e.g.*, silicone grease) was common. Such media are not, of

[2] J. T. Wallmark, "Influence of surface oxidation on alpha$_{cb}$ of germanium *p-n-p* transistors," *RCA Rev.*, vol. 18, pp. 255–271; June, 1957.

[3] G. D. Parkes and J. W. Mellor, "Mellor's Modern Inorganic Chemistry," Longmans, Green and Co., London, Eng., p. 284; 1943.

[4] E. de B. Barnett and C. L. Wilson, "Inorganic Chemistry," Longmans, Green and Co., London, Eng., p. 224; 1957.

[5] I. L. Kalnin and A. P. Pisarchik, "Encapsulation of Silicon Diodes in Low Melting Glass," presented at 118th Meeting of Am. Electrochem. Soc. (I. Materials and Processes), Houston, Tex.; 1960.

[6] W. F. J. Hare, "Encapsulation of Silicon Diodes in Low Melting Glass," presented at 118th Meeting of Am. Electrochem. Soc. (II. Electrical Characteristics), Houston, Tex.; 1960.

course, completely impervious. Apart from the practical difficulties involved it did not appear that sufficiently reliable data on the permeabilities of the media existed to enable this method to be used with confidence for devices where the highest reliability was called for. In addition, the possibility of reaction between the medium and the transistor could not be rigorously excluded.

The most practical method is generally acknowledged to be to fill the package with a dry gas and to ensure the exclusion of moisture from subsequent ingress by means of a mechanical seal. One of the practical advantages of such a package is, as will be shown later, that it can easily be subjected to nondestructive testing and hence some confidence can be established in its use. The choice of gas is governed partly by considerations of economy and partly by the consideration that it will generally be the case that some oxygen is already present (as "oxide") on the transistor and thus exclusion of oxygen at great expense would not be warranted. Arguments may be advanced in special cases for a particular ambient which is supposed to produce a surface of given type with desirable electrical properties; however, whether such a special choice would also produce the same desirable properties after several years' ageing of the surface is a matter requiring examination, to say the least.[7]

Consequently, the most usual choice of encapsulate is very dry air. Considerations governing the definition "very dry" will next be considered.

Control of Encapsulated Ambient

In the preceding section the reasons for choosing to encapsulate a long-life transistor in "very dry" air have been given. It will now be necessary to consider how the air is to be introduced into the encapsulation and the choice of method will to some extent govern the quantitative definition of "very dry." There are three basic choices to be considered:

1) After the encapsulation (by a cold process) a tubulation is left through which the package may be exhausted and refilled with air of the desired quality.
2) The act of encapsulation traps only air of the desired quality.
3) The act of encapsulation traps inferior air which is subsequently brought to the desired quality by the action of physical or chemical adsorbents included in the package (moisture "getters").

In each case it has to be considered whether previous or subsequent processing to free the package and its contents from water held at different levels of tenacity can conveniently be carried out.

Since there are limitations to the temperature at which bake-out of mounted transistors can be carried out (about 130°C for a germanium alloy type and 300°C for a diffused silicon type) the principal means of improving dehumidifying processes is by incrasing the time and lowering the ultimate vacuum attainable. From this point of view method 1) would appear to be the most attractive. However, it is also the slowest and most expensive and suitable only for special applicatons (*e.g.*, the Bell 2N559 germanium diffused transistor and A2100 silicon diffused transistor). Process 2) is obviously the most difficult, because the baked-out mounts and cans have to be transferred from their vacuum ovens into an enclosure in which manipulation is possible. Such an enclosure, *e.g.*, a glove-box, will generally not be maintainable at the degree of dryness ultimately necessary.

Thus the most attractive method for quantity production is process 3). This still involves the cans and mounts spending some considerable period inside a glove-box, but any water inevitably enclosed at encapsulation, partly adsorbed on the can and mount during their stay in the box after removal from the vacuum ovens, and already partly contained in the getter should ultimately be retained in the getter. It is still obviously desirable to reduce the demand on the getter as far as possible and so the glove-boxes used should be maintained at as low a humidity as is practically convenient. This is typically below −40°C dew point but cannot be made, *e.g.*, as low as −60°C in operational practice without great difficulty.

The dew point eventually prevailing inside the package will depend on a great many factors connected with the details of the processing; these factors are discussed below. However, it is possible to achieve a figure better than −60°C by good practice.

Humidity Control in Glove-Boxes

To monitor a dew point between −35°C and −45°C at a point in a large enclosure, in which the pressure varies randomly between $\pm\frac{1}{4}$ pound-inch^{-2} of atmospheric, places special restrictions on the type of instrument suitable. If a finite-sampling type of instrument is used (*e.g.*, an element whose electrical capacity or conductivity is affected by the humidity of the ambient with which it is in equilibrium), it must have a recovery time of less than 10–20 minutes (the interval during which a batch of mounts is encapsulated); be relatively insensitive to temperature variations; maintain a stable calibration over long periods; and be robust in recovery from accidental contaminations or exposure to normal atmospheric humidity during servicing of the glove-boxes. If a continuous-sampling type of instrument is used (*e.g.*, an element giving a reading dependent on the rate of arrival of moisture), then a special arrangement will be necessary to withdraw a small, constant, continuous flow of gas from the desired point in the glove-box. In practice, the author has found no element of the finite-sample type with satisfactory ultimate stability and relative insensitivity to ambient temperature changes, and the extra trouble of the con-

[7] J. J. A. Ploos van Amstel has recently described a method of "buffering" the ambient humidity to a constant value. (Colloque International sur les Dispositifs a Semiconducteurs, Paris 1961.) This technique does not dispose of the effects of sudden temperature change nor of slow prolonged water-catalyzed surface changes.

tinuous-sampling system (*e.g.*, the Faraday's Law electrolytic moisture cell)[8] has been repaid in terms of the consistency and reliability of the calibration.

In practice, the humidity in a cluster of glove-boxes may be conveniently recorded at a central station and instant warning is given of the need to repair any one. The limitation of the dew point attainable appears to lie partly in the enforced use of flexible gloves which are permeable and partly in the inevitable leaks which arise at the joints between glove-box and encapsulating engine. Maintenance of a high flow-rate of dry gas is advantageous against both troubles and several satisfactory commercial equipments exist for this purpose. Nevertheless, considerable humidity gradients can exist within a glove-box, and all elements or sampling exit-tubes must be consistently placed in each box with respect to the distributed air-flow. Usually the sampled point will be chosen to be as close as possible to the point where actual encapsulation takes place.

Moisture Gettering

A natural suggestion to make is to use for this purpose a chemical which reacts irreversibly with water; for example, sodium or calcium hydride. Apart from possible technical difficulties which would arise in the handling and introduction of such reactive materials there would always be some danger that the chemical could escape inside the package and contaminate the device. A way around this would be to use a chemical reacting only at a high temperature and to fire the getter as in vacuum-tube practice. However, the noxious vapors which might be given off would exclude this possibility for transistors. Furthermore, as will be seen later, it is the capacity to continue to adsorb water exuded slowly from the device or leaking in through the seal which is the criterion of a desirable getter.

Physical adsorbents such as silica gel, Vycor glass or molecular sieve are therefore preferred.

The desirable properties of a practical moisture getter of this type may be listed as follows:

1) as high a ratio as possible of moisture content to equilibrium water vapor pressure;
2) maintenance of this property up to temperatures well in excess of the operating temperature;
3) sudden loss of this property at temperatures still higher, but not inconveniently high, to facilitate activation;
4) relaxation time of several hours, to permit ease of loading, but no longer, so that the transistor reaches equilibrium in time for reliable final testing;
5) mechanical properties suitable for the getter to be easily inserted and kept clear of the transistor itself.

In establishing criteria for the prime qualities, 1) and 2), the total situation inside the package must be considered. The getter will have to compete in a dynamic situation with the other materials present for the available water. To predict the outcome, the adsorption isotherms of all these materials would have to be known at all relevant temperatures and the starting condition would need to be specified. Furthermore, even if there are no leaks, tenaciously occluded water may seep into the package from the walls and the glass of the seal over periods of years. Provided that the capacity of the getter is greater by a large factor than the total of all water present, or likely to be added, then the only thing that matters is the competition between the getter and the semiconductor surface. Probably no getter is ultimately so tenacious as a perfectly clean silicon or germanium surface for which the binding energy of a water molecule might be of the order of 1 eV. Application of the Frenkel equation[9] gives mean times of stay of the order of seconds and, assuming unit sticking coefficient, this corresponds to an equilibrium vapor pressure of 10^{-6} Torr or less. It is accordingly out of the question to maintain a clean surface in the package. For real surfaces the binding energy is much less and appeal must be made to data on the variation of electrical performance with ambient water-vapor pressure and temperature. The argument is going to be that if the getter can be arranged to maintain a pressure such that no critical electrical effects arise in the operational temperature range and if it has sufficient reserve capacity in comparison with all other sources of water including the transistor surface, then the problem is solved.

In the case of alloy-junction germanium devices, numerous investigators have made observations on such phenomena,[10–15] largely with a view to accounting for excess reverse currents at the transistor junctions. No very clear criterion emerges from such investigations beyond that of requiring the dew point to be safely less than the lowest temperature of operation. Even then it does not follow without further enquiry that the amount of water present in the package, even if all frozen on to the device (*i.e.*, none on to the can or the stem wires), would cause harmful electrical effects when the device was warmed up again. However, if it is arranged that the getter preserves a partial water-vapor pressure

[8] F. A. Keidel, "Determination of water by amperometric measurements," *Anal. Chem.*, vol. 31, pp. 2043–2048; December, 1959.

[9] J. Frenkel, "Theorie der Adsorption und verwandter Erscheinungen," *Z. Physik*, vol. 26, pp. 117–138; 1924.

[10] J. T. Law, "A mechanism for water induced excess reverse dark current on grown germanium *n-p* junctions," Proc. IRE, vol. 42, pp. 1367–1370; September, 1954.

[11] E. N. Clarke, "Nature of the water-vapour-induced excess current on grown germanium *p-n* junctions," *Phys. Rev.* (Corres.), vol. 99, pp. 1899–1900; September, 1955.

[12] J. T. Law and P. S. Meigs, "Effect of water-vapour on grown germanium and silicon *n-p* junction units," *J. Appl. Physics*, vol. 26, pp. 1265–1274; May, 1955.

[13] A. J. Wahl and J. J. Kleimack, "Factors affecting reliability of alloy junction transistors, Proc. IRE, vol. 44, pp. 494–502; April, 1956.

[14] A. R. F. Plummer, "Observations on the growth of excess current in germanium *p-n* junctions," *Proc. Phys. Soc.* (*London*), vol. 69B, pp. 539–547; May, 1956.

[15] J. J. Forster and H. S. Veloric, "Effect of variations in surface potential on junction characteristics," *J. Appl. Phys.*, vol. 30, pp. 906–914; June, 1959.

equivalent to a dew point below the lowest temperature of storage, such effects need not cause concern. In the remainder of this paper a figure of −60°C dew point will be used for the purpose of illustrative calculations. This pressure must be maintained with the getter at the highest operating temperature of the device, for example, 150°C. This is because sudden cooling might otherwise cause formation of ice on the junctions.

The amount of stray water likely to be present in the package as tenaciously occluded water on metallic surfaces can be assessed to an order of magnitude by multiplying a reasonable estimate of surface area by a thickness of a few monolayers. The result is certainly less than 1 μg. Probably an equal amount of tenaciously occluded water might be desorbed from the glass. The solubility of water in glass has been the subject of numerous researches and an extensive discussion can be found, for example, in Morey.[16] A typical figure is 0.04 per cent by weight, which would give 40 μg of water contained in an average hermetic seal. However, this water is not fully desorbed until temperatures considerably above the softening-point of the glass are reached. The gases either adsorbed or absorbed in near-surface layers and which come off below the softening-point account for only about one or two per cent of this amount,[17] and it would be more reasonable to take this as a measure of the water which the getter has to cope with. Glass appears to be highly impermeable to water. The question of water leaking in is dealt with later.

A getter must accordingly be sought such that, once inside the package, and in an ambient of dew point better than −60°C, the addition of one or 2 μg of water would, after a few hours to be fully adsorbed by the getter, cause only an insignificant change in this dew point. The corresponding partial pressure is 0.008 Torr. This condition should furthermore be met when the getter is at the highest operating temperature quoted for the transistor.

Table I gives some data for three examples of physical-adsorption getters.

The advantage of molecular sieve is clear. If, for example, 10 mg of sieve is activated and contains 120 μg of residual water, at 150°C the equilibrium pressure of water vapor in the package is 0.008 Torr. If 2 μg of water are now added to the system, thus increasing the value of W/W from 12 to 12.2×10^{-3}, the equilibrium pressure increases only insignificantly. If Vycor glass were in use the corresponding amount of moisture would be less than 10 μg and the addition of 2 μg of water would cause an appreciable increase in partial pressure.

It should be noted that the actual amount of water vapor in the encapsulate is minute compared with these amounts. (1 cm^3 at a partial water-vapor pressure of 0.008 Torr contains about 0.01 μg of moisture.) Consequently the partial pressure is related to temperature under virtually isosteric conditions; *i.e.*, constant amount adsorbed. Thus, in the above example, if a temperature of −60°C is approached slowly enough to maintain equilibrium, no ice will form because the molecular sieve progressively removes all excess moisture from the package. Only a sudden rapid fall from +150°C to below −60°C would form a possibly harmful amount of moisture on the transistor.

TABLE I

VALUES OF $W/W^*\times10^3$ FOR THREE ADSORBENTS AT THREE TEMPERATURES—CORRESPONDING TO $p=0.008$ TORR

	25°C	75°C	150°C
Silica Gel[18]	6	1	—
Vycor[19]	2	0.6	—
Molecular Sieve[20]	48	23	12

* W/W = fractional weight of adsorbed water.

The technical problem of "activation," *i.e.*, driving off the moisture to below one per cent by weight, is not difficult. But difficulties appear at the operation of encapsulation. The sieve is appropriately retained in the can and the operation of inserting the can-plus-sieve into the encapsulation tool has to be carried out sequentially in the glove-box used for encapsulation. Consequently, the activated sieve is exposed to an atmosphere with a dew point which we have seen may possibly be as high as −40°C. It becomes of importance to reduce the time of exposure, *i.e.*, to increase the speed of assembly as much as possible. Fortunately, the dynamic behavior of molecular sieve is such as to make the operation feasible. There is considerable advantage in this respect in the use of pieces of sieve of as large a size as is convenient since the process of resorption has been found to obey the usual laws of diffusion and the time-constant for large pieces is accordingly greater.

Ensuring Low-Leakage Encapsulation

Whatever process be used it must not involve such an application of heat as to raise the temperature of the enclosed component dangerously, nor to cause the injection of unwanted contaminate, *e.g.*, soldering fluxes and vapors. In general, this rules out hot-soldering or all-glass encapsulation methods. Reference has already been made to the use of low-melting glasses. Cold-soldering or "push-fitting" has been widely used but now appears to lead to a fairly high percentage of leakers when submitted to the rigors of military humidity-cycling tests.

One of the most successful methods is the cold-weld-

[16] G. W. Morey, "The Properties of Glass" (Reinhold, New York, N. Y.), pp. 87–97; 1954.

[17] R. H. Dalton, "Gases in glass," *J. Am. Ceram. Soc.*, vol. 16, pp. 425–432; December, 1933.

[18] S. S. Hubard, "Equilibrium, data for silica gel and water-vapour," *Ind. Engrg. Chem.*, vol. 46, p. 356; February, 1954.

[19] M. J. Rand, "Moisture Gettering with Porous Vycor," Recent News Paper presented at 118th Meeting of Am. Electrochem. Soc., Houston, Tex.; October 9–13, 1960.

[20] Data Sheets provided by the Linde Div. of the Union Carbide Co., New York, N. Y., Sheet No. 4A-1; 1957.

ing process as described, *e.g.*, by Boswell and Humphrey.[21] In this process, two extremely clean and carefully prepared surfaces of pure copper are brought together under great pressure and distorted in a controlled manner. Sometimes the surfaces are plated. The result can be an almost perfect metal-to-metal join and usually the leakage through such a join is very low indeed, of the order 10^{-11} standard cm³/sec of krypton for an excess pressure of 100 psi. It is necessary to specify the leak in this way since there is still controversy over the dependence of such very small leaks upon the pressures and the nature of the leaking gas.

Projection-welding has also received some attention and is widely used. Great precision is required in the manufacture of the two halves of the case to ensure uniform welding around the periphery of the joint. The method also appears to be open to the objection that hot vapors may be evolved from the metal and retained in the encapsulation.

Leak-Testing Methods

In setting a limit to the permissible leak-rate, not only must a very large extrapolation be made in the time, but also some rather insecurely founded assumptions must be made to connect the leak rate measured in a particular test with the leak rate inwards of water from a damp atmosphere. It has been shown above that 1 or 2 μg is the permissible limit of water ingress over the anticipated life of the transistor which might be put at 60 years, let us say 2×10^9 seconds. If the rate of ingress is constant over such a time the corresponding figure in water molecules per second is 2.5×10^7.

To compare this, for example, with the inward leakage of krypton at 100 psi which was referred to above, the Knudsen formula (see Roberts,[22] for example) could be used, which is appropriate to orifices of dimensions small compared with one mean free path. (It is easy to show that this is the case here.) In this way the equivalent krypton leak rate is found to be about 10^{-10} standard cm³/sec. It is by no means clear that it is legitimate to regard the leak as "an orifice." It seems likely that the process is fast enough to be regarded as kinetically controlled; *i.e.*, that the basic process is the dislodgement of the leaking molecular species from loosely bound sites by molecular bombardment in accordance with the usual laws applicable to gas-phase kinetics. The other process which might be suggested as an alternative, solid-state diffusion, depends on interaction at tightly bound sites between the leaking molecular species and the metal lattice and would be expected to be much slower. Thus extrapolation by means of the Knudsen formula is unlikely to lead to widely erroneous conclusions and the adoption of a limit of "krypton leak rate" of 10^{-11} standard cm³/sec, when testing packages by this method seems to be fairly safe and is widely accepted.

In considering other possible methods of leak detection, some weight should be given to the desirability that what is detected is water leaking into the can, rather than an arbitrary gas, or an outward leak. Any practically useful method should also be nondestructive and permit individual measurement. Methods of leak detection which have been proposed are

1) appearance of bubbles after vacuum test under silicone oil,
2) electrical effect on transistor after pressure test using water, alcohol, or detergent,
3) fluoroscopic detection of dyes,
4) mass-spectrometry,
5) radioactive count.[23]

The following comments are made:

1) This method is surprisingly sensitive provided that the bubbles are viewed at low pressure, *e.g.*, a few centimeters head of oil, to give a volume magnification of several hundred times. However, it probably cannot be pressed as far as the detection of 10^{-11} standard cm³/sec and also is open to the objection that the leak measured is that of air outward.
2) This is the only method which may be made to depend on an inward leak of water, and it also depends on observing the actual property of interest. However, there is no known means of ascertaining the level at which a leak would be detected, and the method can only be confidently used to detect gross leakers. In the sense that it would be wished to remove these in any case, the method is non-destructive, but some damage is almost certainly caused also to the packages of those which pass the test.
3) This is a destructive test and useful only for sampling purposes as the can has to be punctured to insert the dye. It also measures outward leakage of a rather large molecule.
4) This is insufficiently sensitive to measure single transistors but is useful for lot acceptance in cases where other work has established that the percentage of defectives is likely to be low. It depends on outward leakage of an independent gas, *e.g.*, helium, and so may not measure what is required.
5) this is nondestructive, sensitive enough for measurement on individual devices, adaptable for automatic inspection and depends on inward leak.

[21] D. Boswell and R. Humphrey, "Cold pressure-welding and its influence on device design and manufacture," *Proc. IEE*, vol. 106, pt. B, Suppl. 15, pp. 454–458; May, 1959.

[22] J. K. Roberts, "Heat and Thermodynamics," Blackie, London, England, p. 73, *et seq.*; 1933.

[23] P. F. Berry and J. F. Cameron, "The Use of ^{85}Kr for the Measurement of Leaks in Hermetically Sealed Components," presented at Conf. on Use of Radioisotopes in the Physical Sciences and Industry, Paper R.I.C.C. 16, Copenhagen, Denmark; September 6–17; 1960.

The single objection to it is that an independent radioactive gas has to be used, usually krypton.

The half-life of ^{85}Kr is 3.2×10^{8} seconds: it emits a penetrating gamma X ray suitable for detection in a scintillation crystal detector. If a counting rate of about twice background is accepted as corresponding to the limit of detection, *i.e.*, about two counts per second, about 10^{11} molecules of ^{85}Kr could be detected if the counting efficiency is one per cent. At the U.K. AEA's Establishments at Harwell and Wantage, activities of 14.5 curie per liter of ^{85}Kr are employed and soaking pressures of 150 psi are used for periods of up to four days.[24] This is a dangerous concentration and very effective precautions for handling the test gas, and in particular for removing as much of it as possible from the transistors after soaking, have to be ensured. The use of a sodium iodide crystal of minimum possible size together with a band-pass filter reduces the background count considerably. In this way sensitivity can be raised to the level at which a leak of 10^{-13} standard cm^3/sec can be detected.

The equipment is calibrated by preparing a test transistor containing an encapsulate of known dilution derived by simple metering from the same gas source as is used for soaking. Counting efficiences usually turn out to be of the order of one per cent. A small amount of ^{85}Kr is inevitably left adsorbed on the outside of the packages after soaking but, as this gives off β rays without adsorption by the walls of the package, it may easily be distinguished from the Kr85 which has leaked in and a correction may be applied. For unpainted pure OFHC copper cans this correction does not usually amount to more than 0.5 count per second on a Geiger counter having a counting efficiency of three per cent. Dirt films and organic materials (*e.g.*, paint) can adsorb enough krypton to interfere with the test, however.

Counting is easily carried out on a rate meter and many highly satisfactory commercial instruments are available for this purpose. A counting time of 10 seconds is adequate for most purposes.

A complete installation is available commercially and fully automatic counting and sorting equipment can be optionally supplied.

Applications to Quality Control

The Quality Control problem is twofold:

1) to eliminate all but a percentage of leakers from the product given to the customer, insignificant in comparison with the least percentage rejection for other causes, and

2) to monitor continuously the percentage of leakers among the product from the encapsulation process in order to control the latter (*e.g.*, to detect tool-wear).

[24] These facilities have been offered as a service to manufacturers, including Standard Telephones and Cables Ltd.

It is clear that 100 per cent inspection by radioactive leak-testing can fulfill both requirements. However, where cost is an important factor, tests on a sampling basis may have to be considered. A very satisfactory and economical scheme is to test on a sampling basis for purpose 2), using radioactive leak-testing, but to submit the whole product to some cheaper form of leak-elimination test much as a high-pressure water test. It is obvious that the amount of confidence to be placed in this test as a means of guaranteeing the product depends very largely on having a very accurately controlled distribution of leak rates, achievable only by careful design of the sampling system chosen for b). If, for example, a large percentage of the product had leaks in the intermediate range between 10^{-11} standard cm^3/sec and the value detectable by test 1), little confidence could be expressed in the product passed to the customer. Fortunately, radioactive leak-testing shows that for cold-welded packages the percentage of leakers which would escape detection by such an inspection system can be made less than 0.1 per cent.

It is only by large numbers of measurements on single devices that an economical combination of sampling inspection for process control and cheap, 100 per cent inspection for defect elimination can be set up with confidence. For this purpose, the radioactive method alone appears to be satisfactory. By its use it has been shown that only one or two per cent of cold-welded packages have leaks exceeding 10^{-11} standard cm^3/sec, and these are almost entirely gross leakers.

Conclusions

Encapsulation of long-life transistors in a mechanically sealed enclosure containing very dry air has many important advantages, not the least of which is the ability to subject the product to satisfactory quality control. The required dryness of the air can be preserved in practice only by the use of a getter. Molecular sieve has many advantages for this purpose. The capacity of this material is sufficient to adsorb water entering through a leak equivalent to 10^{-11} cm^3/sec of krypton at 100 psi pressure and still maintain dryness of the required degree for over 50 years. Such a leak can be measured satisfactorily on an individual basis only by a radioactive counting method, although other leak-testing methods are adequate for batch-testing once sufficient individual measurements are available.

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Microwave-Carrier Modulation-Demodulation Amplifiers and Logic Circuits*

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Summary—**Modulation-demodulation (M-D) amplifiers are four-terminal unidirectional amplifiers in which amplitude modulators with conversion gain are followed by demodulators. This paper describes wideband video M-D amplifiers (voltage-gain-bandwidth product = 1.5 Gc), low-noise RF M-D amplifiers (signal frequency = 1.2 Gc, voltage-gain-bandwidth product = 70 Mc, noise figure < 3 db), and ultra-high-speed M-D logic circuits (possible pulse repetition rates of the order of 10^9 per second). The wideband video M-D amplifier is analyzed and satisfactory agreement between theory and experiment is obtained.**

I. Introduction

The maximum rate of information flow in digital information handling systems is primarily determined by the maximum pulse rate the components of the system can handle, and the time the pulses are delayed between input and output of the individual components. The maximum pulse repetition rates $(PRR)_{max}$ and the corresponding delay times t_d reported so far for digital circuits are as follows:

	$(PRR)_{max}$ (Mc)	t_d (Nsec)	*Reference*
Transistorized baseband[1] circuits:	250	5	[1]
Tunnel-diode baseband circuits:	100	1.5	[2], [3]
Microwave parametric subharmonic oscillators:	100	6	[4]
Logic circuits using microwave traveling-wave tubes:	700	10	[5], [6]

This paper describes solid-state microwave components capable of handling pulse repetition rates of the order of 1000 Mc with time delays of a fraction of a nanosecond. These components can be used as baseband pulse amplifiers, regenerators, and logic circuits, as well as low-noise RF amplifiers. Experimental results obtained with these circuits and the basic theory of their operation have previously been presented orally [7], [8].

Fig. 1 illustrates the basic configuration used for obtaining amplification. The input signal (frequency $= f_s$, available power $= P_{is}$) and pump (carrier) power (frequency $= f_p > 2f_{s,max}$, available power $= P_{ip} \gg P_{is}$) are applied to an amplitude modulator with gain, and sideband frequencies $f_{p\pm s} = f_p \pm f_s$ (where $P_{p+s} + P_{p-s} > P_{is}$) are produced. The output of the modulator is fed through a high-pass filter (which rejects f_s, but passes f_p and $f_{p\pm s}$) into a demodulator, and output power P_{os} at frequency f_s is obtained. Net gain is produced at the signal frequency ($P_{os} > P_{is}$) as long as the conversion gain of the modulator exceeds any conversion loss in the demodulator. This gain exists between separate input and output ports, and is in general unidirectional.[2]

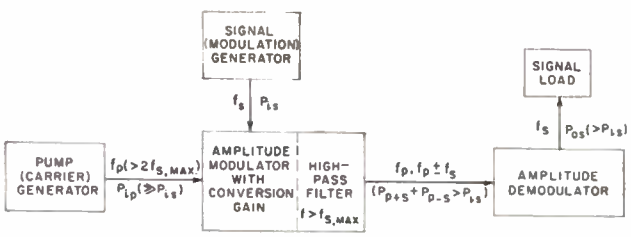


Fig. 1—Block diagram of modulation-demodulation amplifier.

The first part of this paper presents the theory of the basic modulation-demodulation (M-D) amplifier, and discusses bilinear amplitude modulators, lossless and lossy nonlinear-reactance modulators, and linear amplitude demodulators. The application of M-D devices to high-speed logic circuits is then considered, and finally, experimental results obtained with wideband M-D video amplifiers, low-noise M-D RF amplifiers, and ultra-high-speed M-D logic circuits are presented.

While this paper is primarily concerned with M-D devices using pumps at microwave frequencies, pumps outside the microwave range might also prove to be of interest. For example, the microwave light modulators and demodulators currently being developed for optical maser systems can be used in microwave M-D amplifiers pumped with light if sufficient modulator and demodulator sensitivity is achieved.

II. Theory of the Basic M-D Amplifier

The Binlinear Amplitude Modulator

Consider [Fig. 2(a)] a pump generator of frequency f_p and internal resistance R_{gp} feeding power P_{ip} into a lossless transmission line of characteristic impedance

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‡ Electron Tube Div., RCA, Princeton, N. J.

[1] In a baseband system the signal occupies a frequency band starting at or near zero and extending to an upper limit $f_{s,max}$. If the transition from one binary state to the other has a quasi-sinusoidal time dependence, then the spectrum of this transition contains signal frequencies f_s ranging from dc up to a 3-db frequency $f_{s,max} \approx 0.7/t_t$ (t_t = transition time).

[2] A similar amplifier has been described by H. Brett, *et al.*, "A varactor diode parametric standing-wave amplifier," PROC. IRE, vol. 49, pp. 509–510; February, 1961.

$Z_{ip}=R_{gp}$.[3] This line is terminated in port MI of a 3-port network, the internal properties of which will be discussed later. P_{rp} represents the power reflected back into the input line, and P_t the power transmitted out of port MII into a second transmission line (characteristic impedance Z_{om}), which is terminated in the load resistance $R_{lm}=Z_{om}$; a third fraction of P_{ip}, the power $P_{ip}-(P_{rp}+P_t)$, is dissipated resistively within the network.

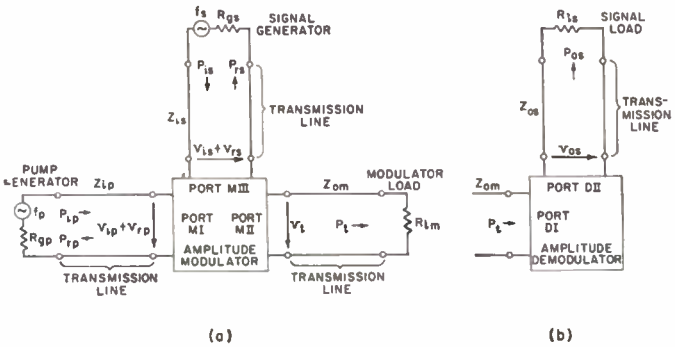


Fig. 2—Equivalent circuits of M-D amplifier input and output.

A signal generator of frequency f_s (where $f_{s,\max}<f_p/2$) and internal resistance R_{gs} feeds power P_{is} into another lossless transmission line (characteristic impedance $Z_{is}=R_{gs}$) connected to port MIII of the network. Power P_{rs} is reflected back, and power $(P_{is}-P_{rs})$ is resistively dissipated in the network. The respective voltages across the three terminal pairs of the network can then be written as follows:

$$v_{ip} = V_{ip}\cos(\omega_p t+\alpha_i), \tag{1}$$

$$v_{rp} = r_p V_{ip}\cos(\omega_p t+\alpha_i+\rho_p), \tag{2}$$

$$v_t = V_t\cos(\omega_p t+\alpha_t), \tag{3}$$

$$v_{is} = V_{is}\cos\omega_s t, \tag{4}$$

$$v_{rs} = r_s V_{is}\cos(\omega_s t+\rho_s). \tag{5}$$

In these equations, r and ρ denote the magnitude and phase angle, respectively, of reflection coefficients, and α denotes the phase angle of the incident and transmitted waves.

If the 3-port network is now specified as a bilinear voltage-controlled amplitude modulator, the voltage V_t varies as a linear function of both V_{ip} and $(v_{is}+v_{rs})$. The properties of this modulator are described by the following equations: (The subscripts u and m denote unmodulated and modulated quantities, respectively.)

[3] If this line is not operated in a TEM mode, the characteristic impedance and the RF voltages are defined such as to describe the power flow in the usual manner.

$$v_{tu} = V_{tu}\cos\omega_p t, \tag{6}$$

$$V_{tu} = aV_{ip}, \tag{7}$$

$$v_{tm} = V_{tm}(t)\cos\omega_p t, \tag{8}$$

$$V_{tm}(t) = V_{tu}\{1+b[v_{is}(t-t_{dm})+v_{rs}(t-t_{dm})]\}. \tag{9}$$

For simplicity, α_i is chosen so that $\alpha_t=0$. In a bilinear modulator, the coefficients a and b are, by definition, constants. We shall call a the (magnitude of the) transmission coefficient, and b the modulation sensitivity. The time t_{dm} is the delay between the application of v_{is} and the resulting change in V_{tm}.

Substitution of (4) and (5) into (9) leads to

$$V_{tm}(t) = V_{tu}\{1+m\cos[\omega_s(t-t_{dm})+\phi_s]\}, \tag{10}$$

where

$$\phi_s = \arcsin\frac{r_s\sin\rho_s}{\sqrt{1+2r_s\cos\rho_s+r_s^2}},$$

and

$$m = b\sqrt{1+2r_s\cos\rho_s+r_s^2}\,V_{is}; \tag{11}$$

m is the degree of modulation, and it is assumed throughout this paper that $m\leq 1$. Spectral resolution of (10) yields

$$v_{tm} = V_{tu}\cos\omega_p t+V_{tu}\frac{m}{2}\cos[(\omega_p-\omega_s)t+\omega_s t_{dm}-\phi_s]$$

$$+V_{tu}\frac{m}{2}\cos[(\omega_p+\omega_s)t-\omega_s t_{dm}+\phi_s]. \tag{12}$$

If it is assumed that Z_{om} and R_{lm} have the same values at the two sideband frequencies $f_{p\pm s}$ as at the pump frequency f_p, the following expressions are obtained for the power at the various frequencies:

$$P_{is}=\frac{V_{is}^2}{2Z_{is}},\qquad P_{ip}=\frac{V_{ip}^2}{2Z_{ip}}, \tag{13}$$

$$P_{tu}=\frac{V_{tu}^2}{2Z_{om}}=P_{ip}a^2\frac{Z_{ip}}{Z_{om}}, \tag{14}$$

$$P_{tm}=P_{tu}+P_{t(p-s)}+P_{t(p+s)}, \tag{15}$$

$$P_{t(p-s)}=P_{t(p+s)}=P_{tu}\frac{m^2}{4}$$

$$=P_{ip}a^2\frac{Z_{ip}}{Z_{om}}\frac{b^2(1+2r_s\cos\rho_s+r_s^2)V_{is}^2}{4}$$

$$=P_{ip}P_{is}\frac{a^2b^2}{2}\frac{Z_{ip}}{Z_{om}}Z_{is}(1+2r_s\cos\rho_s+r_s^2). \tag{16}$$

The transducer gain of the modulator (*i.e.*, the ratio of sideband output power to available signal input power) is then,

$$g_{tm} = \frac{P_{t(p-s)} + P_{t(p+s)}}{P_{is}}$$

$$= P_{ip}a^2b^2 \frac{Z_{ip}}{Z_{om}} Z_{is}(1 + 2r_s \cos \rho_s + r_s^2). \quad (17)$$

Unless $P_{rs}=0$, the actual gain (*i.e.*, the ratio of sideband output power to the part of the available signal input power actually absorbed by the modulator) is, of course, greater than the transducer gain. However, the transducer gain determines the practical utility of the device. (For bandwidth considerations, see Appendix I.)

Eq. (17) shows that the gain of a bilinear voltage-controlled modulator is proportional to the pump power. In principle, therefore, it is always possible to obtain arbitrarily large gain, regardless of how small the transmission coefficient and the modulation sensitivity are, by using sufficiently large pump power. In practice, however, modulators with low modulation sensitivity are of little interest, since they are in general limited by saturation effects and by the noisiness of the pumps.

The Linear Amplitude Demodulator

Amplification with input and output at the same frequency f_s can be obtained by replacing the modulator load R_{lm} with the input port DI of an amplitude-demodulator, as indicated in Fig. 2(b). The demodulator feeds the signal output power P_{os} through a lossless transmission line (characteristic impedance Z_{os}) into the signal load resistance $R_{ls}=Z_{os}$. The demodulator transducer gain is defined by

$$g_{td} = \frac{P_{os}}{P_{t(p-s)} + P_{t(p+s)}} \quad (18)$$

and the total amplifier transducer gain is then given by

$$g_t = g_{tm}g_{td} = \frac{P_{os}}{P_{is}}. \quad (19)$$

Only the case of a passive demodulator (characterized by $g_{td} \leq 1$) will be considered. The condition for net gain is, therefore, $g_{tm} > 1/g_{td}$.

The highest demodulator efficiency (maximum g_{td}) can be obtained with linear rectification, which is possible if $P_t \gg kT2f_{s,\max}$. For appreciable modulator gains, this condition is always satisfied. The transducer gain of linear half-wave and full-wave rectifiers can be evaluated by use of the equivalent circuits shown in Fig. 3. The rectifying elements in these circuits are considered to be ideal, *i.e.*, the voltages defined in Fig. 3(c) are related by the equation

$$v_h = \tfrac{1}{2}(v + |v|), \quad (20)$$

where the subscript h indicates half-wave rectification. Forward losses in the demodulators are represented by the resistances R_d.

The filters in Figs. 3(a) and 3(b) are assumed to be lossless, and the phase-splitting 1:1 transformer in Fig. 3(b) is considered to be lossless and frequency independent within the range $(f_p - f_{s,\max}) \leq f \leq (f_p + f_{s,\max})$. For the lossless ($R_d=0$) full-wave rectifier (subscript f) it follows, therefore, that

$$v_f = |v|. \quad (21)$$

The auxiliary load resistances R_{la} (in cascade with high-pass filters), as well as the load resistance R_{ln} for the negative half-wave [in series with a second rectifying element, Fig. 3(a)], are used to obtain a constant input impedance of the demodulator equivalent circuit so that matched operation can be assumed. If reflections from port DI are permissible, R_{la} and R_{ln} may be omitted.

For the matched case, characterized by

$$Z_{os} = R_{la} = R_{ln} = Z_{om} - R_d, \quad (22)$$

a Fourier expansion of the unmodulated, rectified voltages yields the following results:

$$v_{tuh} = V_{tu} \frac{Z_{os}}{Z_{om}} \left(\frac{1}{\pi} + \frac{1}{2} \cos \omega_p t + \frac{2}{3\pi} \cos 2\omega_p t - \cdots \right), \quad (23)$$

$$v_{tuf} = V_{tu} \frac{Z_{os}}{Z_{om}} \left(\frac{2}{\pi} + \frac{4}{3\pi} \cos 2\omega_p t - \cdots \right). \quad (24)$$

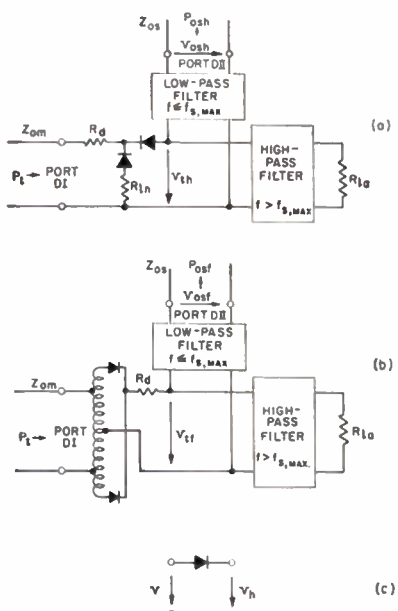


Fig. 3—Equivalent circuits of half-wave (a) and full-wave (b) linear amplitude demodulators. Fig. 3(c), in conjunction with (20), defines v and v_h.

To obtain a spectral resolution of the modulated, rectified voltages, it should be noted that, in general,

$$v_m = v_u[1 + F(t)], \quad \text{with} \quad |F(t)| \leq 1, \tag{25}$$

which gives, by applying (20) and (21),

$$v_{mh} = v_{uhm}, \qquad v_{mf} = v_{ufm}, \tag{26}$$

where the sequence of the subscripts indicates the order in which the processes of modulation and linear rectification are carried out. Therefore,

$$v_{tmh} = V_{tu}\frac{Z_{os}}{Z_{om}}\frac{1}{\pi} + V_{tu}\frac{Z_{os}}{Z_{om}}\frac{m}{\pi}\cos\left[\omega_s(t - t_{dm}) + \phi_s\right]$$
$$+ V_{tu}\frac{Z_{os}}{Z_{om}}\frac{1}{2}\left\{\cos\omega_p t + \frac{m}{2}\cos\left[(\omega_p - \omega_s)t + \omega_s t_{dm} - \phi_s\right] + \frac{m}{2}\cos\left[\cdots(+)\cdots\right]\right\}$$
$$+ V_{tu}\frac{Z_{os}}{Z_{om}}\frac{2}{3\pi}\left\{\cos 2\omega_p t + \frac{m}{2}\cos\left[(2\omega_p - \omega_s)t\cdots\right] + \frac{m}{2}\cos\left[\cdots(+)\cdots\right]\right\} - \cdots, \tag{27}$$

$$v_{tmf} = V_{tu}\frac{Z_{os}}{Z_{om}}\frac{2}{\pi} + V_{tu}\frac{Z_{os}}{Z_{om}}\frac{2m}{\pi}\cos\left[\omega_s(t - t_{dm}) + \phi_s\right]$$
$$+ V_{tu}\frac{Z_{os}}{Z_{om}}\frac{4}{3\pi}\left\{\cos 2\omega_p t + \frac{m}{2}\cos\left[(2\omega_p - \omega_s)t\cdots\right] + \frac{m}{2}\cos\left[\cdots(+)\cdots\right]\right\} - \cdots. \tag{28}$$

The signal-frequency components of these voltages are then given by

$$v_{osh} = V_{tu}\frac{Z_{os}}{Z_{om}}\frac{m}{\pi}\cos\left[\omega_s(t - t_{dm}) + \phi_s\right], \tag{29}$$

$$v_{osf} = V_{tu}\frac{Z_{os}}{Z_{om}}\frac{2m}{\pi}\cos\left[\omega_s(t - t_{dm}) + \phi_s\right]. \tag{30}$$

From these, it follows for the signal output power

$$P_{osh} = V_{tu}^2\frac{m^2}{2\pi^2}\frac{Z_{os}}{Z_{om}^2} = P_{tu}\frac{m^2}{\pi^2}\frac{Z_{os}}{Z_{om}}, \tag{31}$$

$$P_{osf} = V_{tu}^2\frac{2m^2}{\pi^2}\frac{Z_{os}}{Z_{om}^2} = P_{tu}4\frac{m^2}{\pi^2}\frac{Z_{os}}{Z_{om}}, \tag{32}$$

and for the maximum transducer gain of a matched passive demodulator,

$$g_{tdh} = \frac{2}{\pi^2}\frac{Z_{os}}{Z_{om}} \quad (\sim -7 \text{ db for } R_d = 0), \tag{33}$$

$$g_{tdf} = \frac{8}{\pi^2}\frac{Z_{os}}{Z_{om}} \quad (\sim -1 \text{ db for } R_d = 0). \tag{34}$$

The Lossless Nonlinear-Reactance Modulator

There are many devices that approximate the behavior of a bilinear voltage-controlled modulator over certain frequency ranges. However, for ultra-high-speed computing applications, signal frequencies range from dc to about 1 Gc, and the time delay t_{dm} in the modulator has to be minimized. The nonlinear junction capacitance exhibited by semiconductor diodes appears to be the most promising modulation mechanism under such conditions.

In this section a lossless nonlinear-capacitance modulator, as shown in the equivalent circuit of Fig. 4(a), will be considered. The inductance L is assumed to be tuned for parallel resonance (at f_p) with the sum of the circuit capacitance C'' and the average value C_0 of the nonlinear capacitance; C''' represents additional low-frequency circuit capacitance. (Only single-resonance circuits will be considered in this paper.) The coupling of ports MI and MII to the resonant circuit is assumed to be variable.

The power flow at different frequencies in a lossless nonlinear reactor is governed by the Manley-Rowe equations [9]

$$\sum_{m=0}^{\infty}\sum_{n=-\infty}^{\infty}\frac{mP_{mp+ns}}{mf_p + nf_s} = 0, \tag{35}$$

$$\sum_{m=-\infty}^{\infty}\sum_{n=0}^{\infty}\frac{nP_{mp+ns}}{mf_p + nf_s} = 0, \tag{36}$$

(where P_{mp+ns} is the net power absorbed by the nonlinear reactor at frequency $mf_p + nf_s$). Many authors have pointed out that these equations show that double-sideband upconversion (*i.e.*, power input at f_p and f_s, power output at f_p and $f_p \pm f_s$) is a particularly advantageous mode of operation. If only the lower sideband $f_p - f_s$ is allowed to exist, negative resistance is exhibited by the pumped nonlinear reactor at both f_s and

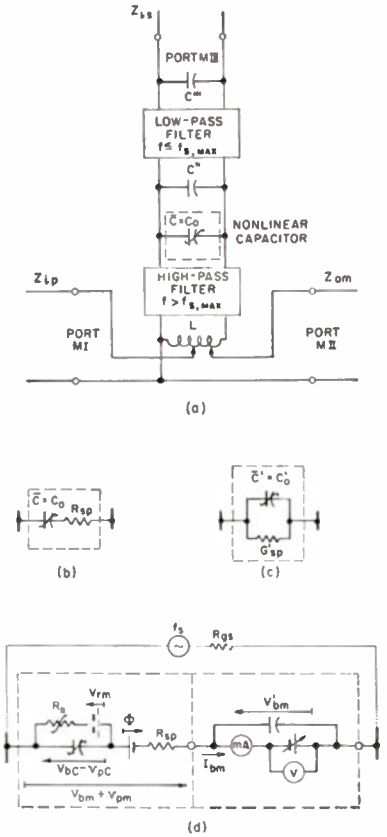


Fig. 4—Equivalent circuits used in the analysis of a nonlinear-capacitance modulator.

$f_p - f_s$, resulting in basically unlimited gain, but also in only conditional stability. If, on the other hand, only the upper sideband exists, the pumped nonlinear reactor represents a positive resistance at both f_s and $f_p + f_s$, but the conversion gain is limited to the ratio $(f_p + f_s)/f_s$. When both sidebands exist in proper proportion, however, it is possible to combine the advantages of the two single sideband cases, *i.e.*, to obtain basically unlimited and yet unconditionally stable gain. This can be shown by specializing the Manley-Rowe equations for the case in which power flows only at the frequencies f_p, f_s, and $f_{p\pm s}$ (*i.e.*, for $m, n = 0, \pm 1$):

$$\frac{P_{p-s}}{f_p - f_s} + \frac{P_p}{f_p} + \frac{P_{p+s}}{f_p + f_s} = 0, \tag{37}$$

$$-\frac{P_{p-s}}{f_p - f_s} + \frac{P_s}{f_s} + \frac{P_{p+s}}{f_p + f_s} = 0. \tag{38}$$

For $P_s = 0$, (37) and (38) assume the following forms:

$$\frac{P_{p-s}}{f_p - f_s} = \frac{P_{p+s}}{f_p + f_s}, \tag{39}$$

$$P_p = -(P_{p-s} + P_{p+s}). \tag{40}$$

Eq. (39) proves that unlimited unconditionally stable gain is indeed possible, because $P_s = 0$ is compatible with sideband powers > 0. Eq. (40) indicates that all of the sideband power leaving the nonlinear reactor (in the case of $P_s = 0$) is supplied by the pump. At this point of the analysis, one difference between a lossless nonlinear reactor, operated as a double-sideband upconverter, and a bilinear voltage-controlled modulator becomes apparent: the latter, according to (16), produces equal powers in both sidebands, regardless of the ratio f_s/f_p, while, for the nonlinear reactor, according to (39), P_{p+s}/P_{p-s} is a function of frequency and approaches unity only if $f_s \ll f_p$. Consequently, distortion-free wideband amplification with a nonlinear-reactance M-D amplifier requires a pump frequency which is considerably higher than the highest signal-frequency component. Gain and bandwidth for given operating conditions can be determined by means of a small-signal theory if it is assumed that $P_{ip} \gg P_{is}$. This assumption is usually valid in practical amplifiers.

A suitable small-signal analysis for the double-sideband upconverter using a lossless nonlinear reactor has been carried out in an unpublished work by Hines [10]. The terms entering into the gain expression derived by Hines require explanation. For a small input signal, application of pump power to the nonlinear capacitance produces an effective time-varying capacitance which (for a proper choice of the time origin) can be written as an even Fourier series:

$$C(t) = C_0 + 2C_1 \cos \omega_p t + 2C_2 \cos 2\omega_p t + \cdots. \tag{41}$$

The following abbreviations are formed from C_0 and the circuit capacitances shown in Fig. 4(a):

$$C_p = C_0 + C'', \tag{42a}$$

$$C_s = C_0 + C'' + C'''. \tag{42b}$$

It is assumed that the coupling of the pump input port MI is so small that its contribution to the total shunt conductance is negligible, while ports MII and MIII are considered to be coupled without impedance transformation. Thus, the shunt conductances are given by

$$G_p = Y_{om} = 1/Z_{om}, \tag{43a}$$

$$G_s = Y_{is} = 1/Z_{is}. \tag{43b}$$

Using these expressions, the following quantities can be defined:

$$Q_p = \frac{\omega_p C_p}{G_p}, \tag{44}$$

$$Q_s = \frac{\omega_p C_s}{G_s}, \tag{45}$$

$$Q_1 = \frac{\omega_p C_1}{G_p}, \tag{46}$$

$$Q_2 = \frac{\omega_p C_2}{G_p}. \tag{47}$$

The transducer gain for the two sidebands is then given by Hines as follows:

$$\frac{P_{t(p\pm s)}}{P_{is}} = 4\frac{G_p}{G_s}Q_1^2 \cdot \left| \frac{j \mp Q_2 + (f_s/f_p)(\pm j - 2Q_p)}{1 - Q_2^2 + j\frac{f_s}{f_p}\left(4Q_p + Q_s + 2Q_1^2Q_2\frac{G_p}{G_s} - Q_2^2Q_s\right)} \right|^2. \quad (48)$$

Eq. (48) can also be written in the following form:

$$\frac{P_{t(p\pm s)}}{P_{is}} = 4\omega_p^2\frac{C_0^2}{G_pG_s}\left(\frac{C_1}{C_0}\right)^2 | \cdots |^2. \quad (49)$$

Because, for all practical cases, the absolute-square factor in (48) and (49) is of the order of unity if $f_s \ll f_p$, it can be concluded from (49) that the midband conversion gain of a lossless nonlinear-capacitance modulator is approximately proportional to the square of the pump frequency, and to the pump power. The proportionaltiy with the pump power is caused by the factor $(C_1/C_0)^2$, because the modulator capacitance is primarily voltage-dependent; this relation will be discussed in further detail later.

The Nonlinear-Capacitance Semiconductor-Diode Modulator

Because nonlinear-capacitance semiconductor diodes are not ideal lossless capacitors, but contain lossy elements, the next step in the analysis is to include the effect of these losses.

The losses in a diode may be either forward-conduction losses or losses due to the spreading resistance R_{sp}, which is in series with the nonlinear capacitance. The forward-conduction losses can be avoided by limiting the voltage across the diode to values which produce no forward conduction. This restriction, however, limits the choice of bias point and pump amplitude, and, therefore, also limits the available gain.

The losses due to the spreading resistance cannot be avoided; these losses are included in the analysis by replacing the dashed box in Fig. 4(a) by the equivalent circuit of Fig. 4(b). For the frequency band $f_p - f_s$ to $f_p + f_s$ (assuming $f_s \ll f_p$) the more convenient equivalent circuit of Fig. 4(c) can be used. With the definition

$$Q_{sp} = \frac{1}{\omega_p C_0 R_{sp}}, \quad (50)$$

one gets

$$G_{sp}' = \omega_p C_0 \frac{Q_{sp}}{Q_{sp}^2 + 1} \quad (51)$$

and

$$C_0' = C_0\frac{Q_{sp}^2}{Q_{sp}^2 + 1}. \quad (52)$$

Transformation (52) applies also to a good approximation to all other Fourier coefficients in (41), so that in general

$$C_\nu' = C_\nu\frac{Q_{sp}^2}{Q_{sp}^2 + 1}, \qquad \nu = 0, 1, 2, \cdots. \quad (53a)$$

The substitution $C_\nu \rightarrow C_\nu'$ leads to the replacement of (42a) and (42b) by

$$C_p' = C_0' + C'' \quad (53b)$$

and

$$C_s' = C_0' + C'' + C'''. \quad (53c)$$

For maximum gain with a lossy diode, the coupling of port MII must be selected so that the loss conductance G_{sp}' is matched to the transform of the load conductance Y_{om}. Eq. (43a) is then replaced by $G_p' = 2G_{sp}'$. If $\omega_{s,\max}$ is a fixed parameter, however, one may have to deviate from the adjustment for maximum gain if this adjustment does not provide sufficient loading of the modulator to obtain the required bandwidth. Approximating the actual bandwidth by the passive bandwidth, this case can be included by putting[4]

$$G_p' = \text{Max}\,[2\omega_{s,\max}C_p' \,|\, 2G_{sp}']. \quad (54)$$

Furthermore, to account for the additional loading of port MIII, (43b) has to be replaced by

$$G_s' = Y_{is} + \omega_s^2 C_0^2 R_{sp}^2, \quad (55)$$

(using the assumption that $\omega_{s,\max}^2 C_0^2 \ll 1/R_{sp}^2$). With (54) and (55) one can now form

$$Q_p' = \frac{\omega_p C_p'}{G_p'}, \quad (56)$$

$$Q_s' = \frac{\omega_p C_s'}{G_s'}, \quad (57)$$

$$Q_1' = \frac{\omega_p C_1'}{G_p'}, \quad (58)$$

and

$$Q_2' = \frac{\omega_p C_2'}{G_p'} \quad (59)$$

to replace (44) through (47), respectively. Finally, to account for the fact that only part of the power de-

[4] $\text{Max}\,[A|B] = \begin{cases} A \text{ if } A > B \\ B \text{ if } B > A \end{cases}$.

livered to G_p' is actual output power, the gain expression must include the factor $(G_p'-G_{sp}'/G_p')$. Thus, the conversion gain obtainable with actual nonlinear-capacitance diodes is determined as follows:

$$\frac{P_{t(p\pm s)}}{P_{is}} = 4\frac{G_p' - G_{sp}'}{G_p'}\omega_p^2\frac{C_0'^2}{G_p'G_s'}\left(\frac{C_1}{C_0}\right)^2\left|\cdots'\right|^2. \quad (60)$$

Because all primed quantities except G_s' are functions of ω_p, the conversion gain of a modulator using a lossy diode is not (as in the lossless case) proportional to the square of the pump frequency. Instead, if the reasonable assumption is made that $Q_{sp}^2 \gg 1$, two regions can be distinguished: as long as ω_p is so low that G_p' has to be made equal to $2\omega_{s,\max}C_p'$, the gain is proportional to $2\omega_{s,\max}C_p\omega_p^2 - C_0^2R_{sp}\omega_p^4$. If, however, ω_p is high enough so that G_p' can be made equal to $2G_{sp}'$, the gain becomes independent of ω_p. The transition between these two regions occurs where

$$\omega_{s,\max}C_p = \omega_p^2C_0^2R_{sp}. \quad (61)$$

This equation defines a critical pump frequency,

$$\omega_{p,\text{crit}} = \sqrt{\frac{\omega_{s,\max}C_p}{C_0^2R_{sp}}}. \quad (62)$$

For $\omega_p < \omega_{p,\text{crit}}$ the modulator does not operate with maximum gain; on the other hand, it is usually not advantageous to choose $\omega_p \gg \omega_{p,\text{crit}}$ because the conversion loss of demodulator diodes increases with increasing pump frequency. Within the limits defined by $Q_{sp}^2 \gg 1$, the conversion gain at or above the critical pump frequency is given by

$$\left.\frac{P_{t(p\pm s)}}{P_{is}}\right|_{\omega_p \geq \omega_{p,\text{crit}}} = \frac{1}{R_{sp}(Y_{is} + \omega_{s,\max}^2C_0^2R_{sp})}\left(\frac{C_1}{C_0}\right)^2\left|\cdots'\right|^2. \quad (63)$$

If the gain expressions are to be useful for numerical calculations, and if the gain limitation caused by forward-conduction is to be included, the coefficients in (41) must be expressed in terms of the actual voltage dependence of the diode capacitance. The notation used is explained in Fig. 4(d), which replaces the former diode equivalent circuit of Fig. 4(b). The right-hand part of Fig. 4(d) shows the biasing circuit external to the diode; V_{bm}' denotes the externally applied bias voltage. The additional elements in the diode equivalent circuit are: the junction contact potential source (Φ), the nonlinear parallel resistance $R_\parallel$, and a virtual source to represent the rectified-pump voltage V_{rm}. A dc (bias) voltage component V_{bm} and a pump-frequency voltage component v_{pm} are across the entire modulator diode; the corresponding voltages across the nonlinear capacitance are V_{bC} and v_{pC}. I_{bm} is the dc (bias) current through the modulator. The following relations hold:

$$V_{bm} = V_{bm}' - I_{bm}R_{gs}, \quad (64)$$

$$V_{bC} = \Phi - (V_{bm} - I_{bm}R_{sp}), \quad (65)$$

$$I_{bm} = \frac{V_{rm} - V_{bC}}{R_\parallel}. \quad (66)$$

It should be noted that I_{bm}, and, therefore, also V_{bm} and V_{bC}, may contain components which are caused by the application of pump power.

The depletion-layer capacitance at the biasing point, C_b, can be written as follows [11]:

$$C_b = C(V_{bm}, I_{bm}) = KV_{bC}^{-n} = K[\Phi - (V_{bm} - I_{bm}R_{sp})]^{-n}, \quad (67)$$

where

$$K = C(0, 0)\Phi^n, \quad (68)$$

and $n = \frac{1}{2}$ for abrupt junctions. This relationship is illustrated in Fig. 5. If pump power is applied, one gets

$$C(V_{bm} + v_{pm}, I_{bm}) = K(V_{bC} - v_{pC})^{-n} = C_b + nC_b\frac{v_{pC}}{V_{bC}} + \frac{n(n+1)}{2}C_b\left(\frac{v_{pC}}{V_{bC}}\right)^2 + \cdots. \quad (69)$$

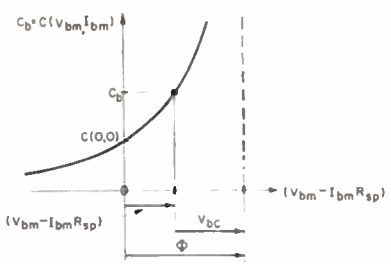


Fig. 5—Graph defining the notation used in the analysis of a nonlinear-capacitance modulator.

With

$$v_{pC} = V_{pC}\cos\omega_p t, \quad (70)$$

(69) becomes

$$C(t) = C_b + nC_b\frac{V_{pC}}{V_{bC}}\cos\omega_p t + \frac{n(n+1)}{2}C_b\left(\frac{V_{pC}}{V_{bC}}\right)^2\frac{1}{2}(1 + \cos 2\omega_p t) + \cdots, \quad (71)$$

or

$$C(t) = C_b\left[1 + \frac{n(n+1)}{4}\left(\frac{V_{pC}}{V_{bC}}\right)^2\right] + nC_b\frac{V_{pC}}{V_{bC}}\cos\omega_p t + \frac{n(n+1)}{4}C_b\left(\frac{V_{pC}}{V_{bC}}\right)^2\cos 2\omega_p t + \cdots. \quad (72)$$

Comparison of (72) with (41) shows that, when terms of higher than second order are neglected,

$$C_0 = C_b\left[1 + \frac{n(n+1)}{4}\left(\frac{V_{pC}}{V_{bC}}\right)^2\right], \tag{73}$$

$$C_1 = C_b \frac{n}{2}\frac{V_{pC}}{V_{bC}}, \tag{74}$$

$$C_2 = C_b \frac{n(n+1)}{8}\left(\frac{V_{pC}}{V_{bC}}\right)^2. \tag{75}$$

It is now necessary to determine the maximum useable value of V_{pC}/V_{bC}. First, it should be noted that the expansion (69) holds only for $V_{pC} < V_{bC}$; also, (67) breaks down if the voltage across the nonlinear capacitance becomes zero or negative. However, these limitations apply to the mathematical description (which is only an approximation), and are not physical restrictions.

An actual physical restriction has been imposed by the requirement that forward conduction losses should be avoided. Because of the rectifier characteristic of $R_{\|}$, the bias current is given by

$$I_{bm}\begin{cases} >0 \\ \approx 0 \end{cases} \quad \text{for} \quad V_{rm} - V_{bC}\begin{cases} >0 \\ \leq 0; \end{cases} \tag{76}$$

when $I_{bm} \approx 0$, peak rectification takes place and $V_{rm} = V_{pC}$. Therefore, the condition for $I_{bm} \approx 0$, or negligible forward-conduction losses, is

$$V_{pC} \leq V_{bC}, \tag{77}$$

and maximum gain will result if $V_{pC} \approx V_{bC}$. We shall introduce this condition by using the following approximation: (73) through (75) (containing only terms up to second order in V_{pC}/V_{bC}) will be retained, but V_{pC}/V_{bC} will be set equal to unity. The resulting relations are (with $n = \frac{1}{2}$, for abrupt junctions),

$$C_0 = 1.19C_b, \tag{78}$$

$$\frac{C_1}{C_0} = 0.21, \tag{79}$$

$$\frac{C_2}{C_0} = 0.079. \tag{80}$$

Substitution of these expressions in the gain equations (60) and (63) makes these equations applicable to practical problems. A comparison between calculated and measured relative performance data for several diodes is given in Section IV, Fig. 15. Also, it can now be seen that, by proper adjustment of the bias voltage, approximately the same amount of small-signal gain should be obtainable with a small pump amplitude as with a large one, as long as the required bias voltage alone does not drive the diode into conduction. The validity of this prediction is demonstrated by the experimental results plotted in Fig. 11.

It should be remembered at this point that, by avoiding the region of forward conduction in our calculations, we have also avoided the region where a reactive mechanism different from the depletion-layer capacitance becomes important: the voltage-dependent capacitance due to the storage of injected minority carriers [11]. Thus, it is theoretically quite possible that, in spite of forward-conduction losses, a further increase in gain will result from the storage-capacitance effect if V_{pC}/V_{bC} exceeds unity. The experimental results indicate that, for some diodes, this is actually the case.

III. Theoretical Considerations of Logic Circuits

Many elementary logic functions, including the three basic ones (AND, INCLUSIVE OR, NOT),[5] can be realized by using modulators and demodulators in simple passive linear circuits [12], [13]. In these circuits, binary ones are represented by the presence of dc or RF pulses in "time slots," and binary zeros by the absence of such pulses.

In the schematic diagram of an AND circuit shown in Fig. 6, each modulator is adjusted to transmit pump power only if a baseband signal is applied. Consequently, there will be an output from the demodulator if, and only if, both input signals are present.

The circuit of Fig. 7, which uses two reflection-type modulators in a balanced arrangement, performs the EXCLUSIVE OR function. This circuit produces an output only if the hybrid is unbalanced, *i.e.*, if an input pulse is applied to one modulator, but not to both. A NOT circuit can be obtained by biasing one input with the voltage representing a binary one, or by replacing one input arm with a network having the same impedance as a modulator biased with a binary one.

Fig. 8 illustrates several logic functions which can be performed by the linear superposition of the RF outputs of modulators (A) and RF bias signals (B). Modulator pumps and bias inputs must have the same frequency and a fixed relative phase, *i.e.*, they must be fed from a common RF source. In most logic circuits of this type, the phase of the resultant RF output (R) depends on the input conditions. Such logic circuits can be cascaded if the RF output is demodulated and the output of the demodulator is used to drive subsequent modulators.

A NOT function is obtained by superimposing the output of a modulator (A) and a constant RF bias (B) which has the same amplitude but opposite phase [Fig. 8(a)]. If $|A| = |B|$, then, because of destructive interference, the resultant output is $R = 0$. If $A = 0$, however, the output is $R = B$. Similarly, if two input signals (A_1 and A_2) are combined 180° out of phase, the EXCLUSIVE OR function is obtained [Fig. 8(b)]. In Fig. 8(c) (INCLUSIVE OR), A_1 and A_2 are combined

[5] In principle, any combinatorial logic circuit can be built using a combination of basic logic circuits.

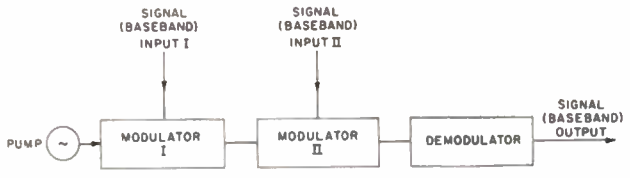


Fig. 6—AND circuit using two cascaded modulators. Each modulator transmits an RF signal only if a baseband signal is applied to it.

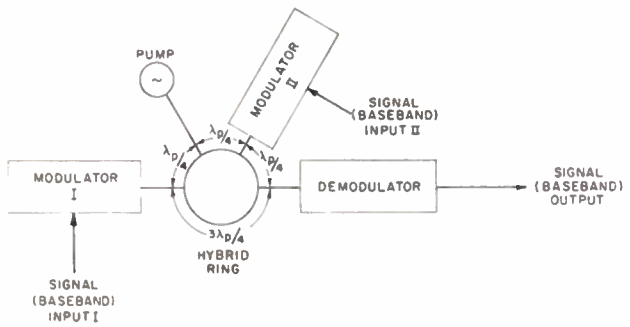


Fig. 7—Logic circuit using two reflection-type modulators in a balanced arrangement.

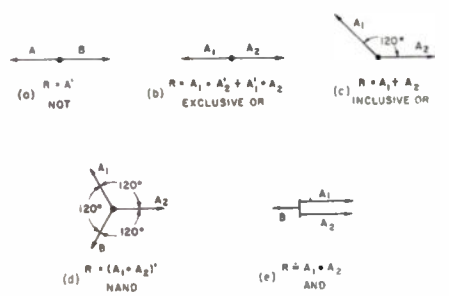


Fig. 8—Examples of logic functions performed by the linear superposition of the RF signal outputs of modulators (A) and RF bias signals (B). R is the resultant RF signal output. The arrows indicate the relative amplitude and phase of the RF signals.

120° out of phase, so that the output has the same amplitude if either one or both inputs are present; if $A_1=A_2=0$, $R=0$. A NAND function is obtained by adding an additional bias signal 120° out of phase with either input [Fig. 8(d)]. In Fig. 8(e) two input signals of equal phase and amplitude are superimposed with a bias signal of opposite phase and half the amplitude of either input signal. In this arrangement, $R=3|B|$ if $|A_1|=|A_2|=2|B|$, and $|R|=|B|$ if one or both inputs are zero. An AND gate can be obtained by rectifying R in a demodulator which is dc-biased to give an output with RF of amplitude $3B$, but no output with RF of amplitude B.

The superposition of RF signals required to perform the logic functions of Fig. 8 can be accomplished in passive linear circuits. However, even if the circuits themselves are lossless, these functions cannot be performed without loss of power, *i.e.*, the available RF input to the circuits must be larger than the resultant RF output. Consider, for example, a linear passive network with N input ports and one output port. Let P_{in} be the available input power at port n, and let $P_{on}=\alpha_n P_{in}$ be the fraction of P_{in} transmitted to the output with all inputs except the nth turned off. Then it can be shown (see Appendix II) that

$$\sum_{n=1}^{N} \alpha_n \leq 1. \tag{81}$$

The application of inequality (81) can be illustrated by reference to the EXCLUSIVE OR and NOT circuits symbolized by Fig. 8(a) and (b). In the EXCLUSIVE OR circuit the available power of the two input signals is usually equal, so that α_1 and α_2 must also be made equal. Inequality (81) then states $\alpha_1=\alpha_2\leq\frac{1}{2}$, *i.e.*, the available power at each input must exceed the output power by at least 3 db. In the NOT circuit [Fig. 8(a)], the input-signal loss can be reduced at the expense of increasing the available power from the RF bias source, *i.e.*, the fraction of signal power transmitted to the output can be made greater than one half by correspondingly decreasing the fraction of transmitted bias power. The loss of RF signal power introduced by any of the logic circuits can, of course, be compensated for by the gain of the modulators.

In the circuits used to perform the logic functions symbolized by Fig. 8, the excess input power is either reflected or absorbed. Usually it is more desirable to absorb the power. Simple circuits of the absorptive type are directional couplers with proper terminations. An advantage of ideal directional couplers is that RF signals can be superimposed with minimum loss ($\alpha_1+\alpha_2=1$).

Regeneration

In applications which require more than a few logic operations, it is essential to reshape the pulses representing the information to prevent serious distortion due to bandwidth limitations and noise. The width and timing of the pulses can be reshaped by use of a pulsed pump supply for the modulators. The pump pulses must then be in synchronism with the pulses representing the information. The amplitude of the pulses can be reshaped by use of the nonlinear gain characteristics of the M-D amplifier for large signals (see Fig. 13).

IV. Experimental Results

Wide-Band Video Amplifiers

Experimental M-D amplifiers were built with 4-, 6-, 10- and 34-Gc pump generators. The amplifiers using the 4-Gc pumps were made of strip transmission line; the other amplifiers used waveguides. A complete amplifier using a 10-Gc pump is shown in Fig. 9. The pump source is a reflex klystron capable of delivering up to 300 mw of RF power. The klystron is follwed by a ferrite isolator which prevents modulation of the klystron output

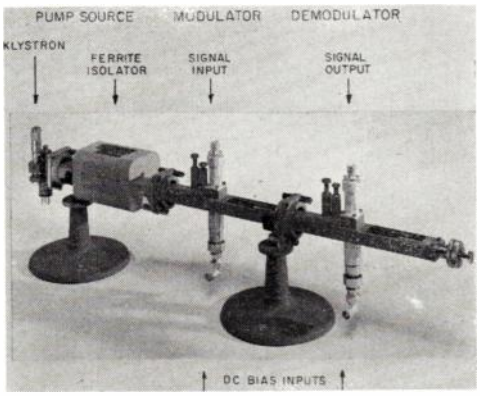


Fig. 9—Photograph of M-D amplifier pumped at X band.

power due to variations of the load impedance.[6] The pump power is fed into the modulator, which consists of a nonlinear-capacitance diode coupled to a short piece of waveguide. The diode is mounted either directly across the guide (as in conventional waveguide detector mounts for diodes of standard 1N21 geometry), or at the end of a short coaxial line which shunts the guide (as in mounts for 1N53 diode geometries). The first method of mounting the diode was used with diodes of the Microwave Associates MA 450/460 series; the second method was found to produce the best results with experimental micro-encapsulated RCA germanium junction diodes [14]. The diode is preceded by two tuning screws which are adjusted for parallel resonance with the diode at the pump frequency. The signal voltage is applied between the waveguide wall and one end of the diode through a band-rejection filter consisting of stagger-tuned microwave chokes. This filter effectively short-circuits the frequency band from $f_p - f_{s,\max}$ to $f_p + f_{s,\max}$, and represents a 50-Ω line for the signal-frequency band. The dc biasing voltage is applied between the waveguide wall and the other end of the diode through a low-pass filter (consisting of a broadband microwave choke and a feed-through capacitor) which provides a very low impedance for the frequency band $f_p - f_{s,\max}$ to $f_p + f_{s,\max}$, as well as for all but the lowest signal frequencies. A bias source having low internal resistance is used to prevent a drop in gain for these low signal frequencies.

The RF power transmitted by the modulator is demodulated in a nonlinear-resistance (rectifier) diode mounted in a second short length of guide. The design of the demodulator is similar to that of the modulator, except that a tuning plunger shorts the end of the guide. The delay time t_d between input and output of the amplifier depends primarily on the distance between the modulator and demodulator diodes, and can be made less than 0.5 nsec.

In the amplifier shown in Fig. 9, the line impedance of both signal input and signal output is 50 Ω. However, both impedances can also be adjusted to any other desired value. For load impedances as low as 50 Ω, the effect of the forward-loss resistance R_d of the demodulator diode becomes important. At X band, and with a 50-Ω load, germanium welded-contact diodes (1N263 and 1N1838) were found to represent the optimum compromise between high cutoff frequency and low forward loss. At higher pump frequencies, and for higher load-impedance levels, silicon point-contact diodes are more advantageous because their rectifying contacts are shunted by lower capacitances than those of the available germanium welded-contact diodes.

Eqs. (33) and (34) suggest the use of a full-wave (push-pull) demodulator. Although the experimental results obtained with such an arrangement agreed reasonably well with the predicted value of g_{tdf}/g_{tdh} (6 db theoretical, 5 db measured), the requirement of two rectifier diodes having closely coincident characteristics cancels most of the advantage because it is usually possible to select a single outstanding diode which performs almost as well as two lesser (but equal) diodes in push-pull.

The performance of the amplifier shown in Fig. 9 is summarized in Figs. 10–15. Fig. 10(a) shows a short-rise-time input pulse to the M-D amplifier, and Fig. 10(b) the resulting output pulse. Both pulses were photographed on a sampling oscilloscope. Figs. 11–14 illustrate the dependence of gain and linearity on pump power, load impedance, input amplitude, and modulator bias. Because the relative effect of these parameters is similar for all the modulator diodes used, a normalized representation was chosen to facilitate the application of these graphs to similar amplifiers. The following absolute performance data were measured (all with the experimental RCA modulator diode listed in Table I and with $P_{ip} = 300$ mw):

Maximum transducer gain between 50-Ω generator and 50-Ω load, with tuning for 0.5 nsec rise time: 6.5 db.

Corresponding voltage-gain-bandwidth product: 1.5 Gc.

Reverse loss under above conditions: 40 db.

Minimum rise time between 50-Ω generator and 50-Ω load, with unity transducer gain: 0.25 nsec.

Maximum transducer gain between 10-kΩ generator and 10-kΩ load: 24 db.

Fig. 15, together with Table I, illustrates the dependence of gain on the modulator diode properties, and at the same time shows to what extent this dependence can be predicted theoretically.

The design of the M-D amplifier operating with a pump frequency of 34 Gc was similar to the unit with X-band pump. Coaxial-line coupling of the modulator diodes to the waveguide was used exclusively. The high-

[6] Isolation between the pump generator and the modulator is required only if a pump generator sensitive to load variations is used, and if the depth of modulation of the pump output is not the same for all signal frequencies.

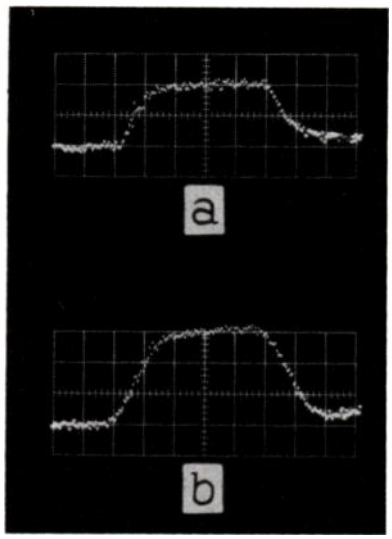


Fig. 10—Photographs of sampling oscilloscope tracings. (a) Input pulse to amplifier (from generator with 50-Ω internal impedance). (b) Output pulse across 50-Ω load. (Horizontal axis 1 division = 1 nsec, vertical axis 1 division = 20 mv.)

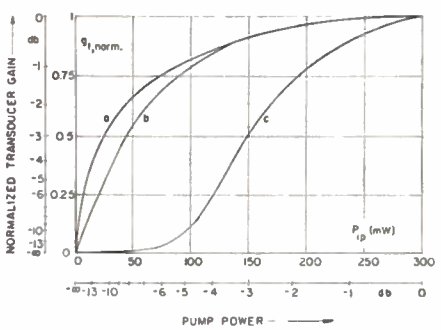


Fig. 11—Plot of normalized transducer gain versus pump power. (a) Tuning of modulator and demodulator and bias voltages readjusted for maximum gain at each pump power level. (b) Tuning adjusted for maximum gain at $P_{ip}=300$ mw, bias readjusted for maximum gain at each pump power level. (c) Tuning and bias adjusted for maximum gain at $P_{ip}=300$ mw. ($R_{gs}=R_{ls}=50\ \Omega$.)

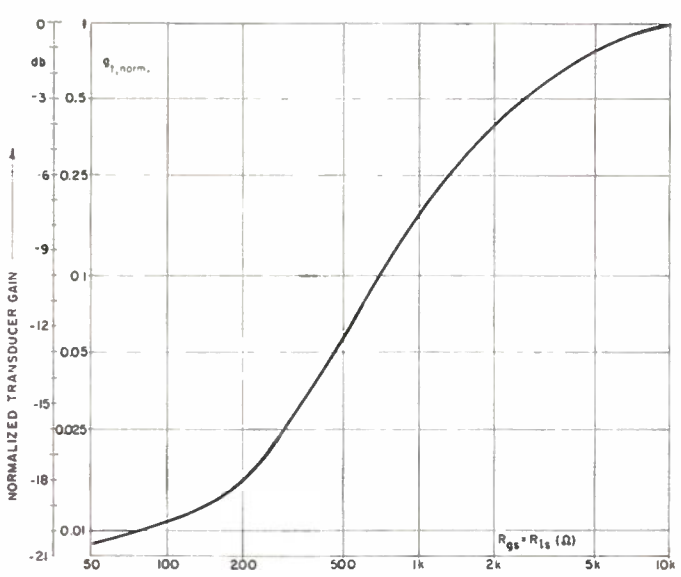


Fig. 12—Plot of normalized transducer gain as a function of generator and load resistance. ($P_{ip}=300$ mw.)

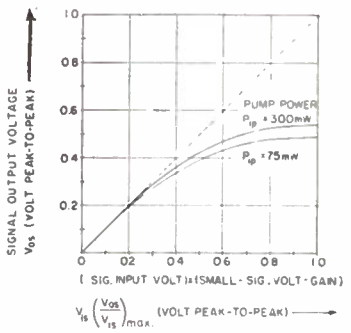


Fig. 13—Plot of normalized amplitude response versus signal input voltage for two values of pump power. ($R_{gs}=R_{ls}=50\ \Omega$.)

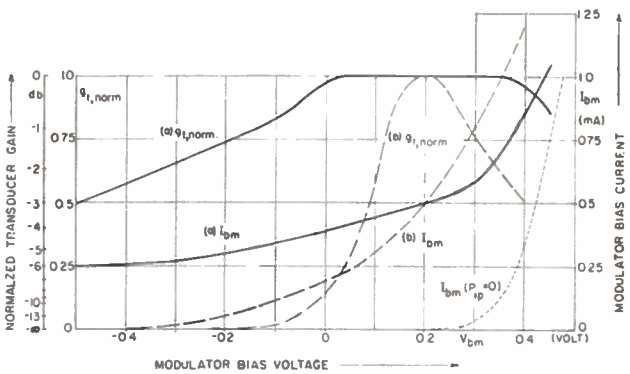


Fig. 14—Plot of normalized transducer gain and modulator bias current as a function of modulator bias voltage ($P_{ip}=300$ mw). (a) Tuning of modulator and demodulator readjusted for maximum gain at each bias voltage. (b) Tuning of modulator and demodulator adjusted for maximum gain at 0.2-volt bias. Also shown is the modulator bias current for $P_{ip}=0$.

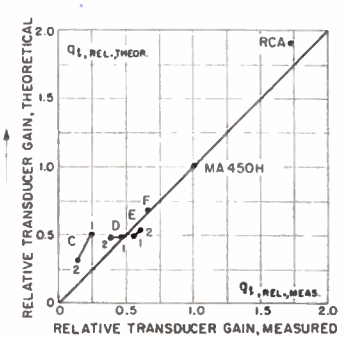


Fig. 15—Plot of theoretical versus experimental relative transducer gain. The theoretical values were calculated from (60). Points for diodes of the type MA450/460 C, D, E, F, and H, as well as for one experimental RCA modulator diode, are shown. (MA450H serves as reference point.)

TABLE I

PROPERTIES OF MODULATOR DIODES USED IN BASEBAND M-D AMPLIFIER

Diode	MA 450/460							RCA
	H	F	E		D	C		
			1	2	1,2	1	2	
$C_b(pf)$	0.6	0.7	1.1	1.2	1.5	2.6	1.6	0.7
Q_{sp}	5.75	3.8	2.1	2.1	1.7	1.25	1.25	8.3

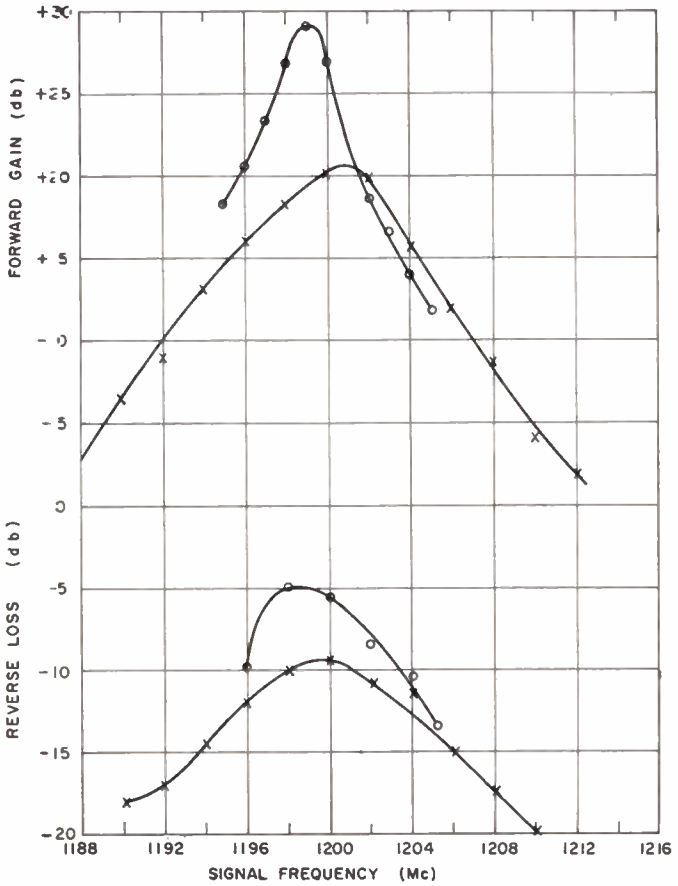


Fig. 16—Plot of forward gain and reverse loss versus signal frequency for two different bandwidth adjustments of a low-noise L-band M-D amplifier.

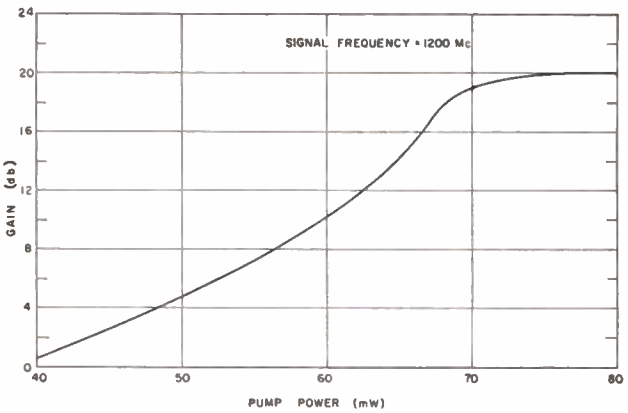


Fig. 17—Plot of gain versus pump power for a low-noise L-band M-D amplifier.

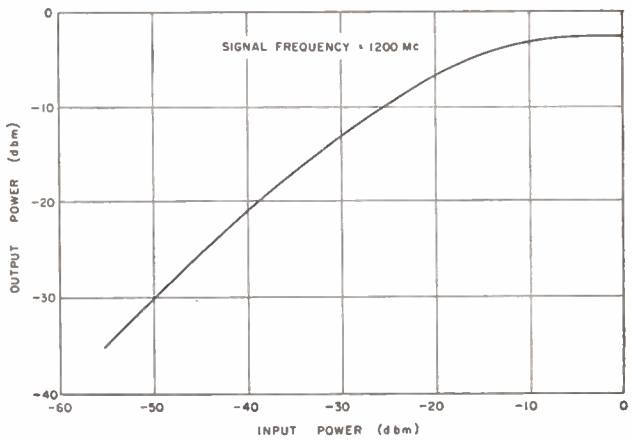


Fig. 18—Plot of input power versus output power for a low-noise L-band M-D amplifier.

est demodulator efficiency at load impedances between 50 Ω and 1 kΩ was obtained with 1N1132 point-contact diodes (coaxial cartridges). The performance of this amplifier was significantly inferior to that of the X-band unit: the maximum transducer gain with 1-kΩ generator and load impedances was 10 db; there was loss with 50-Ω generator and load impedances. The gain reduction is caused partially by the insufficient Q_{sp} of the available modulator diodes, and partially by the high internal resistance of the demodulator diodes suitable for this pump frequency range. A considerable improvement of both modulator and demodulator diodes would be needed, therefore, to make pump frequencies much higher than X-band advantageous for the M-D amplifier.

Low-Noise L-Band Amplifier

The low-pass M-D amplifier shown in Fig. 9 can be converted to a band-pass amplifier if a tuning network is inserted at its signal input. By adjustment of this tuning network, and of the tuning elements associated with the modulator and the demodulator, the center of the pass band of the amplifier can be moved to any frequency up to approximately 2 Gc. No signal input power is reflected from such an amplifier if properly adjusted; therefore, transducer gain and actual gain are equal.

The performance of such a band-pass amplifier, operating at a signal frequency of about 1.2 Gc, is summarized in Figs. 16–18. Fig. 16 shows the forward gain and reverse loss for two different adjustments of the tuning elements. The maximum voltage-gain-bandwidth product measured was about 70 Mc.[7] Fig. 17 shows the power gain as a function of pump power. As in the wide-band case (Fig. 11), the gain of the amplifier is relatively insensitive to pump-power variations at high pump-power levels. The curve of input power as a function of output power in Fig. 18 shows that the amplifier saturates with an input power of about 0.1 mw.

The equivalent noise temperature of the nonlinear-capacitance modulator is small compared to the noise temperature of the nonlinear-resistance demodulator. For minimum noise figure of an M-D amplifier, therefore, the gain in the modulator should be made very high so that the contribution of the demodulator to the over-all amplifier noise figure becomes negligible. For a tuning adjustment giving a gain of 30 db, the measured noise figure was below 3 db.

[7] J. Napoleon (RCA Electron Tube Div., Harrison, N. J.) has obtained a voltage-gain-bandwidth product of about 180 Mc with a similar amplifier operating at a signal frequency of 1 Gc.

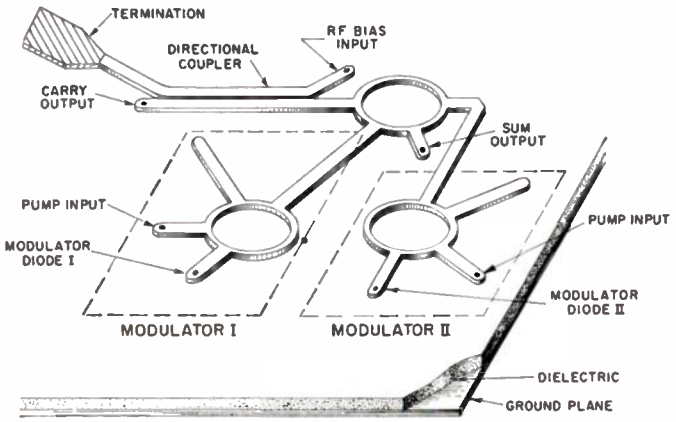


Fig. 19—Experimental half-adder built in strip transmission line. Coaxial connectors for the inputs and outputs, as well as for the modulator and demodulator diodes, are mounted on the rear of the board and are not visible.

Ultra-Fast Logic Circuits

Half-Adder: Fig. 19 shows a schematic diagram of an experimental binary half-adder which performs the logic functions according to Figs. 8(b) and (e). This half-adder is built in strip transmission line. The two inputs, coded in dc pulses, are applied to two nonlinear-capacitance diodes and modulate the RF leaving the hybrids at their output arms. Each modulator transmits RF only if a dc pulse is present. The relative phase of the two pumps is adjusted so that the outputs of the modulators arrive in opposite phase at the circumference of a third hybrid ring. One terminal of this hybrid feeds directly into a demodulator (SUM output). The output of the fourth arm feeds through a directional coupler into another demodulator (CARRY output). An RF bias source is connected to the CARRY output through this same directional coupler. The amplitude and phase of the RF bias source are adjusted so that, at the CARRY terminal, the amplitude due to the bias signal is one half the amplitude of the signal from either modulator, and the phase of the two signals is opposite. The demodulator at the CARRY terminal is dc-biased as described in Section III.

This half-adder was tested with a 4-Gc carrier. The ratios of the ONE output amplitude to the ZERO output amplitude at the SUM and CARRY terminals exceeded 8:1. The rise time was about 2.5 nsec. For faster operation a higher pump frequency would be required.

AND Circuit: The high-speed capabilities of M-D logic circuits were demonstrated by a simple AND gate operating according to Fig. 6 and using X-band carrier components of the type shown in Fig. 9. Output rise times of less than 0.5 nsec were measured. This result indicates that logic can be performed at rates of about 10^9 operations per second.

Memory Element

If a sufficient part of the signal output power of a basic M-D amplifier is fed back with proper polarity into its signal input port, a bistable dynamic memory element is obtained. This element permits nondestructive readout, and can be triggered either in its RF portion (by use of additional modulators) or on the baseband side. These possibilities were demonstrated by an experimental unit using a 6-Gc pump and waveguide circuits.

V. Conclusions

M-D amplifiers combine the advantages of parametric upconverters and negative resistance amplifiers. With only a single parametric element, and without a circulator, the amplifiers have the following characteristics: input and output are at the same frequency, unidirectional gain is obtained between separate input and output ports, and the gain is relatively insensitive to variations in pump power, generator and load impedance.

M-D video amplifiers are particularly suitable for ultra-high-speed computing circuits because they combine large voltage-gain-bandwidth products ($\approx$1.5 Gc) with short delay times ($\approx$0.5 nsec). These amplifiers can, in conjunction with simple passive circuits, perform all logic functions, and, at the same time, regenerate the pulses representing the information. The experimental results obtained with the M-D video amplifier are in satisfactory agreement with analysis.

M-D RF amplifiers feature noise figures of less than 3 db up to frequencies of at least 1.3 Gc. While the voltage-gain-bandwidth product of these amplifiers is only modest ($\approx$180 Mc at f_s=1 Gc), they can be readily tuned over a large fraction of an octave.

Appendix I

Bandwidth Consideration for the Bilinear Amplitude Modulator

If it is assumed that the input admittance of port MIII in Fig. 2 is independent of both the applied pump and signal powers, and that it consists of the conductance $G_{ms}(f_s)=1/R_{ms}(f_s)$, shunted by the susceptance

$\Xi_{ms}(f_s)$, the reflection coefficient (with $Y_{is}=1/Z_{is}$) is given by

$$r_s(f_s) = \frac{\sqrt{Y_{is}^4 + G_{ms}^4 + B_{ms}^4 - 2Y_{is}^2G_{ms}^2 + 2Y_{is}^2B_{ms}^2 + 2G_{ms}^2B_{ms}^2}}{Y_{is}^2 + G_{ms}^2 + B_{ms}^2 + 2Y_{is}G_{ms}},$$

$$\rho_s(f_s) = \arctan \frac{2Y_{is}B_{ms}}{G_{ms}^2 + B_{ms}^2 - Y_{is}^2}.$$

If $B_{ms}(0)=0$, and G_{ms} is independent of f_s or a monotonically increasing function of f_s, maximum gain is obtained for $f_s \to 0$:

$$g_{tm}(0) = P_{ip}a^2b^2 \frac{Z_{ip}}{Z_{om}} Z_{is} \frac{4R_{ms}^2(0)}{[R_{ms}(0) + Z_{is}]^2}.$$

To determine the 3-db bandwidth, one has to solve the equation

$$g_{tm}(f_{s,3\text{ db}}) = \tfrac{1}{2}g_{tm}(0)$$

for $f_{s,3\text{ db}}$. In the case $B_{ms}(0)=0$, and G_{ms} independent of f_s, the problem is reduced to solving

$$B_{ms}(f_{s,3\text{ db}}) = G_{ms} + Y_{is}.$$

Appendix II

Proof of Inequality (81)

Consider a general linear (reciprocal or nonreciprocal) passive network with two input ports and one output port. The input ports are fed by two independent coherent power sources of frequency f and arbitrary phase. The output port is connected to a load resistor R_l. For $n=1,2$, the following powers can be defined:

P_{in} = the available power of the nth source;
P_{on} = the power dissipated in R_l, when only source n delivers power to the network, and the other source is replaced by its internal impedance;
$P_{o\Sigma}$ = the power dissipated in R_l, when both sources deliver power to the network.

The voltages across R_l can be written as follows:

$$v_{on} = V_{on}\cos(\omega t + \psi_n) = \sqrt{2P_{on}R_l}\cos(\omega t + \psi_n), \quad (82)$$

$$v_{o\Sigma} = V_{o\Sigma}\cos(\omega t + \psi_\Sigma) = \sqrt{2P_{o\Sigma}R_l}\cos(\omega t + \psi_\Sigma). \quad (83)$$

Application of the superposition theorem yields

$$V_{o\Sigma} = \sqrt{V_{o1}^2 + V_{o2}^2 + 2V_{o1}V_{o2}\cos(\psi_1 - \psi_2)}, \quad (84)$$

and, therefore,

$$P_{o\Sigma} = P_{o1} + P_{o2} + 2\sqrt{P_{o1}P_{o2}}\cos(\psi_1 - \psi_2). \quad (85)$$

From conservation of energy, it follows that

$$P_{o\Sigma} \leq P_{i1} + P_{i2} \quad (86)$$

for arbitrary values of $\psi_1 - \psi_2$; therefore,

$$P_{i1} + P_{i2} \geq P_{o1} + P_{o2} + 2\sqrt{P_{o1}P_{o2}}. \quad (87)$$

Let

$$\frac{P_{on}}{P_{in}} = \alpha_n \qquad (0 \leq \alpha_n \leq 1) \quad (88)$$

and

$$\frac{P_{i2}}{P_{i1}} = K \qquad (0 \leq K \leq \infty). \quad (89)$$

The α_n's are circuit constants because of the linearity of the network, and K is an arbitrary constant because of the independence of the sources. Inequality (87) can then be rewritten as follows:

$$1 \geq \alpha_1 + (\alpha_2 - 1)K + 2\sqrt{\alpha_1\alpha_2}\sqrt{K}. \quad (90)$$

The right-hand side of inequality (90) is a maximum for

$$K = \frac{\alpha_1\alpha_2}{(1-\alpha_2)^2}. \quad (91)$$

Substitution of (91) into (90) yields

$$1 \geq \alpha_1 + \alpha_2. \quad (92)$$

The output of the 3-port network discussed so far can be used as one of the inputs of a second, similar 3-port network, and so on. By iteration of the above argument, inequality (92) can thus be generalized for an $(N+1)$-port network with N inputs and one output:

$$1 \geq \sum_{n=1}^{N} \alpha_n, \quad \text{q.e.d.}$$

VI. Acknowledgment

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References

[1] A. K. Rapp and J. L. Robinson, "Rapid-transfer principles for transistor switching circuits," Digest of Technical Papers, 1961 Internatl. Solid-State Circuits Conf., Philadelaphia, Pa.; February, 1961.
[2] W. V. Harrison and R. S. Foote, "High speed switching circuitry using tunnel diodes," Digest of Technical Papers, 1961 Internatl. Solid-State Circuits Conf., Philadelphia, Pa.; February, 1961.
[3] J. R. Turnbull, "100-Mc nonsynchronous Esaki-diode computer circuitry," Digest of Technical Papers, 1961 Internatl. Solid-State Circuits Conf., Philadelphia, Pa.; February, 1961.
[4] F. Sterzer, "Microwave parametric subharmonic oscillators for digital computing," PROC. IRE, vol. 47, pp. 1317–1324; August, 1959.
[5] W. C. G. "Nanosecond logic by amplitude modulation at X-band," IRE TRANS. ON ELECTRONIC COMPUTERS, vol. EC-8, pp. 265–271; September, 1959.
[6] W. Sauter and P. J. Isaacs, "Microwave logic circuits using diodes," IRE TRANS. ON ELECTRONIC COMPUTERS, vol. EC-8, pp. 302–307; September, 1959.
[7] W. Eckhardt and F. Sterzer, "A modulation-demodulation scheme for ultra-high-speed computing and wide-band amplification," Digest of Technical Papers, 1960 Internatl. Solid-State Circuits Conf., Philadelphia, Pa.; February, 1960.
[8] W. Eckhardt and F. Sterzer, "A Four-Terminal Low-Noise Parametric Microwave Amplifier," presented at the PGMTT National Symposium, Coronado, Calif.; May, 1960.
[9] 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.
[10] M. E. Hines, "Three-Frequency Theory of Amplification in Nonlinear-Reactance Modulators," unpublished Bell Telephone Lab. Memorandum, Murray Hill, N. J.
[11] W. Shockley, "The theory of *p-n* junctions in semiconductors and *p-n* junction transistors," *Bell Sys. Tech. J.*, vol. 28, pp. 435–489; July, 1949.
[12] W. D. Lewis, "Microwave Logic," presented at the International Switching Symposium, Harvard University, Cambridge, Mass.; April, 1957.
[13] D. J. Blattner and F. Sterzer, "Fast microwave logic circuits," IRE TRANS. ON ELECTRONIC COMPUTERS, vol. EC-8, pp. 297–301; September, 1959.
[14] J. Hilibrand, *et al.*, "Semiconductor parametric diodes in microwave computers," IRE TRANS. ON ELECTRONIC COMPUTERS, vol. EC-8, pp. 287–296; September, 1959.

An Analysis of a Class of Variable-Frequency Sinusoidal Oscillators Using a Single Active Element*

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***Summary*—A class of variable-frequency sinusoidal oscillators requiring only a single active element is analyzed. The frequency of oscillation is approximately proportional (directly or inversely) to the square root of the g_m of the active element. The analysis is carried out in two steps. The first step allows one to write the relation between the frequency and the circuit parameters almost by inspection, and also to determine easily an upper bound on the loading. The second step is a more detailed analysis yielding certain circuit constraints which must be met in order that the relations determined in the first step be valid. Two experimental circuits are described and compared to the theory.**

THERE ARE many applications for an oscillator whose frequency may be varied electronically. Such applications include frequency modulators, automatic frequency control loops, and sweep-frequency generators. Several types of electronically variable oscillators have found use in the present day art, the more notable types being the reactance tube controlled oscillator,[1] the RC oscillator,[2] and the multivibrator.[3] These oscillators require, in general, more than one active element, and in the relaxation oscillators a filter is required to convert a nonsinusoidal oscillation into a sinusoid.

This paper describes a class of variable frequency sinusoidal oscillators whose implementation requires just one active element with an electronically variable mutual conductance, g_m. The frequency of oscillation is generally proportional (inversely or directly) to the square root of the g_m of the active element, and may be made to vary over an octave or more. Similar oscillators have been previously proposed by Chang and Rideout.[4]

The general form of this oscillator is shown in Fig. 1. Z_1, Z_2 and Z_3 are RLC networks with the condition that

* Received by the IRE, September 8, 1961.
† Bell Telephone Labs., Inc., Murray Hill, N. J.

[1] C. Travis, "Automatic frequency control," PROC. IRE, vol. 23, pp. 1125–1141; October, 1935.
[2] C. K. Chang, "A frequency-modulated resistance-capacitance oscillator," PROC. IRE, vol. 31, pp. 21–25; January, 1943.
[3] S. Bertram, "The degenerative positive bias multivibrator," PROC. IRE, vol. 36, pp. 277–280; February, 1948.
[4] H. Chang and V. C. Rideout, "The reactance tube oscillator," PROC. IRE, vol. 37, pp. 1330–1331; November, 1949.

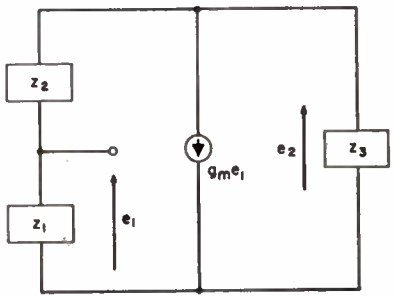


Fig. 1—General form of variable reactance oscillator.

at least one element in Z_1 be an inductor and one element in Z_2 be a capacitor, or vice versa, and that there be at least one resistor in the Z_1-Z_2 network. Z_3 contains at least one reactance of the appropriate type (to be described later), the oscillator load, and a nonlinear element to limit the amplitude of oscillations. The active element, as shown, has a high input and output impedance and a variable g_m. If the terminal impedances of the active element are not infinite, they may be lumped with Z_1, Z_2 and Z_3. The circuit is arranged as a class-A oscillator—that is, the limiting in Z_3 is such that the active element is always in the active region. Furthermore, the amplitude of oscillation should be such that g_m remains approximately constant over a cycle (if g_m is a function of the control voltage) to reduce distortion.

This paper treats a simpler subclass of these oscillators, namely those cases in which either Z_1 or Z_2 is a series RX combination and the other a parallel RX combination, where the reactive elements are of opposite types as stated above. The general ideas behind the theory of operation of these oscillators, their significant features, and a method of detailed analysis will be described. Finally, two realizations of these oscillators will be analyzed and the analytic results compared to experiment.

Theory of Operation

The general approach to be taken in the analysis of this class of oscillators requires two steps. The first step involves making an approximate analysis of the circuit to determine just what is the equation relating g_m and the frequency of oscillation, ω. As will be seen, this relation will hold only if certain conditions are placed on the circuit parameters. The second step is to make a detailed analysis of the circuit and determine, from this rather complicated expression for ω, just what circuit conditions must hold in order that it reduce to the simple desired relation determined by the first step. One also gets, as a by-product of either analysis, an upper bound on the load which the oscillator may drive.

Step One: The Approximate Analysis

The general approach to be taken in the approximate analysis is shown in Fig. 2(a). The portion of the circuit

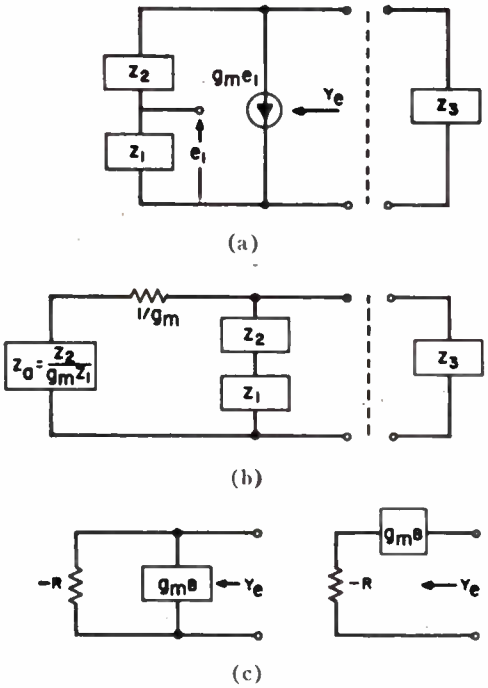


Fig. 2—Equivalent circuits for the variable reactance oscillator.

to the left of the break, including Z_1, Z_2, and the active element, will be replaced by its equivalent admittance Y_e, which is

$$Y_e = \frac{g_m Z_1 + 1}{Z_1 + Z_2} = \frac{1}{\frac{1}{g_m} + \frac{Z_2}{g_m Z_1}} + \frac{1}{Z_1 + Z_2}.$$

This can be represented by the circuit of Fig. 2(b), where $Z_a = Z_2/g_m Z_1$ will be referred to as the active impedance. It will be shown in this section that, for the class of circuits we are considering, Z_a reduces to one of the circuits of Fig. 2(c). That is, Z_a appears as a negative resistance in parallel or series with a susceptance which is proportional to g_m. Then, if the circuit parameters can be chosen such that Z_a and Z_3 are the dominant impedances in the circuit of Fig. 2(b), the oscillator circuit may be reduced approximately to Z_a in parallel with Z_3. In this case, if Z_3 contains a positive resistance load and a susceptance opposite to that contained in Z_a, the circuit will oscillate. The frequency of oscillation will depend on g_m because one of the susceptances is proportional to g_m.

The method of determining the appropriate circuit conditions so that this approximation holds will be discussed later. We first study Z_a.

One-Element Networks for Z_a: To get a feel for what Z_a becomes for different Z_1 and Z_2, let us tabulate Z_a for the various possible cases when Z_1 and Z_2 are single elements (Fig. 3). From this figure it can be seen that the following rules describe Z_a when Z_1 and Z_2 are *one-element networks:*

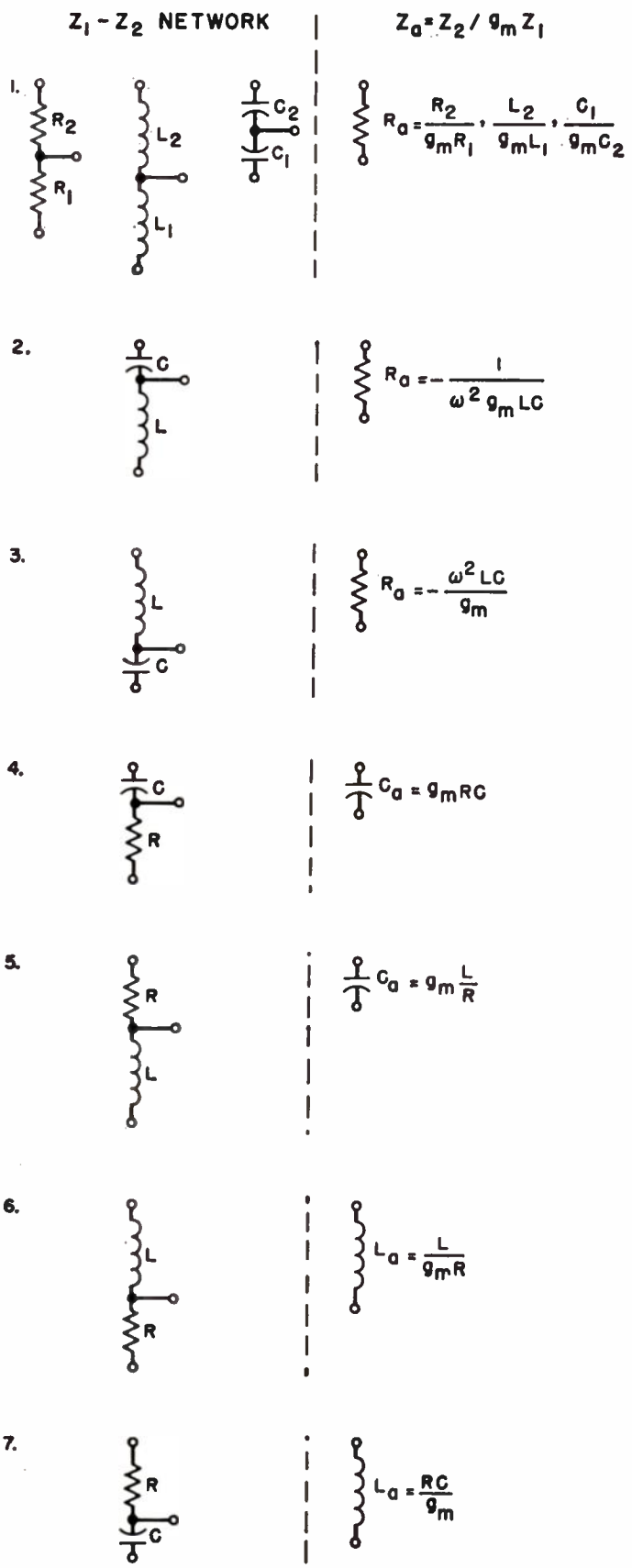


Fig. 3—Z_A for one-element networks.

R1) Like elements cause a positive resistance.
R2) Opposite reactances cause a negative resistance.
R3) All combinations cause an admittance proportional to g_m.
R4) In an RX combination, if the reactance is in the Z_2 position, a like reactance is reflected; if it is in the Z_1 position, an opposite reactance is reflected.
R5) In an RX combination, if Z_a is a capacitor, the capacitance is $g_m\tau$; if Z_a is an inductor, the inductance is τ/g_m, where τ is the time constant of the Z_1-Z_2 network.

It will now be seen that these rules hold for more complicated networks.

Two-Element Networks for Z_a: As mentioned in the Introduction, we will be concerned only with those oscillator circuits in which:

1) Z_1 and Z_2 each contain a resistance and a reactance.
2) The reactance contained in Z_1 is of the opposite type as that contained in Z_2.
3) One network is a series network, and the other a parallel network.

The computation of Z_a for these networks is straightforward, and only the results will be given. There are four possible cases, which are tabulated in Fig. 4. In all cases, the equivalent circuit elements obey the rules [R1)–R5)] given in the previous section, if the following rules are added:

R6) Each combination of one element from Z_1 and one element from Z_2 yields an appropriate element in Z_a.
R7) The connectivity (series or parallel) of Z_a is of the same form as Z_2, and of the opposite form of Z_1.

Now, if the circuit parameters are chosen properly, as described in the next section, the equivalent circuit can be approximated by the parallel connection of Z_a and Z_3. Furthermore, if only one resistor is used (either Z_1 or Z_2 a pure reactance), then the positive resistance in the circuit equivalents for Z_a (Fig. 4) disappears, and Z_a simplifies to a combination of two elements, a negative resistance and an admittance proportional to g_m, as previously indicated in Fig. 2(b).

If a load (Z_3) consisting of an opposite reactance, a limiter,[5] and a load resistance (connected in parallel if Z_a is a parallel network, series if Z_a is series) is connected across Z_a, then the circuit will oscillate. The amplitude of the oscillations will adjust so that the magnitude of the negative resistance equals the magnitude of the

[5] It is important to prevent the limiter from appearing in the Z_1–Z_2 network. Otherwise, the varying resistance of the limiter over a cycle would introduce a like variation in an equivalent susceptance over a cycle, thus causing distortion.

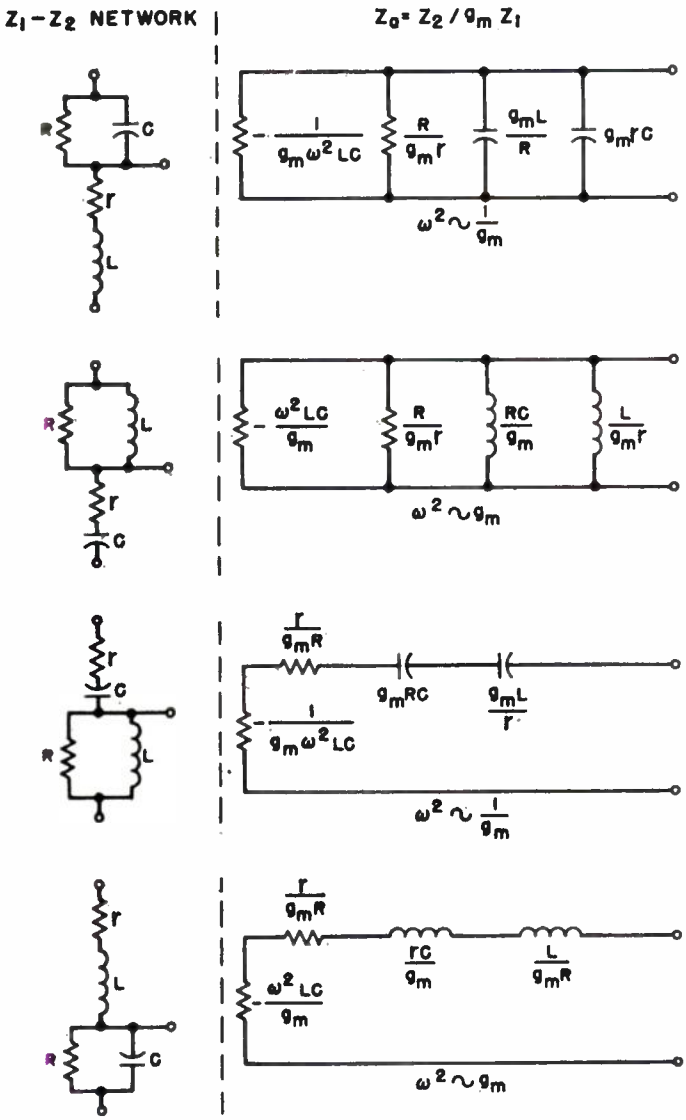


Fig. 4—Z_A for some two-element networks.

equivalent load resistance, and the dependence of frequency on g_m is as indicated under each circuit in Fig. 4. Note that this variation is such that the negative resistance is not a function of g_m, for if the relation of ω^2 on g_m is substituted into the negative resistance expression in each network, the g_m's cancel out. That is, if the circuit conditions are such that Z_a and Z_3 are the dominant impedances, and Z_3 is as described above, the negative resistance does not depend on the gain of the active element. This generally occurs when g_m is sufficiently large, as will be seen in the detailed analyses given in the appendixes.

For reasons which we hope are apparent from this discussion, we suggest the coining of the phrase "variable reactance oscillator," or VRO, to describe these circuits.

Step Two: The Detailed Analysis

From the approximate analysis described so far, one can easily obtain the frequency of oscillation of the VRO, the dependence of frequency on g_m, and the maximum load which the oscillator will drive (given by the magnitude of the negative resistance). This entire approximate analysis is based on the assumption that the impedances other than Z_a and Z_3 in Fig. 2(b) can be neglected. To determine the conditions under which this assumption is valid, a more detailed analysis of the circuit must be made. One method for doing this will now be described.

Consider the representation of Fig. 1. A pair of node equations in e_1 and e_2 can be written

$$\left(\frac{1}{Z_1} + \frac{1}{Z_2}\right) e_1 - \frac{1}{Z_2} e_2 = 0$$

$$\left(-\frac{1}{Z_2} + g_m\right) e_1 + \left(\frac{1}{Z_2} + \frac{1}{Z_3}\right) e_2 = 0.$$

Since this is a pair of homogeneous equations, e_1 and e_2 will be zero unless the determinant is zero. Therefore, the conditions of oscillation (e_1 and e_2 not zero) are found by setting the determinant equal to zero. Doing this and simplifying, we obtain

$$g_m Z_1 Z_3 + Z_1 + Z_2 + Z_3 = 0. \tag{1}$$

Alternately letting

$$Y_i = 1/Z_i, \qquad i = 1, 2, 3,$$

(1) becomes

$$g_m Y_2 + Y_1 Y_2 + Y_1 Y_3 + Y_2 Y_3 = 0. \tag{2}$$

This determinant will be complex, and therefore both its real and imaginary parts must be separately zero. This gives two equations, from which the total load resistance can be eliminated, leaving an equation in ω^2 and the circuit constants. The solution for ω^2 will in general be quite a complicated expression, but we know from the approximate analysis what it should be. Consequently, the conditions under which this complicated expression reduces to the approximate expression for ω^2 can be determined, giving the circuit conditions as mentioned previously.

Two examples of this type of two-step analysis are given in the following sections.

A VRO Using a Field Effect Transistor

In Fig. 5 is shown a form of a VRO using a field effect transistor[6] as the active element. Briefly, a field effect transistor is a three-terminal semiconductor device with high input and output impedances (measured in the megohms), and a g_m which is typically of the order of

[6] W. Shockley, "A unipolar field effect transistor," Proc. IRE, vol. 40, pp. 1365–1376; November, 1952. See especially p. 1374.

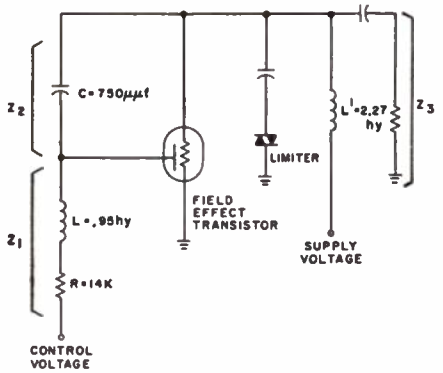


Fig. 5—A VRO using a field effect transistor.

1000 μmhos. The g_m is a function of the voltage on the control element (which is denoted the gate), and may be varied over a wide range (such as 10 to 1).

The equivalent circuit of this VRO is shown in Fig. 6(a). R_L' represents the parallel combination of the fixed resistive load, R_L, and the limiter.

The complete equivalent circuit [see Figs. 2(b) and 4, or rules R1)–R7)] is shown in Fig. 6(b); the approximate form of this is shown in Fig. 6(c). From Fig. 6(c), we see that, under the appropriate circuit conditions,

$$\omega^2 = \frac{1}{g_m RCL'} \tag{3}$$

$$R_e(Z_a) = R_a = -R\frac{L'}{L}. \tag{4}$$

Note that (4) is independent of g_m. The total load resistance R_L' must be equal to the magnitude of the negative resistance. Since the load resistance is made up of the parallel combination of R_L and the limiter, the amplitude of oscillation will adjust itself so that the equivalent value of the limiter resistance over a cycle, in parallel with R_L, gives the above value. Therefore,

$$R_L \geq R\frac{L'}{L}, \tag{5}$$

and a lower bound on R_L (upper bound on the loading) is determined.

The more detailed analysis which yields the circuit conditions under which (3) and (5) are valid is given in Appendix I. These circuit conditions are

$$g_m R \gg \frac{CR^2}{L'}, \tag{6}$$

$$g_m R \gg 1 + 2\frac{L}{L'}. \tag{7}$$

Note that since g_m is varied, conditions (6) and (7) must be met for the minimum g_m to be used.

To get maximum power to the load, we want R_L to be as small as possible. This implies R small and L/L' large [see (5)]. But we can only go so far with this until condition (7) is violated. If this condition does not hold, ω^2 is given by

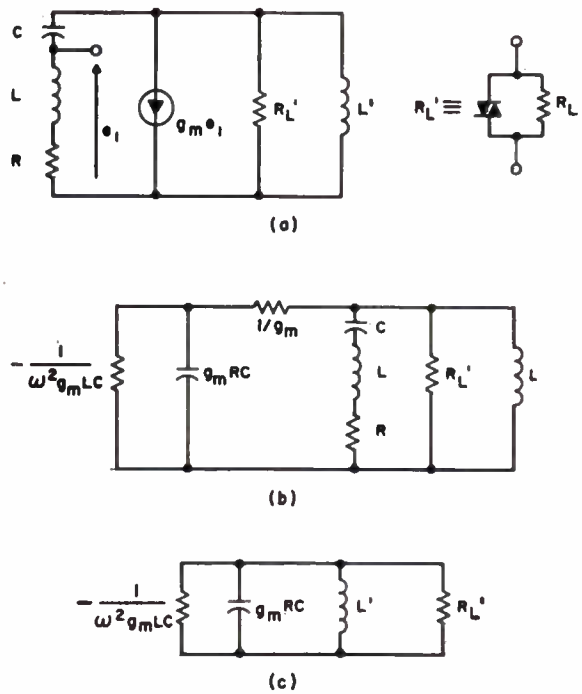


Fig. 6—Analysis of the variable reactance oscillator of Fig. 5.

$$\omega^2 = \frac{1}{g_m RCL' + (L' + 2L)C}, \tag{8}$$

as derived in Appendix I. In this case it is seen that the variable g_m term may be swamped by the constant term. Therefore, a compromise must be struck between the frequency shift obtained and the power delivered to the load. The load power can also be increased by increasing the amplitude of oscillation (controlled by the limiter). This causes more distortion since g_m will vary more over the cycle (since g_m is voltage controlled), and consequently another compromise must be made between load power and distortion.

This circuit was constructed and tested, using the component values indicated in Fig. 6 and a field effect transistor made at the Bell Telephone Laboratories, Inc. Its g_m vs control (gate) voltage characteristic is shown in Fig. 7. With component values as shown in Fig. 5, the measured frequency variation with control voltage is depicted by the dotted curve in Fig. 8; the theoretical curve, based on (8) and Fig. 7, is shown solid.[7]

From condition (5), one notes that $R_L \geq 33$ K. It was found empirically that maximum power output occurred for $R_L = 100$ K. Choosing the limiter so that the output voltage was 1 volt rms, the total harmonic distortion at any frequency was less than 2.5 per cent.

[7] The frequency range used in these experimental examples was dictated by the intended use of these oscillators, *i.e.*, for data communication over the switched telephone network.

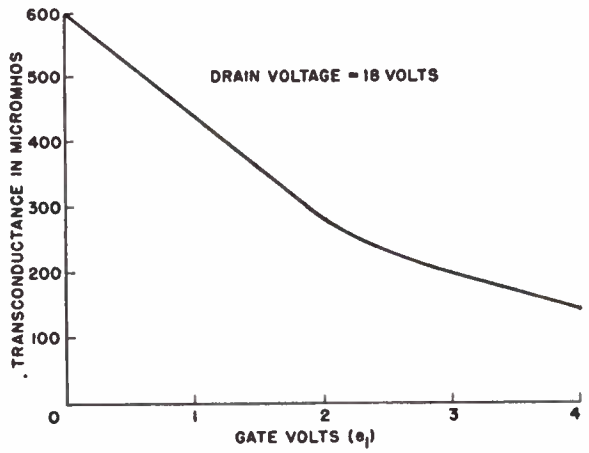


Fig. 7—Variable g_m characteristic of a typical field effect transistor.

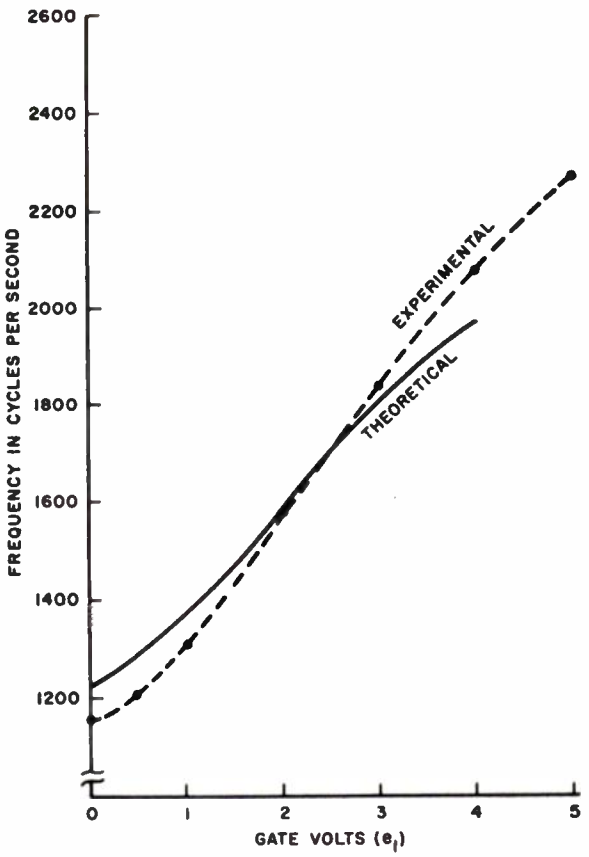


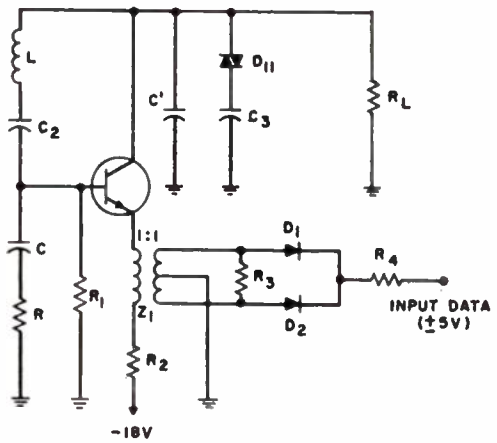
Fig. 8—Frequency characteristic of the VRO of Fig. 5.

A Transistor Frequency Shift VRO

Another example of a VRO is shown in Fig. 9. It uses a conventional transistor as a frequency shift oscillator. The g_m of a transistor stage with a resistance R_E in the emitter circuit is equal to the inverse of the emitter resistance provided that

$$R_E \gg \frac{r_b}{\beta} + r_e,$$

where r_b, β, and r_e are the appropriate transistor parameters. In the circuit of Fig. 9, a positive voltage on the input lead reverse-biases the diodes D_1 and D_2. Consequently, R_3 is reflected into the emitter circuit, causing an effective emitter resistance of R_2+R_3. When the input lead is negative, the diodes are saturated, and a small impedance is reflected into the emitter circuit. The emitter resistance is now essentially R_2. Thus, changing the polarity of the voltage on the input lead switches the emitter resistance from one value to another.



L - 10.5hy
C' - 0.0039μF
R - 9KΩ
R_1 - 2.2 MEGΩ
C - 0.01μF
R_2 - 1KΩ
R_3 - 2030Ω
$C_2 - C_3$ - 1μF
R_L - 40K

Fig. 9—A frequency shift VRO.

This transistor circuit is used as the active element in a VRO in Fig. 9. Since the switching of the effective emitter resistance will cause a pulse in the emitter current (although this is minimized by using balanced AC coupling), a VRO with a capacitive output element C' is used to minimize the voltage transient at the collector.

The ac equivalent circuit for this VRO is shown in Fig. 10; the approximate form of the equivalent circuit is shown in Fig. 11. From this figure, it is clear that

$$\omega^2 = \frac{g_m R}{LC'} \tag{8}$$

and

$$R_L \geq R\frac{C}{C'}. \tag{9}$$

The detailed analysis which yields the circuit conditions under which (8) and (9) are valid is given in Appendix II. These conditions are

$$g_m R \gg \frac{C'R^2}{L} \tag{10}$$

$$g_m R \gg 1 + 2\frac{C'}{C}. \tag{11}$$

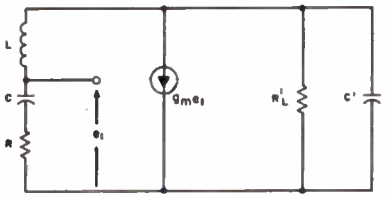


Fig. 10—An ac equivalent circuit of the VRO of Fig. 9.

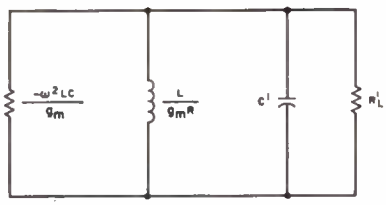


Fig. 11—Approximate equivalent circuit of the VRO of Fig. 9.

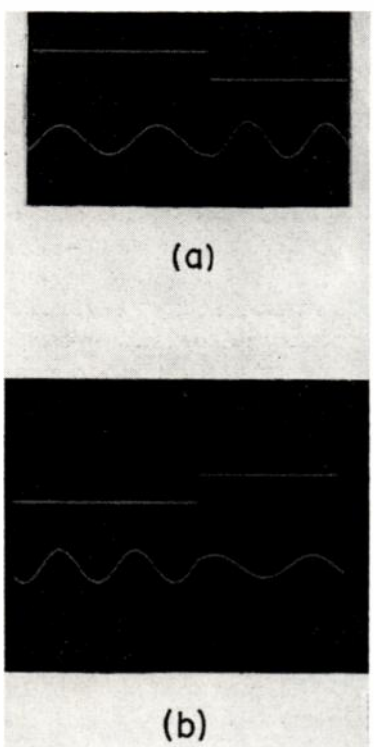


Fig. 12—Switching waveforms for the VRO of Fig. 9.

If condition (11) is not met, ω^2 is better approximated by

$$\omega^2 = \frac{g_m R + \left(1 + 2\frac{C'}{C}\right)}{LC'}. \tag{12}$$

This circuit was constructed, using the circuit values shown in Fig. 9. The two frequencies of oscillation are 1500 cps and 2000 cps. Switching waveforms are shown in Fig. 12; the upper trace is the input signal, the lower trace is the oscillator output.

Conclusion

An analysis of a class of variable frequency sinusoidal oscillators (the variable reactance oscillator) has been given. These oscillators require only one active element whose transconductance may be varied. The frequency is dependent upon this transconductance. By the use of a simple set of rules [R1–R7)], the approximate equations for the frequency and maximum loading may be determined by inspection. The circuit conditions under which these approximations are valid must be determined by a more detailed analysis.

The primary limitation of oscillators of this class is the available power output (as may be appreciated from the experimental examples). A practical limit in the audio range seems to be about 1 mw with present-day semiconductor devices if the full frequency range is to be used and distortion is to be kept to a small value.

However, such an oscillator has several advantages, among which are its simplicity and fairly wide frequency range (a 4 to 1 frequency range appears practical). It is also of interest to note that it is possible to build such an oscillator whose frequency of oscillation depends primarily on the values of two capacitors or of two inductors, rather than on the value of one capacitor and one inductor. A perusal of Fig. 4 will illustrate this point.

Appendix I

Detailed Analysis of the VRO of Fig. 5

In order to determine the circuit conditions under which the (3) and (4) hold, let us evaluate the determinant given by (1). Here

$$Z_1 = R + sL$$

$$Z_2 = \frac{1}{sC}$$

$$Z_3 = \frac{sR'_L}{s + R'_L/L'}.$$

Substituting these into (1), letting $s = j\omega$, and setting the real and imaginary parts to zero yields

$$\text{(real)}\ \frac{1}{R'_L} + \frac{CR}{L'} - \frac{\omega^2 CL}{R'_L} - \omega^2 g_m LC = 0, \tag{13}$$

$$\text{(imaginary)}\ \omega^2 L'C\left(1 + \frac{R}{R'_L} + \frac{L}{L'} + g_m R\right) = 1. \tag{14}$$

Eq. (14) can be solved for R_L', giving

$$\frac{1}{R'_L} = \frac{1}{\omega^2 L'CR} - \frac{1}{R}\left(1 + \frac{L}{L'} + g_m R\right). \tag{15}$$

This value for R'_L can be substituted into (13), and the resulting equation (a quadratic in ω^2) can be solved for ω^2. This gives

$$\omega^2 = \frac{A}{2LC(L+L')}\left[1 \pm \sqrt{1 - \frac{4L(L+L')}{A^2}}\right],$$

where

$$A = g_m RL' + L' + 2L - CR^2.$$

To simplify this into its desired form, let

$$g_m R \gg \frac{CR^2}{L'}.$$

Then A becomes

$$A = g_m RL' + (L' + 2L). \tag{16}$$

The numerator of the fraction in the radical can be written

$$4L(L+L') = (L'+2L)^2 - L'^2,$$

which is much less than A^2 if

$$g_m R \gg \frac{L'+2L}{L'} = 1 + 2\frac{L}{L'}. \tag{17}$$

If this is the case, then the fraction in the radical is much less than 1, and the approximation $\sqrt{1-x} \approx 1 - x/2$ can be used.

Doing this, ω^2 becomes simply

$$\omega^2 = \frac{1}{AC},$$

or, using (16),

$$\omega^2 = \frac{1}{g_m RCL' + (L'+2L)C}. \tag{18}$$

The negative sign associated with the radical is used here to obtain agreement with the approximate solution. Under condition (16), this becomes

$$\omega^2 = \frac{1}{g_m RCL'}.$$

R'_L can now be determined. Substituting the more accurate expression for ω^2 given by (18) into (15) yields

$$R'_L = R\frac{L'}{L},$$

exactly as precidted from the initial simple analysis. Since R'_L is the parallel combination of R_L and the limiter, the maximum loading (minimum R_L) is

$$R_L \geq R\frac{L'}{L}.$$

Appendix II

Detailed Analysis of the VRO of Fig. 9

Note that, in the VRO of Fig. 9, the following relations hold:

$$Z_1 = \frac{sRC+1}{sC}$$

$$Z_2 = sL$$

$$Z_3 = \frac{R'_L}{sC'R'_L + 1},$$

where R'_L is the parallel combination of R_L and the limiter. The real and imaginary parts of (1) are then

$$\text{(real)} \qquad g_m R'_L - \omega^2 RCC'R'_L + 1 - \omega^2 LC = 0, \tag{19}$$

$$\text{(imaginary)}$$

$$g_m RCR'_L + RC + R'_L C' - \omega^2 LC'CR'_L + R'_L C = 0. \tag{20}$$

From (20),

$$R'_L = \frac{RC}{\omega^2 LCC' - (g_m RC + C + C')}. \tag{21}$$

Substituting (21) into (19) yields a quadratic equation in ω^2, whose solution is

$$\omega^2 = \frac{A}{2LCC'}\left[1 \pm \sqrt{1 - \frac{4C'(C+C')}{A^2}}\right],$$

where

$$A = g_m RC + C + 2C' - \frac{R^2CC'}{L}.$$

This is analogous to the equation for ω^2 developed in Appendix I, and by an argument similar to that used in Appendix I reduces to

$$\omega^2 = \frac{g_m RC + C + 2C'}{LCC'} \tag{22}$$

providing that

$$g_m R \gg \frac{C'R^2}{L}$$

and

$$g_m R \gg 1 + 2\frac{C'}{C}. \tag{23}$$

Here the plus sign associated with the radical is used to obtain agreement with the approximate solution. Substituting (22) into (21) for R'_L yields

$$R'_L = \frac{-RC}{g_m RC + C' + C - \left(\dfrac{g_m RC + 2C' + C}{LCC'}\right)LCC'},$$

which reduces to

$$R'_L = R\frac{C}{C'}.$$

Since R'_L consists of the limiter and load resistance R_L in parallel, then

$$R_L \geq R\frac{C}{C'}.$$

From (22) the approximate equation for ω^2 is obtained by using the condition (23):

$$\omega^2 \approx \frac{g_m R}{LC'}.$$

ACKNOWLEDGMENT

The authors would like to thank M. A. Flavin, J. R. Davey, and J. O. Edson for their helpful discussions on this topic.

Relativistic Beam-Wave Interactions*

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Summary—**The results of the special theory of relativity have been used to modify the nonlinear interaction equations for linear-beam devices to account for relativistic mass changes when the electron velocity is a significant fraction of the velocity of light. The effect of these mass changes on gain and efficiency is evaluated and it is found that the gain is reduced and the efficiency increases rapidly as k_e approaches one. The velocity spread in the stream is independent of the voltage. The harmonic current amplitudes and optimum drift distances are calculated for a relativistic klystron and one finds that as k_e approaches one the klystron behaves according to a ballistic theory.**

INTRODUCTION

RECENT REQUIREMENTS for very high-power RF generators and amplifiers have led scientists and engineers to reexamine presently used methods of microwave power generation and envision new means to obtain the most efficient, feasible systems. Large powers require high voltages and high currents; in some cases the voltages are sufficiently high that the electron velocity of a stream is a significant fraction of the velocity of light. Under these conditions it is pertinent to inquire how the gain and efficiency of a klystron or traveling-wave amplifier, for example, are affected by relativistic mass changes of the electrons. This interesting question provides the motivation for this paper.

A consideration of the basic mode of operation of beam-wave devices immediately suggests that the gain of such a device should decrease as the particle velocity is increased into the relativistic region, since in the limit an infinite force would be required to move or bunch the beam particles. On the other hand the efficiency of a relativistic nonlinear beam-wave device would be expected to increase, since the velocity spread in the stream is not a significant fraction of the average velocity and hence it would have a less deleterious effect on bunching and the energy exchange process. Furthermore, for a given percentage decrease in velocity a relativistic beam gives up more energy than does a nonrelativistic beam.

In this paper the nonlinear analysis of linear-beam devices will be modified to conform to the requirements of the special theory of relativity and various gain and efficiency calculations will be made for the traveling-wave amplifier as well as harmonic current calculation in a relativistic klystron. Also a further qualitative understanding of the relativistic interaction will be presented in terms of a linear theory (small-signal). The totality of analyses will be one-dimensional simply because they are easier to handle and it is likely that confined flow conditions would exist in any practical embodiment.

CONSEQUENCES OF THE SPECIAL THEORY OF RELATIVITY

For the above defined problem it is worthwhile examining the consequences of the special theory of relativity with regard to the invariance of the several fundamental quantities and formulas of electrodynamics as they apply. Several significant treatises have been written on relativistic electrodynamics and its consequences [1]–[4] but unfortunately each author chooses to define a representation and transformation system of his own. The reader is cautioned to avoid the pitfalls of sign changes and different fundamental definitions of four-vector representation as used by the authors. The Lorentz transformation will be written here as a linear transformation of the components of the vector $x^i(x^1, x^2, x^3, x^4 = ct)$ in a four-dimensional space.

Implicit in the special theory is the invariance of cer-

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tain proper quantities such as the velocity of light and the combined space-time interval. The Lorentz transformation is a proper one which satisfies the above principles. A fundamental postulate of the special theory is that the conservation of momentum for particles retains its classical form and hence is Lorentz covariant (independent of the choice of inertial frame). Thus it is seen that the particle mass cannot be invariant but must depend directly on the particle velocity in the particular frame of reference.

Current density and charge density are components of a four-vector and are simply different aspects of the same thing. The electronic charge, on the other hand, is a universal constant of relativity and thus in a Lagrangian analysis, such as is used in the nonlinear beam-wave interaction theory, a counting operation on the fundamental unit of charge is also an invariant.

Other useful consequences of the special theory are that both Maxwell's equations and the scalar and vector potentials lend themselves to a covariant description. It should be clear then that the formulation of the nonlinear equations for beam-wave interactions can proceed in the laboratory system with the major modifications to be made in the force equations and the fact that both the scalar-electric and magnetic-vector potentials must be considered along with a replacement of Poisson's equation by the inhomogeneous wave equation. Since the D'Alembertian operator is an invariant, the homogeneous equation governing electromagnetic wave propagation is in covariant form.

Formulation of Lagrangian Interaction Equations

Introduction

It was pointed out above that the principal modifications to be made in the formulation of the Lagrangian interaction equations for klystrons and traveling-wave amplifiers under relativistic operating conditions are that the Lorentz equation needs to be changed to account for mass changes and that the magnetic vector potential is significant in the laboratory system. The circuit equation is in covariant form. The analysis will proceed on a one-dimensional basis and confined flow conditions will be assumed so that no appreciable radial motion is present.

Circuit Equation

The usual procedure in the formulation of a circuit equation is to make use of the similarity and correspondence between Maxwell's equations and the Telegrapher's equations and write a second-order inhomogeneous differential equation for the voltage along the line where the forcing function is the bunched RF current in the stream. Characteristic phase velocity and impedance for a particular space harmonic field are chosen to specify the coupling and degree of interaction. Even though this circuit equation was developed on a low-velocity assumption it needs no modification in the relativistic case since the relative difference in circuit wave velocity and stream velocity is small and the component of electric field parallel to particle motion is invariant to a proper transformation. On the other hand the observer in the laboratory does see something altogether different in the relativistic case as opposed to the nonrelativistic case. The same form of circuit equation will be used as was previously used [5].

Force Equation

In a multidimensional analysis there would be a term relating to $J_z \times B_\theta$ in the Lorentz equation. In the case of a one-dimensional analysis, however, this would result in a vector force in the radial direction and for the assumed confined flow conditions the term is negligible provided that $B_\theta < B_z$.

The general Lorentz transformation of the potential four-vector $(-A_x, -A_y, -A_z, V_c/c)$ between two coordinate systems where the primed system is moving at a relativistic velocity with respect to the unprimed system (laboratory system) is given by

$$\begin{bmatrix} -A_x \\ -A_y \\ -A_z \\ V_c/c \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & p_0 & -k_0 p_0 \\ 0 & 0 & -k_0 p_0 & p_0 \end{bmatrix} \begin{bmatrix} 0 \\ 0 \\ 0 \\ V_c'/c \end{bmatrix} \tag{1}$$

where

$k_0 \triangleq v_0/c$,
$v_0 \triangleq$ electromagnetic wave phase velocity relative to the laboratory system,
$c \triangleq$ velocity of light,
$p_0 \triangleq 1/\sqrt{1-k_0^2}$, and
$A \triangleq$ magnetic vector potential.

Evaluation of the above matrix equation for the scalar potential yields

$$V_c = p_0 V_c' = \frac{V_c'}{\sqrt{1-k_0^2}}. \tag{2}$$

The components of the magnetic vector potential transform as follows:

$$\begin{aligned} A_x &= 0 \\ A_y &= 0 \\ A_z &= \frac{k_0 p_0 V_c'}{c}. \end{aligned} \tag{3}$$

Combining (2) and (3) gives

$$A_z = \frac{k_0}{c} V_c. \tag{4}$$

The electric field in the one-dimensional laboratory system for relativistic systems is

$$E_z = -\left(\nabla_z V + \frac{\partial A_z}{\partial t}\right). \tag{5}$$

Fortunately the time derivative of the vector potential in (5) may be handled in another manner since the field is carried by the moving charge. An observer in the laboratory system will see the same change in any field quantity during the time dt as will an observer who moves a distance $-vdt$ from the same position without any lapse of time. The magnetic vector potential is taken to be zero in the primed system, or in the laboratory system if v/c is sufficiently small. It is easy to show that the component of electric field parallel to the particle motion is invariant to a Lorentz transformation and hence the time derivative of A_z may be treated in a simple manner. Thus, for any field quantity

$$\frac{\partial}{\partial t} + v_0 \cdot \nabla = 0. \tag{6}$$

Hence (5) with the aid of (4) becomes

$$E_z = -\nabla_z V_c + v_0 \cdot \nabla_z A_{cz}$$

$$= -(1 - k_0^2)\frac{\partial V_c}{\partial z}. \tag{7}$$

Eq. (7) has been written for the circuit electric field. Of course, the space-charge component of electric field may be handled in exactly the same way.

The Lagrangian as a function of time for a charged particle in an electromagnetic field is given by

$$L(t) = \left[-mc^2\sqrt{1 - k_e^2} - eV + \frac{e}{c}v \cdot A\right], \tag{8}$$

where

$k_e \triangleq u_0/c$, and
u_0 = electron velocity relative to the laboratory system.

The generalized momentum is given by

$$p = \frac{\partial L(t)}{\partial v} \tag{9}$$

and the action integral

$$I = \int_{t_1}^{t_2} L(t)dt \tag{10}$$

is a scalar invariant. Applying Hamilton's variational principle to the action integral such that

$$\delta I = \delta \int_{t_1}^{t_2} L(t)dt = 0 \tag{11}$$

leads to the Euler-Lagrangian equations

$$\frac{d}{dt}\left(\frac{m_0 u^s}{\sqrt{1 - k_e^2}}\right) = eE^s + \frac{e}{c}v^r H^{sr}, \tag{12}$$

where

H^{sr} = the electromagnetic field tensor and the term $(e/c)v^r H^{sr}$ is zero for a one-dimensional confined flow, and
m_0 = the rest mass of the electron.

Thus

$$f = \frac{dp}{dt} = m\frac{dv}{dt} + v\frac{dm}{dt} \tag{13}$$

and the time rate of change of mass is

$$\frac{dm}{dt} = \frac{m_0 v \dfrac{dv}{dt}}{c^2(1 - k_e^2)^{3/2}}. \tag{14}$$

Combining (13) and (14) yields

$$\frac{dv}{dt} = -\eta_0(1 - k_e^2)^{3/2}(E_c + E_{sc}) \tag{15}$$

where

$\eta_0 \triangleq e/m_0$, $e > 0$, the charge-to-mass ratio.

In order to relate k_0 of (7) and k_e of (15), it will be assumed that the electromagnetic wave on the circuit travels at a velocity v_0 and thus since $u_0/v_0 = 1 + Cb$

$$k_0^2 = \frac{k_e^2}{(1 + Cb)^2}, \tag{16}$$

where

$C \triangleq$ gain parameter of Pierce [6], and
$b \triangleq (u_0 - v_0)/Cv_0$, the relative velocity parameter of Pierce.

Thus (7) becomes

$$E_{cz} = -\left[1 - \frac{k_e^2}{(1 + Cb)^2}\right]\frac{\partial V_c}{\partial z}. \tag{17}$$

The space-charge component of potential is now given by the inhomogeneous wave equation as

$$\Box^2 V_{sc} = -\rho/\epsilon. \tag{18}$$

Since the D'Alembertian is invariant to a proper transformation and the space-charge potential of (15) is measured in the laboratory system, the derivation of the space-charge field expression proceeds as in the nonrelativistic case. The complete derivation of the space-charge field expression is given in the Appendix. The basic result is that the expression for E_{sc} remains in the same form except that the arguments of the Bessel functions become modified as follows:

$$I_1(n\beta b') = I_1(n\beta b'\sqrt{1 - k_e^2}) = I_1(nB_r)$$

and the plasma frequency reduction factor is given by

$$R_n^2 = 1 - \frac{n\sqrt{1 - k_e^2}\cdot\beta b'}{I_{0na'}}(I_{1nb'}K_{0na'} + I_{0na'}K_{1nb'}) \tag{19}$$

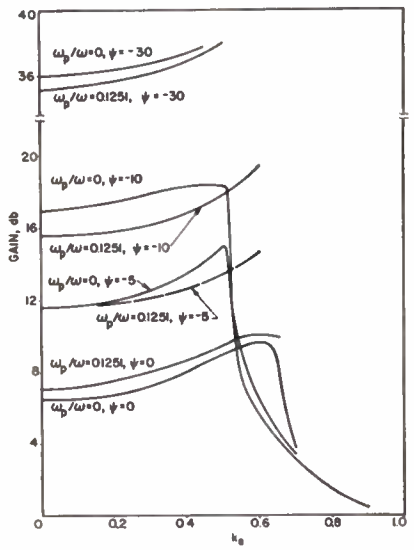


Fig. 1—Saturation gain vs k_e for a relativistic TWA. ($C=0.1$, $B=1.0$, $d=0$, $b=1.75$.)

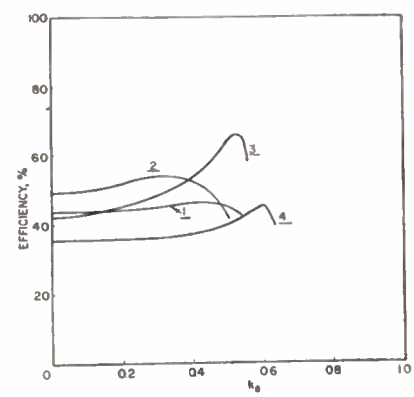


Fig. 2—Efficiency vs k_e for a relativistic TWA as a function of RF drive, ω_p/ω and b. (1. $\omega_p/\omega=0$, $\psi=-30$, $b=1.75$. 2. $\omega_p/\omega=0$, $\psi=-10$, $b=1.75$. 3. $\omega_p/\omega=0$, $\psi=-5$, $b=1.75$, 4. $\omega_p/\omega=0.1251$, $\psi=-10$, $b=1.75$.)

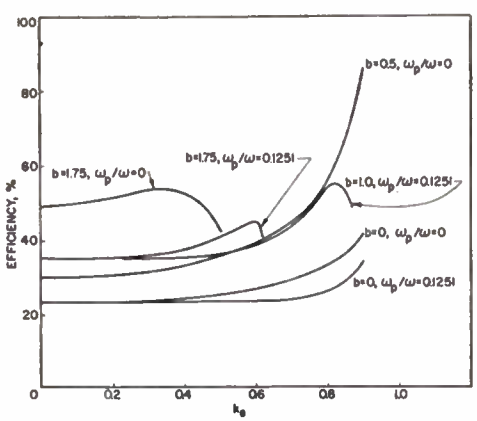


Fig. 3—Efficiency vs k_e for a relativistic TWA. b and ω_p/ω are the parameters. ($C=0.1$, $B=1.0$, $d=0$, $\psi=-10$.)

where the above quantities are defined in Rowe [5], the Appendix and $B_r \triangleq B\sqrt{1-k_e^2}$. The space-charge fields must also be modified by the inclusion of the magnetic vector potential when transforming from the primed system to the laboratory system. The result is

$$E_{sc-s} = -(1-k_e^2)\frac{\partial V_{sc}}{\partial z}. \tag{20}$$

The final result is that the new force equation for a relativistic one-dimensional electron beam in confined flow is given by

$$\begin{aligned}\frac{\partial u(y,\Phi_0)}{\partial y}&[1+2Cu(y,\Phi_0)]\\ &=(1-k_e^2)^{3/2}\left\{\left[1-\frac{k_e^2}{(1+Cb)^2}\right]\right.\\ &\cdot\left[A(y)\left(1-C\frac{d\theta(y)}{dy}\right)\sin\Phi(y,\Phi_0)\right.\\ &\left.-C\frac{dA(y)}{dy}\cos\Phi(y,\Phi_0)\right]\\ &+\frac{1}{(1+Cb)}\left(\frac{\omega_p}{\omega C}\right)^2\left.\int_0^{2\pi}\frac{F(\Phi-\Phi')d\Phi_0}{1+2Cu(y,\Phi_0')}\right\}.\end{aligned} \tag{21}$$

Relations Between Variables

It is recalled that the dependent variables in the nonlinear Lagrangian analysis were measured relative to the electron velocity; *i.e.*, the total electron velocity and phase positions are measured with respect to either the electromagnetic wave or the average stream velocity u_0 as was the wave phase shift. Hence there is no relativistic correction to be made in the equation relating them. Therefore, as before,

$$\frac{\partial\Phi(y,\Phi_0)}{\partial y}+\frac{d\theta(y)}{dy}=\frac{2u(y,\Phi_0)}{1+2Cu(y,\Phi_0)}. \tag{22}$$

Nonlinear Results

The relativistic interaction equations of the previous section may be used to calculate gain and efficiency as a function of various beam-structure characteristics. The calculated results presented here serve to illustrate the important effects on gain and efficiency in a collinear beam-wave relativistic interaction when the beam is traveling at relativistic velocities. Under low-stream-velocity conditions the optimum nonlinear performance is obtained when relative injection velocities of $b=1.75$ or so are used, *i.e.*, when the stream is initially moving some 17.5 per cent faster than the wave for $C=0.1$.

Typical saturation gain calculations for a particular RF power input level are shown in Fig. 1 as a function of u_0/c, with efficiency data given in Fig. 2. The RF power input assumed is fairly high, representing one per cent of the beam power. It is seen that the satura-

tion gain increases for increasing stream velocities but that the gain cuts off around $k_e = 0.6$. That this happens is due to a lack of synchronism between the stream and the wave; it would not have happened if the stream were initially synchronous with the RF wave, as is illustrated in Fig. 3.

The effect of finite space-charge forces ($\omega_p/\omega \neq 0$) is to reduce the gain and efficiency as expected if B is kept constant independent of k_e. Since the space-charge field derivation shows that $B_r = \sqrt{1-k_e^2} \cdot B$, the effective B is reduced at relativistic velocities and the efficiency increases, approaching the $\omega_p/\omega = 0$ results at large k_e.

At large values of k_e the optimum injection condition on the stream is that it be made synchronous with the wave. A proof of this will be given in the next section.

As would be expected, the effect of higher-circuit-impedance and higher-perveance streams would be to increase the efficiency at any given k_e value. Typical results for a variety of conditions are shown in Figs. 4 and 5 when B is considered constant. Replacing B by B_r yields finite space-charge results essentially equal to the zero space-charge results since the longitudinal space-charge debunching is nonexistent at high velocities. Unfortunately at relativistic velocities the needed interaction length becomes excessive for low current and low impedance. At high perveance or circuit impedance, however, the length is reasonable and practical for high-efficiency power amplification. Gain vs distance for representative cases is shown in Figs. 6 and 7 as a function of k_e. Clearly there is a great deal of advantage of working at 1-mev beam energy levels, as the efficiency increases 2 or 3 times over the low-velocity value. High-energy beams would permit the use of physically larger circuits at all frequencies which affords particular advantages to the millimeter and sub-millimeter wavelength regions. A simple prediction of the efficiency of a relativistic tube, based on assuming that the velocity distribution in the spent beam is independent of voltage level, would be an increase of $p_e/2(p_e+1)$ where $p_e \triangleq 1/\sqrt{1-k_e^2}$. Of course this is not valid near a k_e of 0.8 to 1 since it would give greater than 100 per cent efficiency; it is however useful. If this factor is adjusted to 25 per cent efficiency at $k_e = 0$ the dotted curve in Fig. 4 is obtained. The general qualitative agreement is evident.

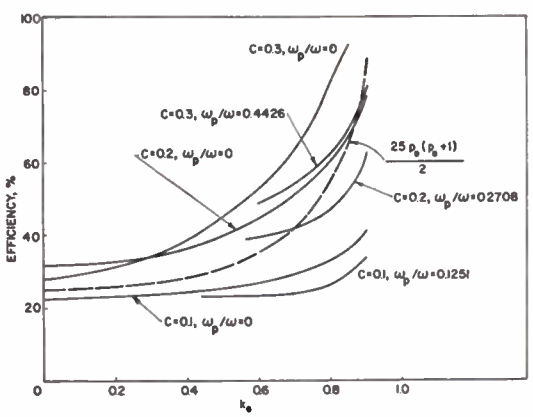


Fig. 4—Efficiency vs k_e for a relativistic TWA with C and normalized plasma frequency as the parameters. ($d=0$, $b=0$, $\psi=-10$, $B=1.0$ (constant).)

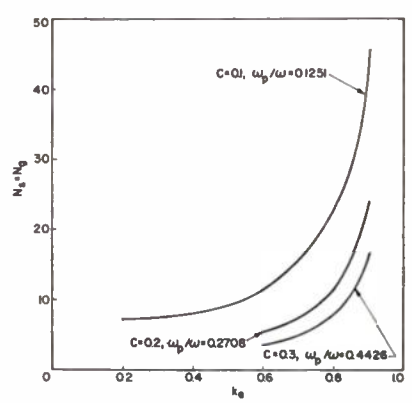


Fig. 5—Length vs k_e for a relativistic TWA. ($d=0$, $b=0$, $\psi=-10$, $B=1.0$ (constant).)

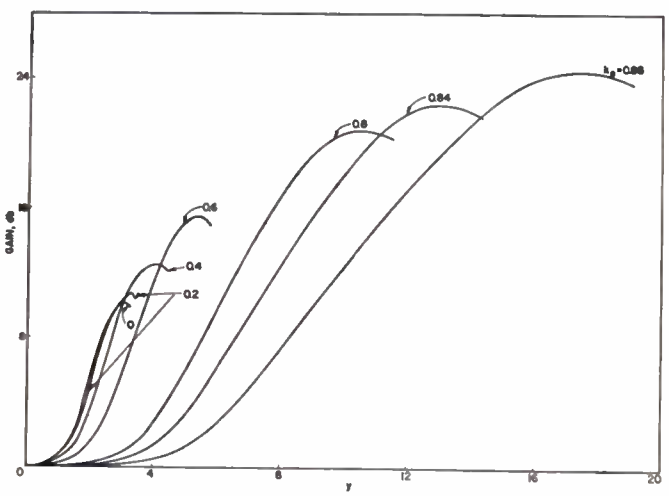


Fig. 6—Gain vs distance for a relativistic TWA. ($C=0.3$, $\omega_p/\omega=0$, $d=0$, $b=0$, $\psi=-10$ db, $B=1.0$.)

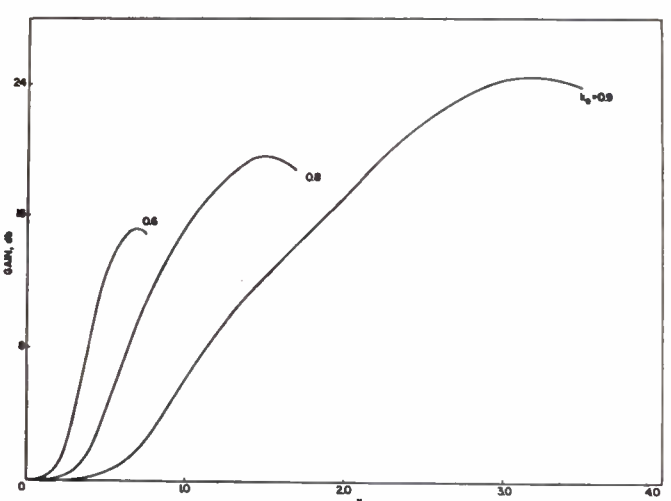


Fig. 7—Gain vs distance for a relativistic TWA. ($C=0.3$, $\omega_p/\omega=0.4426$, $d=0$, $b=0$, $\psi=-10$ db, $B=1.0$).

The calculation of efficiency for these relativistic interactions proceeds directly from the equivalence of mass and energy given by the Einstein relation $\Delta_r = \Delta mc^2$. The decrease in potential energy or increase in kinetic energy of a stream when accelerated through a large potential difference V_r is given by

$$eV_r = m_0c^2(p_e - 1) \tag{23a}$$

and under the low-velocity assumption

$$eV_0 = \frac{m_0}{2} u_0^2. \tag{23b}$$

Thus

$$\frac{V_r}{V_0} = \frac{2p_e^2}{p_e + 1}. \tag{24}$$

In making the relativistic calculations $C^3 = Z_0I_0/4V_0$ was assumed constant and hence there must necessarily be an impedance change as V_r changes in order to maintain C constant. The impedance ratio is given by

$$\frac{Z_{0r}}{Z_0} = \frac{2p_e^2}{p_r + 1}. \tag{25}$$

Linear Relativistic Interaction Equations

The modification of Pierce's [6] linear theory for the traveling-wave amplifier to account for relativistic effects can proceed along the same lines as developed for the nonlinear equations. The circuit equation does not change in the relativistic case even though it was derived on the assumption that $v/c \ll 1$ so that electrostatics were valid.

The continuity equation also does not need modification since the current density is a three-vector component of a general four-vector, the charge density being the fourth component, and is constant in the laboratory system.

The force equation is again modified to include the time-changing electron mass and the magnetic vector potential as in the nonlinear case. Hence the same form of this equation as given in (12) is used.

It will be assumed throughout the linear analysis that all RF quantities vary as $\exp j\omega t - \Gamma z$ with respect to the laboratory system, where the propagation constant Γ is as defined by Pierce. The resulting determinantal equation for a linear one-dimensional relativistic beam-wave interaction is given by

$$\delta^2 = (1 - k_e^2)^{3/2}\left\{\frac{\left[1 - \frac{k_e^2}{(1 + Cb)^2}\right](1 + jC\delta)^2[1 + C(b - jd)]}{\left[-b + jd + j\delta + C\left(jdb - \frac{b^2}{2} + \frac{d^2}{2} - \frac{\delta^2}{2}\right)\right]} - 4QC(1 + jC\delta)^2\right\}, \tag{28}$$

The power output in general is given by

$$P = \left.\frac{V_{rf}^2}{Z_{0r}}\right|_{avg}$$

$$= 2CI_0V_0(A^2_{max} - A_0^2)\frac{\left(1 - C\frac{d\theta(y)}{dy}\right)}{(1 + Cb)}\frac{(p_e + 1)}{2p_e^2}, \tag{26}$$

where the method outlined in Rowe and Hok [7] was followed and

$$A_{max} = A(y)|_{max}$$
$$A_0 = A(0).$$

The saturation efficiency is then

$$\text{Eff} = 2C(A^2_{max} - A_0^2)\frac{\left(1 - C\frac{d\theta(y)}{dy}\right)}{(1 + Cb)}\frac{(p_e+1)^2}{4p_e^4}, \tag{27}$$

since the input power is $I_0V_0[2p_e^2/(p_e+1)]$. Calculations on the spent beam indicate that the velocity spread in the beam is independent of the voltage.

where the symbols used are according to those defined by Pierce.

Linear Results

In a previous section on nonlinear calculations it was pointed out that the small-signal gain decreased as k_e was increased. A typical set of results obtained by solving (28) is shown in Fig. 8, where the real parts of the growing and declining wave propagation constants are shown. As the injected stream velocity is made higher the low-level gain decreases and the relative injection velocity required tends toward synchronism with the characteristic circuit phase velocity. A composite of this information is shown in Figs. 9 and 10, where the maximum of Re (δ_1) and the value of b at which it occurs are plotted vs k_e for a range of C.

A simple method of predicting the effect of relativistic mass changes on gain [8] is to assume that the effective C has been reduced and the gain reduced proportionately. The dependence of C on k_e is as follows

$$\frac{C_R}{C_0} = \left[\frac{2}{p_e(p_e + 1)}\right]^{1/3}.$$

This is plotted in Fig. 9 along with the product of C_R/C_0 and the value of x_1 at $C=0.1$ and $k_e=0$. The agreement is fair at low velocities but deviates markedly as k_e approaches one. This reduction in a small-signal gain has been predicted with a coupled-mode theory.

Klystron Interaction

A much investigated and well understood interaction occurs when an electron beam in a klystron is modulated at a buncher cavity and then passed through a drift tube and subsequently through a catcher cavity. This two-cavity klystron may be analyzed by dropping the circuit field terms from the Lorentz force equation and modifying the phase equation. Thus only the modified forms of (21) and (22) are required to study this interaction under relativistic conditions. It is convenient to relate the bunching parameter X for klystrons to the normalized distance parameter for traveling-wave tubes:

$$X = \frac{\alpha\theta_0}{2} = \pi\alpha N_s \tag{29}$$

$$y = \frac{C\omega z}{u_0} = 2\pi C N_s. \tag{30}$$

These two normalized distances are equivalent provided that $\alpha \triangleq 2C$. Introduction of the above into (21) and (22) gives the following equations for the relativistic klystron after the appropriate circuit field terms are eliminated.

$$\frac{\partial\Phi(X, \Phi_0)}{\partial X} = \frac{2u(X, \Phi_0)}{1 + \alpha u(X, \Phi_0)}, \tag{31}$$

and

$$\frac{\partial u(X, \Phi_0)}{\partial X}\left[1 + \alpha u(X, \Phi_0)\right] = 4(1 - k_e^2)^{3/2}\left(\frac{\omega_p}{\omega\alpha}\right)^2 \int_0^{2\pi} \frac{F(\Phi - \Phi')d\Phi_0'}{1 + \alpha u(X, \Phi_0)}. \tag{32}$$

The relativistic klystron equations may be solved for assumed values of the space-charge integral coefficient $\omega_p/\omega\alpha$ and the initial velocity modulation. For the calculations presented here it is assumed that the stream is velocity-modulated at a zero transit angle gap. The electron phase positions (relative to their entrance phase) and velocities may be plotted at any constant-distance plane and from this information the harmonic structure of the beam current may be determined. The amplitude of the nth harmonic of beam current to the dc is given by

$$\frac{|i_n|}{I_0} = \frac{(A_n^2 + B_n^2)^{1/2}}{\frac{1}{2\pi}\int_0^{2\pi}[1 + \alpha u(0, \Phi_0')]d\Phi_0'}, \tag{33}$$

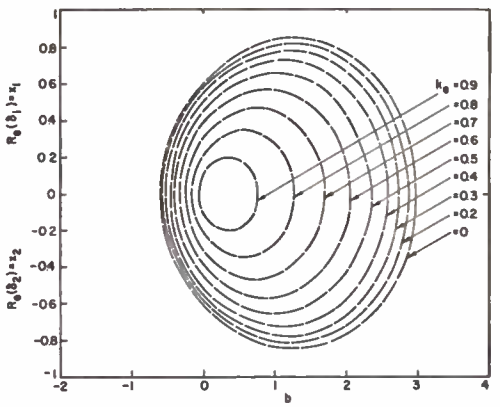


Fig. 8—Small-signal gain vs b for a relativistic TWA. ($C=0.1$, $QC=0.25$, $d=0$.)

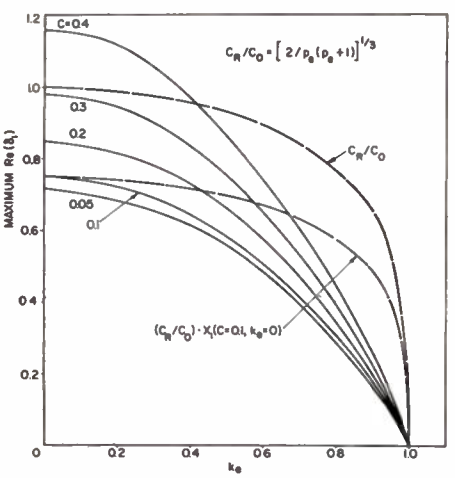


Fig. 9—Maximum Re (δ_1) vs k_e. ($QC=0.25$, $d=0$.)

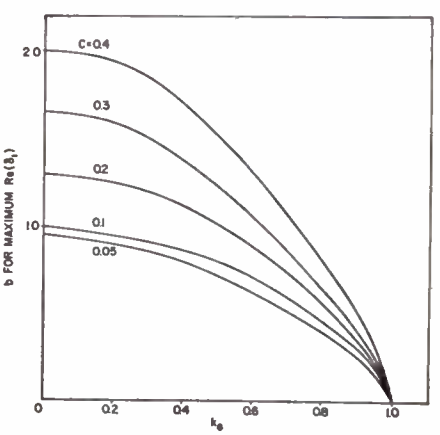


Fig. 10—b for maximum Re (δ_1) vs k_e. ($QC=0.25$, $d=0$.)

where

$$A_n = \frac{1}{\pi}\int_0^{2\pi} [1 + \alpha u(0, \Phi_0')] \cos \Phi(X, \Phi_0') d\Phi_0'$$

and

$$B_n = \frac{1}{\pi}\int_0^{2\pi} [1 + \alpha u(0, \Phi_0')] \sin \Phi(X, \Phi_0') d\Phi_0',$$

where $u(0, \Phi_0')$ represents the initial normalized RF velocity modulation of the individual electrons.

Some representative calculations of the beam RF current structure are illustrated in Fig. 11, where the harmonic-to-dc-current ratio is plotted vs n for a specific value of $\omega_p/\omega\alpha$. As a basis of comparison the original space-chargeless theory of Webster may be used to calculate the harmonic content of the stream. Webster's procedure was to use a ballistic model and to neglect space-charge forces. If one assumes an ideal velocity modulation gap then the harmonic-to-dc-current ratio in the stream is given by

$$\frac{|i_n|}{I_0} = 2J_n(nX)\big|_{\max}, \tag{34}$$

where $J_n(nX)\big|_{\max}$ is the maximum of the J_n Bessel function; this of course also gives information on the distance from the modulating gap where the nth harmonic current reaches a maximum.

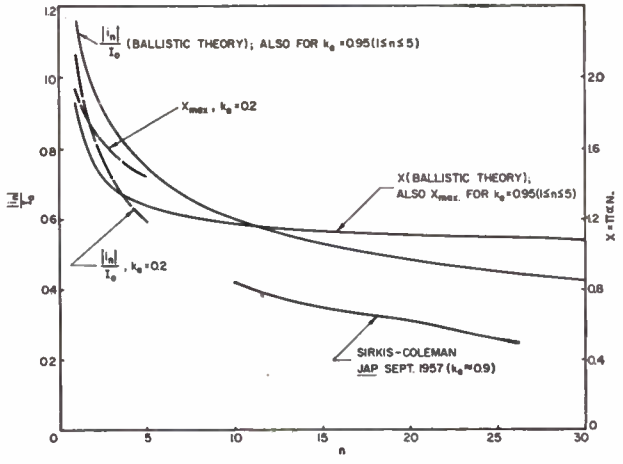


Fig. 11—Klystron harmonic current amplitudes and optimum drift distances. $(\omega_p/\omega\alpha)=0.63$.

The maximum current ratio and the value of X at which it occurs are also given in Fig. 11. It is seen that for low values of k_e the nonlinear relativistic current ratio is less than that for the ballistic calculations. However, independent of $\omega_p/\omega\alpha$, for high values of k_e the harmonic-to-dc-current ratios are just those given by the ballistic theory. The comparison of the optimum drift distances indicates the same thing. This is to be expected from a consideration of (44) since at high relativistic velocities the right-hand side approaches zero independent of $\omega_p/\omega\alpha$ and hence space-charge forces are important. As has been shown by Webber [9], it is probably possible to find combinations of parameters which will give current ratios greater than those predicted by ballistic theory. However, even in these cases very high energy beams would behave ballistically. Even larger harmonic-to-dc-current ratios are achievable through the use of adequately prebunched electron beams. Again it is apparent that high-energy beams would have great utility in millimeter wave generators and amplifiers.

In order to give some comparison with actual operating relativistic beams, some of the results of Sirkis and Coleman are also shown in Fig. 11, where the computed harmonic-to-dc-current amplitudes are given for a rebatron-harmodotron system where k_e is approximately 0.9. These experimental data do not represent the best attainable. The current amplitudes were calculated by Fourier analysis at a distance of 10 cm from the exit aperture of the accelerating cavity. These are not maximum obtainable current ratios and are not meant to correlate directly with the calculations presented here. As in the case of the TWA the velocity spread in the klystron beam is independent of the voltage.

Conclusions

The one-dimensional interaction theory developed above for relativistic beams and waves assumes that the θ-generated magnetic field due to the longitudinal current is small relative to the axial confining field. Hence essentially confined flow conditions must prevail. The high interaction efficiency attainable at large values of k_e suggests that such interaction would be extremely useful in very short wavelength generation and amplification. For $B_z > B_\theta$ two-dimensional field effects probably do not significantly modify the predicted efficiencies providing that the pinch effect does not lead to instabilities.

Appendix

Relativistic Space-Charge Field Expression

The derivation of a useful space-charge field expression for a relativistic one-dimensional confined-flow electron beam will proceed along the lines followed in the corresponding nonrelativistic theory. The principal assumptions to be made are that the RF wave bunches the stream in such a way that the space-charge density is constant in magnitude and varies sinusoidally with distance and that there is positive ion neutralization of the average space charge. The equations of electrostatics are no longer sufficient when $v \simeq c$ and Poisson's equation is generalized to its four-dimensional equivalent, the inhomogeneous wave equation. Thus, it is assumed that

$$\Box^2 V_{sc} = \frac{-\rho(z)}{\epsilon\pi b'^2} = -A\rho e^{j(\omega t - \beta z)}, \tag{35}$$

where b' represents the electron stream radius.

Eq. (35) applies inside the stream whereas between the stream and the surrounding drift tube the source term is zero; hence

$$\square^2 V = 0. \tag{36}$$

The solution of (35) is determined as

$$V(r)e^{j(\omega t-\beta z)} = \left[\frac{A\rho}{\beta^2(1-k_e^2)} + BI_0(\beta r\sqrt{1-k_e^2})\right]e^{j(\omega t-\beta z)}, \tag{37}$$

where $k_e = u_0/c$. Eq. (37) may be verified directly by substituting directly into (35). The magnetic vector potential must be included as follows:

$$E_{zc} = -\left(\nabla V_{zc} + \frac{\partial A}{\partial t}\right), \tag{38}$$

and it was shown in (20) that the total electric field in the direction of the particle motion may be written as

$$E_{zc-z} = -(1-k_e^2)\frac{\partial V_{zc}}{\partial z}. \tag{20}$$

Thus the space-charge field may be obtained from (37) as

$$E_{zc-z} = j\beta(1-k_e^2)\left[\frac{A\rho}{\beta^2(1-k_e^2)} + BI_0(\beta r\sqrt{1-k_e^2})\right] \cdot e^{j(\omega t-\beta z)}. \tag{39}$$

In the charge-free region between the stream and the drift tube E_{zc-z} has a Bessel function dependence. The continuity of radial and longitudinal components of electric field at the stream boundary and the requirement of zero longitudinal field at the drift tube wall define an eigenvalue problem whose solution gives the space-charge wave propagation constants (eigenvalues of the system). The resultant expression for E_{zc-z} is

$$E_{zc-z} = \frac{j\rho A}{\beta}\left\{1 - \beta b'\sqrt{1-k_e^2}\,\frac{I_0(\beta r\sqrt{1-k_e^2})}{I_0(\beta a'\sqrt{1-k_e^2})} \cdot \left[I_1(\beta b'\sqrt{1-k_e^2})K_0(\beta b'\sqrt{1-k_e^2}) + I_0(\beta a'\sqrt{1-k_e^2})K_1(\beta b'\sqrt{1-k_e^2})\right]\right\} e^{j(\omega t-\beta z)}. \tag{40}$$

The space-charge field is evaluated at $r=0$ and the time dependence is suppressed.

$$E_{zc-z} = \frac{j\rho A}{\beta} R^2 e^{-j\beta z}, \tag{41}$$

where

$$R^2 \triangleq 1 - \frac{B_r}{I_{0a'}}(I_{1b'}K_{0a'} + I_{0a'}K_{1b'}),$$

$$I_{1b'} \triangleq I_1(\beta b'\sqrt{1-k_e^2}), \text{ and}$$

$$B_r \triangleq \beta b'\sqrt{1-k_e^2} = B\sqrt{1-k_e^2}.$$

The I and K functions are the modified Bessel functions, first and second kind.

The derivation of the final form for E_{zc} proceeds directly along the lines used in Rowe [5]. The electron plasma frequency may be defined in terms of either the laboratory system particle mass or the relativistic mass, *i.e.*,

$$\omega_p^2 \triangleq \frac{|I_0|\,|\eta_0|}{\pi\epsilon b'^2 u_0}. \tag{42}$$

The resultant relativistic expression for E_{zc-z} is

$$E_{zc-z} = \frac{2\omega u_0}{|\eta_0|(1+Cb)}\left(\frac{\omega_p}{\omega}\right)^2 \int_0^{2\pi} \frac{F(\Phi-\Phi')d\Phi_0'}{1+2Cu(y,\Phi_0')}, \tag{43}$$

where the weighting function $F(\Phi-\Phi')$ may be written in closed form as

$$F(\Phi-\Phi') = \frac{1}{2\pi}\left[\frac{\pi-(\Phi-\Phi')}{2} - 2\tan^{-1}\frac{\sin(\Phi-\Phi')}{e^{B_r f(a'/b')} - \cos(\Phi-\Phi')} + \tan^{-1}\frac{\sin(\Phi-\Phi')}{e^{2B_r f(a'/b')} - \cos(\Phi-\Phi')}\right]. \tag{44}$$

Variables and parameters introduced above and in the body of the paper are defined below:

$\Phi(y, \Phi_0) \triangleq$ electron phase relative to the RF wave as a function of position and initial phase,

$u(y, \Phi_0) \triangleq$ normalized electron velocity defined by $u_t(y, \Phi_0) = u_0[1+2Cu(y, \Phi_0)]$,

$A(y) \triangleq$ normalized RF voltage along the circuit,

$\theta(y) \triangleq$ phase lag of the RF circuit wave, and

$a' \triangleq$ RF circuit radius.

References

[1] C. Moller, "The Theory of Relativity," Oxford University Press, London, England, 1955.

[2] P. G. Bergmann, "An Introduction to the Theory of Relativity," Prentice Hall, Inc., New York, N. Y.; 1942.

[3] A. Einstein, H. A. Lorentz, H. Minkowski, and H. Weyl, "The Principle of Relativity," Dover Publications, Inc., New York, N. Y.; 1923.

[4] W. Panofsky and M. Phillips, "Classical Electricity and Magnetism," Addison Wesley Publications, Inc., Reading, Mass., Chs. 14–19; 1955.

[5] J. E. Rowe, "A large-signal analysis of the traveling-wave amplifier: theory and general results," IRE Trans. on Electron Devices, vol. ED-3, pp. 39–57; January, 1956.

[6] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

[7] J. E. Rowe and G. Hok, "When is a backward-wave not a backward wave?" Proc. IRE, vol. 44, pp. 1060–1061; August, 1956.

[8] W. R. Ayers, "High-Power Applications of the Connected-Ring Structure in Traveling-Wave Tubes," Stanford University, Stanford, Calif., Scientific Rept. No. 9; December, 1958.

[9] S. E. Webber, "Ballistic analysis of a two-cavity finite beam klystron," IRE Trans. on Electron Devices, vol. ED-5, pp. 98–109; April, 1958.

Truncation Error of Sampling-Theorem Expansions*

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Summary—Upper bounds are obtained for the error, termed *truncation error*, which arises in reconstituting a band-limited function by summing over only a finite number (instead of the requisite infinite number) of samples of this function in an appropriate sampling-theorem expansion. Upper bounds are given for the truncation errors of the Cardinal and Fogel sampling expansions and for "self-truncating" versions of these two sampling expansions; these latter sampling expansions are "self-truncating" in the sense that the upper bounds on their truncation errors are almost always much lower than the upper bounds on the truncation errors of their prototype sampling expansions.

All of the upper bounds are given as functions of three parameters: M, the maximum magnitude of the band-limited function; q, the per unit guard band (assuming that the band-limited function is sampled at a rate greater than the Nyquist rate); and N, the measure of the number of samples in the finite summation of terms of the sampling expansion.

Introduction

THE WELL-KNOWN *cardinal series*[1] sampling expansion for the function $h(t)$ may be written as

$$h(t) = \sum_{n=-\infty}^{\infty} h\left(\frac{n}{2W}\right) \operatorname{sinc} 2W\left(t - \frac{n}{2W}\right), \tag{1}$$

where the form sinc x is defined by

$$\operatorname{sinc} x \triangleq \frac{\sin \pi x}{\pi x}. \tag{2}$$

It has been shown[2] that the function $h(t)$, whose samples are employed in (1), is uniquely determined by this equation provided $h(t)$ is band limited in the sense that it contains no frequencies greater than W cps.

The expansion of (1) permits a continuous band-limited signal to be represented by a discrete set of samples of the signal. The original signal $h(t)$ can be recovered through (1), which allows any value of the signal to be interpolated exactly between the samples $h(n/2W)$. Theoretically, a delayed version of $h(t)$ can be "realized" by passing an infinite train of impulses proportional to the $h(n/2W)$, $n = \cdots, -1, 0, 1, \cdots$, through an ideal low-pass filter. This filter introduces only a linear phase shift and has an amplitude response equal to $1/2W$ for all frequencies f for which $f \leq W$ cps and equal to zero for $f > W$ cps.

In practical situations, the right side of (1) usually differs from $h(t)$. The difference is called *error* in this paper and may arise from several sources. For example, *aliasing error*[3] occurs when the original function $h(t)$ is not precisely band limited to W cps. Thus Stewart[4] and Spilker[5] have derived average values of the mean-squared error which occurs when a nonband-limited function is reconstituted from its samples.

Another type of error, which may be called *truncation error*, arises when only a finite number of terms are used in the series of (1). This latter type of error can be considered to arise in any physical operation involving the mechanization of (1). In the first place, only a finite number of samples can be summed. Secondly, any attempts to realize (1) by low-pass filters involves a truncation of the impulse responses of the filters. In this paper, an upper bound is obtained for the truncation error which occurs when a low-pass band-limited function is reconstituted at any instant of time.

Truncation Error for the Cardinal Series

Let $h(t)$ be approximated by the expansion

$$h_2(t) = \sum_{n=K-N}^{K+N} h\left(\frac{n}{2W}\right) \operatorname{sinc} 2W\left(t - \frac{n}{2W}\right), \quad 0 < N < \infty. \tag{3}$$

Here K is an integer which is assumed to be a function of t [see (4)] and N is a fixed integer. The difference $h(t) - h_2(t)$ is the truncation error, which approaches zero in the limit as N approaches infinity provided $h(t)$ is bandlimited to W cps.

Actually, for this treatment, it will be necessary to restrict the function $h(t)$ so that it contains no frequency components greater than rW cps, where $0 < r < 1$. The *guard band*, $(1-r)W$ cps, thus provided may be made arbitrarily small, but the truncation-error bound will increase correspondingly.

Let the integer K of (3) be chosen according to the inequalities

$$2Wt - \tfrac{1}{2} \leq K(t) \leq 2Wt + \tfrac{1}{2}. \tag{4}$$

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[1] E. T. Whittaker, "On the functions which are represented by the expansions of the interpolation theory," *Proc. Roy. Soc. (Edinburgh)*, vol. 35, pp. 181–194; 1914–1915.

[2] C. E. Shannon, "Communication in the presence of noise," PROC. IRE, vol. 37, pp. 10–21; January, 1949.

[3] R. B. Blackman and J. W. Tukey, "The measurement of power spectra from the point of view of communications engineering," *Bell Sys. Tech. J.*, vol. 37, p. 217; January, 1958.

[4] R. M. Stewart, "Statistical design and evaluation of filters for the restoration of sampled data," PROC. IRE, vol. 44, pp. 253–257; February, 1956.

[5] J. J. Spilker, Jr., "Theoretical bounds on the performance of sampled data communications systems," IRE TRANS. ON CIRCUIT THEORY, vol. CT-7, pp. 335–341; September, 1960.

The truncation error $E(t)$ associated with (3) is by definition

$$E(t) \triangleq h(t) - \sum_{n=K-N}^{K+N} h\left(\frac{n}{2W}\right) \operatorname{sinc} 2W\left(t - \frac{n}{2W}\right). \quad (5)$$

Now let us assume explicitly that the function $h(t)$ has no frequency components greater than rW cps, where $0<r<1$. More precisely, that class of functions $h(t)$ is included whose Fourier-Stieltjes transform[6] is a constant for all frequencies greater in absolute value than rW cps.

Let the number M be defined as

$$M \triangleq \max |h(t)|, \qquad -\infty < t < \infty, \quad (6)$$

so that M is equal to the largest absolute value which the function $h(t)$ assumes. Then it is shown in Appendix I that the truncation error $E(t)$ is bounded by

$$|E(t)| \leqq \frac{4M}{\pi^2 N(1-r)} = \frac{4M}{\pi^2 Nq}, \quad -\infty < t < \infty, \quad (7)$$

where, for convenience, the number q has been defined by

$$q \triangleq 1 - r. \quad (8)$$

As an example, suppose that $N=24$, $W=1000$ cps, and $h(t)$ has a highest frequency of 750 cps. Then $E(t)$ is bounded by $0.068M$. In terms of a normalized $h(t)$ with a largest value of unity, the absolute value of the truncation error is less than 0.068.

It is also possible to give an upper bound to the truncation error for the case where the truncation interval is asymmetrically placed with respect to the time t at which the cardinal series is being evaluated. Thus if

$$h_2(t) = \sum_{n=K-N_1}^{K+N_2} h\left(\frac{n}{2W}\right) \operatorname{sinc} 2W\left(t - \frac{n}{2W}\right), \quad (9)$$

the truncation error $E(t)$ is bounded as

$$|E(t)| \leq \frac{2M}{\pi^2(1-r)}\left[\frac{1}{N_1} + \frac{1}{N_2}\right], \quad -\infty < t < \infty. \quad (10)$$

The truncation errors given in later sections of this paper can be modified in the same manner by making minor changes in the proofs given in the Appendixes.

Truncation Error for a Self-Truncating Sampling Expansion

It is possible to obtain sampling expansions where the upper bound on the truncation error decreases more rapidly with N than is the case for the cardinal series. The product

$$h(t)\left\{\operatorname{sinc}\left(\frac{2qW}{m}(t'-t)\right)\right\}^m,$$

for constant t' and m a positive integer, is bandlimited to W cps if $h(t)$ is bandlimited to rW cps. This fact is easily shown by performing m repeated convolutions of the Fourier transform of

$$\operatorname{sinc}\left[\frac{2qW}{m}(t'-t)\right]$$

with the Fourier transform of $h(t)$.

Now if

$$h(t)\left\{\operatorname{sinc}\left[\frac{2qW}{m}(t'-t)\right]\right\}^m$$

is substituted for $h(t)$ in (1), the result is the expansion

$$h(t) = \sum_{n=-\infty}^{\infty} h\left(\frac{n}{2W}\right)\left\{\operatorname{sinc}\left[\frac{2qW}{m}\left(t - \frac{n}{2W}\right)\right]\right\}^m \cdot \operatorname{sinc} 2W\left(t - \frac{n}{2W}\right). \quad (11)$$

Let the truncation error $E_1(t)$ for (11) be given by

$$E_1(t) \triangleq h(t) - \sum_{n=K-N}^{K+N} h\left(\frac{n}{2W}\right)\left\{\operatorname{sinc}\left[\frac{2qW}{m}\left(t - \frac{n}{2W}\right)\right]\right\}^m \cdot \operatorname{sinc} 2W\left(t - \frac{n}{2W}\right), \quad (12)$$

where K is again an integer chosen according to (4). If the integer m is set equal to the approximately optimum value[7] $Nq\pi/e$, then, as is shown in Appendix II, the truncation error $E_1(t)$ is bounded for all values of t by

$$|E_1(t)| \leq 1.48M(qN - 1.74)^{-1/2}(3.2)^{-qN}, \quad (13)$$

where M has been defined previously by (6).

For the same constants used in the example following (8), where $N=24$, $W=1000$ cps, and $q=\frac{1}{4}$, the bound on $E_1(t)$ is approximately $6.8\times10^{-4}M$, which is 1/100 as large as the bound of the previous example.

Truncation Error of the Fogel Expansion

A sampling expansion which operates on simultaneous samples of a function $h(t)$ and its first derivative $h^{(1)}(t)$, each sampled at half of the rate of the cardinal series (1), has been obtained by Fogel.[8] This sampling expansion is

$$h(t) = \sum_{n=-\infty}^{\infty}\left[h\left(\frac{n}{W}\right) + \left(t - \frac{n}{W}\right)h^{(1)}\left(\frac{n}{W}\right)\right] \cdot \left[\operatorname{sinc} W\left(t - \frac{n}{W}\right)\right]^2. \quad (14)$$

[6] S. Bochner, "Lectures on Fourier Integrals," Princeton University Press, Princeton, N. J., pp. 317–331; 1959.

[7] Since m was assumed to be a positive integer, the product Nq must be adjusted so that $Nq\pi/e$ is such an integer for these results to hold.

[8] L. J. Fogel, "A note on the sampling theorem," IRE Trans. on Information Theory, vol. IT-1, pp. 47–48; March, 1955.

Again $h(t)$ is assumed to be a function whose Fourier-Stieltjes transform is a constant for all frequencies greater than rW cycles, where $r<1$. If the same process used in transforming the cardinal series (1) into (11) is applied to expansion (14), the result is

$$h(t) = \sum_{n=-\infty}^{\infty} \left[h\left(\frac{n}{W}\right) + \left(t - \frac{n}{W}\right) h^{(1)}\left(\frac{n}{W}\right) \right] \cdot \left[\operatorname{sinc} \frac{2qW}{m}\left(t - \frac{n}{W}\right) \right]^m \left[\operatorname{sinc} W\left(t - \frac{n}{W}\right) \right]^2, \quad (15)$$

which is a self-truncating version of the Fogel sampling expansion.

Let the truncation error $E_2(t)$ of expansion (14) and the truncation error $E_3(t)$ of expansion (15) be defined in a manner analogous to (12) so that n still assumes the values $n=K-N, \cdots, K+N$, where now inequality (4) is replaced by $Wt-\frac{1}{2}\leq K(t)\leq Wt+\frac{1}{2}$. Then, as is shown in Appendix III, $E_2(t)$ is bounded by inequality (7) and $E_3(t)$ by

$$| E_3(t) | \leq 2.1M(qN - 0.87)^{-1/2}(3.2)^{-2qN}, \quad (16)$$

where M and q have been defined previously and where m is an integer which is set equal to the approximately optimum value[7] $m=Nq2\pi/e$.

For the example used before, where $N=24$, $W=1000$ cps, and $q=\frac{1}{4}$, the bound on $E_3(t)$ is approximately $8\times10^{-7}M$.

Appendix I

Derivation of an Upper Bound on the Cardinal Series Truncation Error

It has been assumed that $h(t)$ is band-limited to rW cps so that[9]

$$h(t) = \int_{-rW}^{rW} S(f)e^{j2\pi ft}df. \quad (17)$$

Let the complex variable s be defined by

$$s \triangleq t + j\sigma. \quad (18)$$

It follows from the Schwarz inequality and (17) that

$$| h(s) |^2 \leq \int_{-rW}^{rW} | S(f) |^2 df \int_{-rW}^{rW} e^{-4\pi f\sigma} df, \quad (19)$$

or, if $S(f)$ is of integrable absolute square,

$$| h(s) |^2 \leq K \frac{\sinh 4\pi rW\sigma}{4\pi rW\sigma} \leq Ke^{4\pi rW|\sigma|}, \quad (20)$$

where K is a constant defined by

$$K \triangleq 2rW \int_{-rW}^{rW} | S(f) |^2 df. \quad (21)$$

[9] The argument that follows can be broadened so that $h(t)$ contains constants and sinusoids by writing $h(t)$ as the Fourier-Stieltjes transform of a function of bounded variation.

Thus, $h(s)$ is of exponential type[10] $2\pi rW$. The lemma of Duffin and Schaeffer[11] states that a function $h(s)$ of exponential type $2\pi rW$ is bounded by

$$| h(s) | \leq Me^{2\pi rW|\sigma|}, \quad (22)$$

where M has already been defined by (6).

It is apparent from (22) that $h(s)$ has no poles in the finite s plane. Cauchy's integral formula may be used to write

$$\frac{h(\xi)}{\sin 2\pi W\xi} = \frac{1}{2\pi j}\oint_C \frac{h(s)}{(s-\xi)\sin 2\pi Ws} ds, \quad (23)$$

where the contour C may be a small circle which surrounds only the pole $s=\xi$. It is assumed that ξ is real as shown in Fig. 1. Now let the contour C be deformed to the new contour C' as shown in the same figure. Since C' still encloses only the single pole $s=\xi$, C may be replaced by C' in the contour integral of (23).

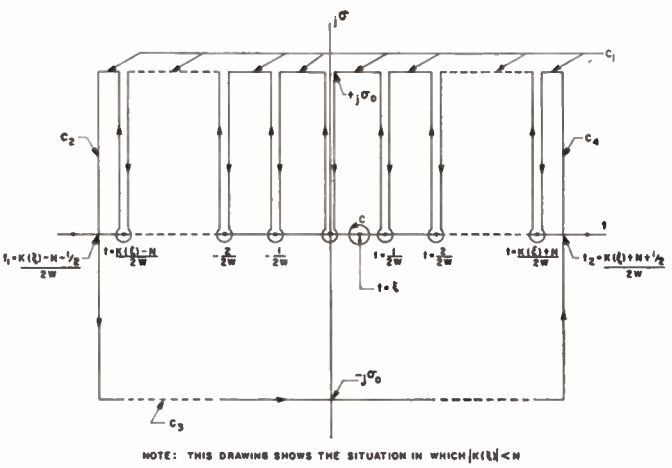


Fig. 1—The contours C and C' in the complex s plane.

Eq. (23) may be conveniently divided into the following parts corresponding to the parts of C':

1) The integrals evaluated on the clockwise contours which encircle each of the poles $s=n/2W$, where $n=K(\xi)-N$, $K(\xi)-N+1, \cdots, K(\xi)+N-1$, $K(\xi)+N$; the sum of these integrals will be denoted by $Q(\xi)$.
2) The integrals along the straight lines leading to and from the poles $s=n/2W$; these integrals cancel each other.
3) The integrals along each of the two horizontal contours C_1 and C_3.
4) The integrals along each of the two vertical contours C_2 and C_4.

[10] R. P. Boas, "Entire Functions," Academic Press, Inc., New York, N. Y., p. 8; 1954.

[11] R. Duffin and A. C. Schaeffer, "Some properties of functions of exponential type," *Bull. Am. Math. Soc.*, vol. 44, pp. 236–240; April, 1938.

The clockwise contour integrals around each of the poles $s=n/2W$ give rise to the sum of terms

$$Q(\xi) = -\sum_{n=K(\xi)-N}^{K(\xi)+N} \frac{(-1)^n h\left(\frac{n}{2W}\right)}{2\pi W\left(\frac{n}{2W}-\xi\right)}. \tag{24}$$

This sum may be subtracted from both sides of (23) to yield

$$\frac{h(\xi)}{\sin 2\pi W\xi} - Q(\xi) = \frac{1}{2\pi j}\int_{C_1+C_2+C_3+C_4} \frac{h(s)ds}{(s-\xi)\sin 2\pi Ws}. \tag{25}$$

Since the left side of (25) is equal to $E(\xi)/\sin 2\pi W\xi$ [where $E(\xi)$ is defined by (5)], an upper bound on the right side of (25) is equivalent to an upper bound on the truncation error $E(\xi)$.

It is easy to show that the integrals along contours C_1 and C_3 approach zero in the limit as σ_0 approaches infinity. The relationship

$$\begin{aligned} |\sin(x+jy)| &= |\sin x\cosh y + j\cos x\sinh y| \\ &= [\cosh^2 y - \cos^2 x]^{1/2} \geq \sinh|y| \end{aligned} \tag{26}$$

indicates that the denominator of the integrand $|h(s)/(s-\xi)\sin 2\pi Ws|$ dominates the numerator as σ_0 approaches infinity since, as proved by (22), $h(s)$ increases no faster than $Me^{2\pi rW|\sigma|}$.

After these preliminaries, the bound on the truncation error can be written as

$$\begin{aligned} |E(\xi)| \leq{}& \frac{1}{2\pi}\int_{C_2}\left|\frac{h(s)}{(s-\xi)\sin 2\pi Ws}\right|d\sigma \\ &+ \frac{1}{2\pi}\int_{C_4}\left|\frac{h(s)}{(s-\xi)\sin 2\pi Ws}\right|d\sigma. \end{aligned} \tag{27}$$

If it is recognized that contour C_2 lies along the line

$$t_1 = \frac{K(\xi) - N - \frac{1}{2}}{2W}$$

and C_4 along

$$t_2 = \frac{K(\xi) + N + \frac{1}{2}}{2W},$$

then, with the use of (22) and (26), (27) may be written as

$$\begin{aligned} |E(\xi)| \leq{}& \frac{1}{2\pi}\int_{-\infty}^{\infty} \frac{Me^{2\pi rW|\sigma|}d\sigma}{[(t_1-\xi)^2+\sigma^2]^{1/2}\cosh 2\pi W\sigma} \\ &+ \frac{1}{2\pi}\int_{-\infty}^{\infty} \frac{Me^{2\pi rW|\sigma|}d\sigma}{[(t_2-\xi)^2+\sigma^2]^{1/2}\cosh 2\pi W\sigma}. \end{aligned} \tag{28}$$

The integer $K(\xi)$ is constrained to satisfy (4), which implies that

$$|t_i - \xi| \geq N/2W, \qquad i = 1, 2. \tag{29}$$

With the additional inequality

$$\cosh 2\pi W\sigma \geq \tfrac{1}{2}e^{2\pi W|\sigma|}, \tag{30}$$

(28) becomes

$$|E(\xi)| \leq \frac{4M}{\pi}\int_0^{\infty} \frac{e^{-2\pi W(1-r)\sigma}}{\left[\left(\frac{N}{2W}\right)^2+\sigma^2\right]^{1/2}}d\sigma. \tag{31}$$

After defining $\alpha \triangleq (2W/N)\sigma$, (31) becomes

$$\begin{aligned} |E(\xi)| &\leq \frac{4M}{\pi}\int_0^{\infty} \frac{e^{-\pi N(1-r)\alpha}d\alpha}{[1+\alpha^2)^{1/2}} \\ &\leq \frac{4M}{\pi}\int_0^{\infty} e^{-\pi N(1-r)\alpha}d\alpha, \end{aligned} \tag{32}$$

or

$$|E(\xi)| \leq \frac{4M}{\pi^2 N(1-r)}, \tag{33}$$

which is identical to (7).

Appendix II

Derivation of an Upper Bound on the Truncation Error for a Self-Truncating Sampling Expansion

Cauchy's integral formula may be used to write

$$\frac{h(\xi)}{\sin 2\pi W\xi} = \frac{1}{2\pi j}\oint_{C'} \frac{h(s)\left[\operatorname{sinc}\frac{2qW}{m}(s-\xi)\right]^m}{(s-\xi)\sin 2\pi Ws}ds, \tag{34}$$

where the contour C' has already been defined by Fig. 1. Exactly as in Appendix I, it follows that

$$\begin{aligned} &|E_1(\xi)| \\ &\leq \left|\frac{1}{2\pi}\int_{C_1+C_2+C_3+C_4} \frac{h(s)\left[\operatorname{sinc}\frac{2qW}{m}(s-\xi)\right]^m}{(s-\xi)\sin 2\pi Ws}ds\right|, \end{aligned} \tag{35}$$

where $E_1(\xi)$ has been defined by (12). Eq. (26) establishes that

$$\left|\sin\frac{2\pi qWs}{m}\right|^m \leq \left[\cosh\frac{2\pi qW\sigma}{m}\right]^m \tag{36}$$

and hence that

$$\left|\sin\frac{2\pi qWs}{m}\right|^m \le e^{2\pi qW|\sigma|}. \tag{37}$$

As before, the limits as $\sigma_0\to\infty$ of the contour integrals on C_1 and C_3 are zero since the integrand of (35) is not greater than $M|\sigma_0|^{-(m+1)}$ along these contours. The result is that (35) becomes

$$|E_1(\xi)| \le \frac{4M}{\pi}\int_0^\infty \frac{d\sigma}{\left(\frac{2\pi qW}{m}\right)^m\left[\left(\frac{N}{2W}\right)^2+\sigma^2\right]^{(m+1)/2}}, \tag{38}$$

where use has been made of (29), (30), and (37). The right side of (38) may be replaced by a larger integral. If p is chosen as the integer such that

$$\frac{m}{2} \le p \le \frac{m+1}{2}, \tag{39}$$

then, with α defined as $\alpha \triangleq (2W/N)\sigma$,

$$|E_1(\xi)| \le \frac{2M}{\pi}\left(\frac{m}{\pi qN}\right)^m\int_{-\infty}^\infty \frac{d\alpha}{(1+\alpha^2)^p}. \tag{40}$$

It should be noted that $m\ge 1$ and, consequently, $p\ge 1$.

The integral on the right side of (40) may be evaluated by contour integration[12]

$$\int_{-\infty}^\infty \frac{d\alpha}{(1+\alpha^2)^p} = \frac{4\pi}{2^{2p}}\binom{2(p-1)}{p-1} \tag{41}$$

where $\binom{a}{b}$ is the binomial coefficient.

It is now necessary to establish the inequality[13]

$$\binom{2n}{n} < 4^n(\pi n)^{-1/2}. \tag{42}$$

The binomial coefficient may be written in the alternative forms

$$\binom{2n}{n} = 2^{2n}\prod_{k=1}^n\left(1-\frac{1}{2k}\right) = \frac{2^{2n}}{2n+1}\prod_{k=1}^n\left(1+\frac{1}{2k}\right), \tag{43}$$

and the square of (43) becomes

$$\left[\binom{2n}{n}\right]^2 = \frac{4^{2n}}{2n+1}\prod_{k=1}^n\left(1-\frac{1}{4k^2}\right). \tag{44}$$

The product formula[14] for $(\sin\pi x)/(\pi x)$ gives, for the case $x=\frac{1}{2}$,

$$\prod_{k=1}^\infty\left(1-\frac{1}{4k^2}\right) = \frac{2}{\pi}. \tag{45}$$

Now (44) may be written as

$$\left[\binom{2n}{n}\right]^2 = \frac{4^{2n}}{2n+1}\,\frac{2}{\pi}\,\frac{1}{\prod_{k=n+1}^\infty\left(1-\frac{1}{4k^2}\right)}. \tag{46}$$

The infinite product in (46) may be written

$$\begin{aligned}\prod_{k=n+1}^\infty\left(1-\frac{1}{4k^2}\right) &= \left[1-\frac{1}{2(n+1)}\right]\prod_{k=n+1}^\infty\left[\left(1-\frac{1}{2(k+1)}\right)\left(1+\frac{1}{2k}\right)\right]\\ &= \left[1-\frac{1}{2(n+1)}\right]\prod_{k=n+1}^\infty\left[\left(1+\frac{1}{4k(k+1)}\right)\right]\\ &> \left[1-\frac{1}{2(n+1)}\right] > 1-\frac{1}{2n+1} = \frac{2n}{2n+1}.\end{aligned} \tag{47}$$

After this substitution, the square root of (46) becomes the inequality (42).

Inequality (42) may be used in (41) to yield

$$\int_{-\infty}^\infty \frac{d\alpha}{(1+\alpha^2)^p} < [\pi/(p-1)]^{1/2}. \tag{48}$$

This result may be substituted in (40) to give

$$|E_1(\xi)| < \frac{2M}{\left[\pi\left(\frac{m}{2}-1\right)\right]^{1/2}}\left(\frac{m}{\pi qN}\right)^m, \tag{49}$$

where use has been made of the fact that $p\ge m/2$.

It is desirable to select a value of m which minimizes the right side of (49). A routine calculation shows that m should satisfy the equation

$$\ln\left(\frac{m}{\pi qN}\right) = \frac{5-2m}{2m-4}. \tag{50}$$

The right side of this equation approaches -1 as m increases, so that, for $m\gg 1$, an approximately optimum value of m is given by[7]

$$m = Nq\pi/e, \tag{51}$$

where e is the base of natural logarithms. Eq. (51) could be obtained by minimizing only the last term (49) and hence is the result of the assumption that $(m/2-1)^{-1/2}$ varies slowly compared to m^m. The use of (51) in (49) gives

$$|E_1(\xi)| \le 1.48M(Nq-1.74)^{-1/2}(3.2)^{-qN}, \tag{52}$$

which was to be proved.

[12] D. Bierens de Haan, "Nouvelles Tables d'Integrales Definies," Hafner Publishing Co., New York, N. Y., Table 17, No. 21; 1957.

[13] The statement and proof of inequality of (42) are due to B. F. Logan and V. A. Vyssotsky.

[14] E. C. Titchmarsh, "The Theory of Functions," Oxford University Press, New York, N. Y., p. 34; 1939.

Appendix III

Derivation of an Upper Bound on the Truncation Error for the Fogel Sampling Expansion and for a Self-Truncating Form of the Fogel Sampling Expansion

In the same manner as in the previous appendixes, Cauchy's integral formula allows the equality

$$\frac{h(\xi)}{[\sin \pi W\xi]^2} = \frac{1}{2\pi j}\oint_{C'''} \frac{h(s)}{(s-\xi)[\sin \pi Ws]^2}\,ds, \quad (53)$$

where the contour C''' is almost the same as the contour C' in Fig. 1, except that $2W$ is to be replaced by W wherever $2W$ occurs in Fig. 1, thereby making C''' twice as wide as C'. By the same arguments used in Appendix I, the truncation error $E_2(\xi)$ is bounded by

$$|E_2(\xi)| \leq \frac{1}{2\pi}\int_{C_1+C_2+C_3+C_4} \frac{|h(s)|\,ds}{|s-\xi|\,|\sin \pi Ws|^2}. \quad (54)$$

As before the contour integrals along C_1 and C_3 are zero, and (54) becomes

$$|E_2(\xi)| \leq \frac{8M}{\pi}\int_0^\infty \frac{e^{-2\pi W(1-r)\sigma}}{\left[\left(\frac{N}{W}\right)^2+\sigma^2\right]^{1/2}}\,d\sigma, \quad (55)$$

where use has been made of (22), (29) [with $2W$ replaced by W], and (30). If the new integer N' is defined by

$$N' \triangleq 2N, \quad (56)$$

then it follows from (31) that

$$|E_2(\xi)| \leq \frac{8M}{\pi^2 N'(1-r)} = \frac{4M}{\pi^2 N(1-r)}, \quad (57)$$

which is identical to (7) as was to be proved.

The truncation error $E_3(\xi)$ is bounded by the same methods as in Appendix II. Instead of (34), the following relationship is used:

$$\frac{h(\xi)}{[\sin \pi W\xi]^2} = \frac{1}{2\pi j}\int_{C'''} \frac{h(s)\left[\operatorname{sinc}\frac{2qW}{m}(s-\xi)\right]^m}{(s-\xi)[\sin \pi Ws]^2}\,ds. \quad (58)$$

By the same procedures as before, it can be shown that

$$|E_3(\xi)| \leq \frac{8M}{\pi}\int_0^\infty \frac{d\sigma}{\left[\frac{2\pi qW}{m}\right]^m\left[\left(\frac{N'}{2W}\right)^2+\sigma^2\right]^{(m+1)/2}}. \quad (59)$$

A comparison of (38) and (59) shows that

$$|E_3(\xi)| \leq \frac{4M}{\left[\pi\left(\frac{m}{2}-1\right)\right]^{1/2}}\left(\frac{m}{\pi qN'}\right)^m. \quad (60)$$

On the assumption that m is an integer and has the optimum value[7] $(\pi qN')/e$, (60) becomes

$$|E_3(\xi)| \leq 2.1M(qN-0.87)^{-1/2}(3.2)^{-2qN},$$

where N' has been replaced by $2N$.

The Fogel expansion given by (14) might be described as a second-order Jagerman-Fogel expansion.[16] An upper bound on the truncation error of the pth order Jagerman-Fogel expansion (which operates on samples of $h(t)$ and its first $p-1$ derivatives each sampled at $t=np/2W$, $n=\cdots,-1,0,1,\cdots$) is obtained by replacing N on the right side of inequality (7) by pN and then multiplying the resulting expression by 2^{p-1}. This new expression gives an upper bound for truncation error resulting from summing only the terms denoted by $n=K-N,\cdots,K+N$, where now inequality (4) is replaced by $(2W/p)t-\frac{1}{2}\leq K(t)\leq(2W/p)t+\frac{1}{2}$. The same procedure can be applied to self-truncating versions of the pth order Jagerman-Fogel expansion.

Acknowledgment

The authors are grateful for the numerous suggestions made by Dr. H. S. McDonald of the Bell Telephone Laboratories.

[16] D. L. Jagerman, and L. J. Fogel, "Some general aspects of the sampling theorem," IRE Trans. on Information Theory, vol. IT-2, pp. 139–146; December, 1956.

Transferred Electron Amplifiers and Oscillators*

C. HILSUM†

Summary—In some semiconductors the conduction band system has two minima separated by only a small energy, and the lower minimum has associated with it a smaller electron effective mass than the upper minimum. At high electric fields it should be possible to transfer electrons to the upper minimum where they will have a lower mobility. The conductivity of a homogeneous crystal bar can therefore decrease as the field is increased, and it is conceivable that a differential negative resistance could be obtained. The conditions needed for obtaining negative resistance are examined, and calculations made for GaSb and semi-insulating GaAs. It appears that negative resistances should be observable in both these materials.

I. Introduction

NEGATIVE RESISTANCE effects in p-n junctions have attracted much interest since the invention of tunnel diodes, and some consideration has also been given to the possibility of obtaining such effects in homogeneous solids. Burgess[1] has analyzed the situation where the carrier density is a function of the electric field, and concludes that the mechanism is not promising. Krömer[2] has exploited the idea of transferring a fraction of the electron population to a part of the conduction band where they have negative effective mass. He has developed the theory of NEMAGS (negative effective mass amplifiers and generators) and shown that it may be possible to make devices for use at frequencies up to 10^{12} cps.

This concept of transferring electrons within the conduction band may be taken a stage further. For example, consider a conduction band system with one minimum U at an energy δE above another minimum L (Fig. 1). In zero electric field the carriers will share themselves between these two branches of the conduction band in a manner determined by δE, the lattice temperature T_0, and the relative densities of states. When a field is applied a redistribution of the population occurs. If the upper minimum U has a large effective mass associated with it, and the effective mass in L is much smaller, an electron with an energy greater than δE above the bottom of the conduction band will probably be scattered into U, where its mobility will be low. The conductivity of the crystal will thus decrease as the field is increased, and it is conceivable that a differential negative resistance could be obtained.[3] In this paper we investigate the conditions needed for obtaining negative resistance in such systems.

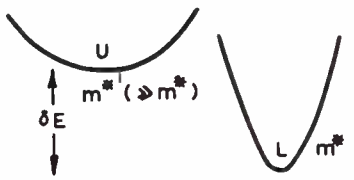


Fig. 1—Suitable conduction band for transferred electron amplifier.

II. Theory

Let us assume that in branch L the electrons have an effective mass m^*, a mobility μ, and a concentration n, and the density of states is N_c. In the upper branch U the corresponding values are $m^{*\prime}$, μ', n' and N_c'. Then if σ is the conductivity and N the total electron concentration, we have

$$\sigma = e[n\mu + n'\mu'] \tag{1}$$

$$\frac{n}{n'} = \frac{N_c}{N_c'} \exp(\delta E/kT_0) \tag{2}$$

$$n + n' = N \tag{3}$$

$$n = \frac{N_c}{N_c'} \frac{1}{1 + (N_c/N_c') \exp(\delta E/kT_0)} N \exp(\delta E/kT_0) \tag{4}$$

$$n' = N \frac{1}{1 + (N_c/N_c') \exp(\delta E/kT_0)} . \tag{5}$$

If the crystal bar used for the Transferred Electron Amplifier (TEA-bar) has a cross-sectional area A and a length d, the current-voltage relationship is

$$I = V\sigma A/d \tag{6}$$

and

$$\frac{dI}{dV} = \frac{A\sigma}{d} + \frac{AV}{d} \cdot \frac{d\sigma}{dV} , \tag{7}$$

if $V/d \equiv F$,

$$\frac{dI}{dV} = \frac{A}{d}\left(\sigma + F\frac{d\sigma}{dF}\right). \tag{8}$$

* Received by the IRE, June 21, 1961; revised manuscript received, October 30, 1961. Permission to publish this work was given by the Admiralty.

† Services Electronics Research Laboratory, Baldock, Hertfordshire, England.

[1] R. Burgess, "Negative resistance effects in homogeneous semiconductors," *Proc. Internatl. Conf. on Semiconductors*, Prague, Czech., pp. 818–821; 1960.

[2] H. Krömer, "Proposed negative mass microwave amplifier," *Phys. Rev.*, vol. 109, p. 1856; March, 1958.
—— "The physical principles of a negative-mass amplifier," PROC. IRE, vol. 47, pp. 397–406, March, 1959.

[3] B. K. Ridley and T. B. Watkins, "The possibility of negative resistance effects in semiconductors," *Proc. Phys. Soc.*, vol. 78, pp. 293–304; August, 1961. Recently these two authors have given a similar analysis of conduction in two bands. They consider the possibilities of obtaining a suitable band structure by straining p-type germanium.

The incremental change of conductivity is

$$\delta\sigma = e\delta n[\mu - \mu']. \tag{9}$$

If we assume that μ and μ' are field independent,

$$\frac{d\sigma}{dF} = e(\mu - \mu')\frac{dn}{dF}. \tag{10}$$

The assumption of field independent mobilities simplifies the algebra considerably, but it will not in general be justified. Its validity in particular cases is considered later. Eqs. (4) and (5) give the distribution of the electron population in thermal equilibrium. To calculate dn/dF we must make some assumptions about the distribution between U and L when there is a considerable departure from thermal equilibrium. We assume that:

1) The electron population as a whole has an energy distribution which can be characterized by an electron temperature T. This may be several times as large as T_0.
2) The distribution between U and L is given by the same equations as for thermal equilibrium [(4) and (5)] with T substituted for T_0. We may find dI/dV from (8) and (10) by expressing dn/dF as $(dn/dT)(dT/dF)$. We then have

$$\frac{d\sigma}{dF} = \frac{-e(\mu - \mu')n}{1 + (N_c/N_c')\exp(\delta E/kT)}\cdot\frac{\delta E}{kT}\cdot\frac{1}{T}\cdot\frac{dT}{dF} \tag{11}$$

$$\frac{dI}{dV} = \frac{A}{d}\left[n\mu e + n'\mu' e + F\frac{d\sigma}{dF}\right]$$

$$= \frac{AeN\mu'}{d[1+Z]}\left\{1 + Z\mu/\mu' - \frac{Z(\mu - \mu')}{\mu'(1+Z)}\cdot\frac{\delta E}{kT}\cdot\frac{F}{T}\cdot\frac{dT}{dF}\right\}, \tag{12}$$

where $Z = N_c/N_c' \exp \delta E/kT$.

A negative resistance will therefore occur if the TEA-bar is made of a semiconductor with both a suitable conduction band structure and a strong field dependence of electron temperature.

III. Application of 3-5 Compounds

The band structure required for a TEA occurs in the gallium and indium arsenides and antimonides. The conduction band in these materials has the lowest minimum at $k=0$, and minima a few tenths of an electron-volt higher along the [100] and [111] directions. These higher minima have associated with them an effective electron mass much larger than that of the lowest minimum. Further, all the 3-5 compounds have a large polar component in their binding, and the lattice mobility is limited by polar optical mode scattering. Stratton[4] has shown that when the transfer of energy to the lattice is through optical mode scattering, the electron temperature can be strongly dependent on field, yet the mobility will not vary strongly with field. We cannot make calculations for InAs and InSb, since no determinations of δE are available, but we have sufficient data for GaAs and GaSb to study them in more detail.

The present conception of the conduction band structure[5] of GaSb is shown in Fig. 2. The effective mass in L is $0.047m_0$, and in U $0.2m_0$. Let us take $\mu = 25{,}000$ cm²/volt sec, $\mu' = 5000$ cm²/volt sec, and $N_c'/N_c = 8$. In fact this ratio would be greater than 8 because of the multiplicity of equivalent minima along [111] directions, but by putting a low value for N_c'/N_c, we are taking a pessimistic view. A higher value will result in a more pronounced negative resistance region. For GaSb δE is 0.075 ev.

The relation between F and T has been given by Stratton[4] as

$$F^2 = F_0 2N_0^2(2/3\pi)G^2 \exp G[\exp(G_0 - G) - 1]K_0(G/2)$$
$$\cdot\{[\exp(G_0 - G) - 1]K_0(G/2)$$
$$+ [\exp(G_0 - G) + 1]K_1(G/2)\} \tag{13}$$

where

$$F_0 = (\epsilon_\infty^{-1} - \epsilon_0^{-1})m^* ek\theta/\hbar^2, \qquad N_0 = [\exp G_0 - 1]^{-1}$$

and $G_0 = \theta/T_0$, $G = \theta/T$, θ being the optical mode temperature and ϵ_0 and ϵ_∞ the static and high frequency dielectric constants respectively.

It is assumed in the derivation of this expression that interelectronic collisions are predominant for momentum and energy interchanges, and for this to be true the electron concentration must be greater than $(eF_0k\theta/2\pi e^{*4}G^{1/2})\exp(-G)$ for the case when T_0 is less than θ. This condition is satisfied for n-type GaSb with an electron concentration greater than 10^{14} cm^{-3}.

The curve in Fig. 3 is the F-T relation for GaSb with $T_0 = 70°$K. It is deduced from (13) taking $\epsilon_0 = 15$, $\epsilon_\infty = 14$, $\theta = 340°$K, and $m^* = 0.047\ m_0$. A current-voltage curve for a GaSb TEA-bar can now be derived and the related curve, that of $J/eN\mu'$ as a function of F, where J is the current density, is shown in Fig. 4. It can be seen that a negative resistance region occurs at fields greater than 500 volt/cm.

Unless GaSb can be made several orders of magnitude purer than at present, it will be difficult to apply such high voltages to it. With 10^{16} carriers per cm³ the conductivity is greater than 10 ohm^{-1} cm^{-1}, and at 500 volt/cm the power dissipated would be about 10 megawatts per cm³. It might be possible to circumvent this difficulty by using pure p-type material, which would

[4] R. Stratton, "The influence of interelectronic collisions on conduction and breakdown in polar crystals," *Proc. Roy. Soc.*, vol. 246A, pp. 406–422; August, 1958.

[5] A. Sagar, "Experimental investigation of conduction band of GaSb," *Phys. Rev.*, vol. 117, pp. 93–100; January, 1960.

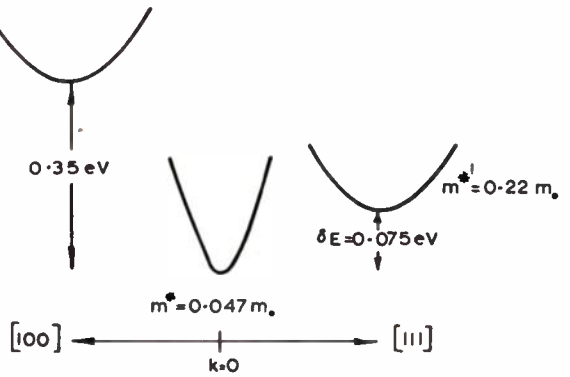


Fig. 2—Conduction band structure of GaSb.

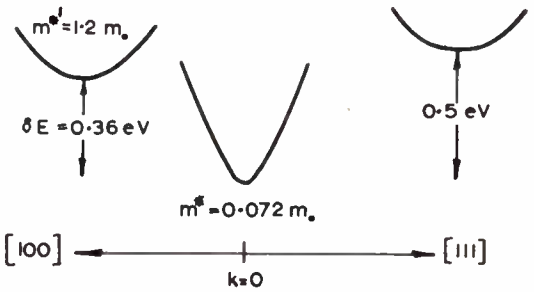


Fig. 5—Conduction band structure of GaAs.

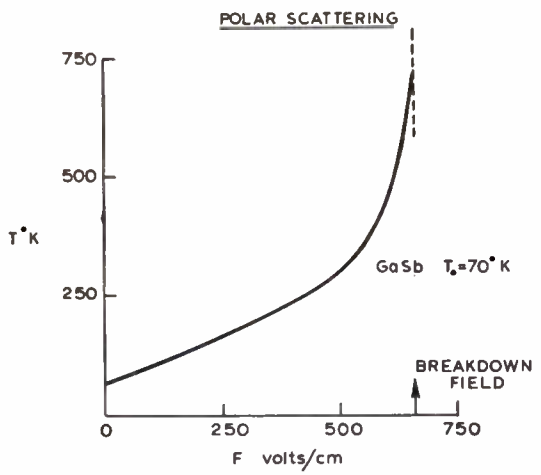


Fig. 3—*F-T* relation for GaSb at 70°K.

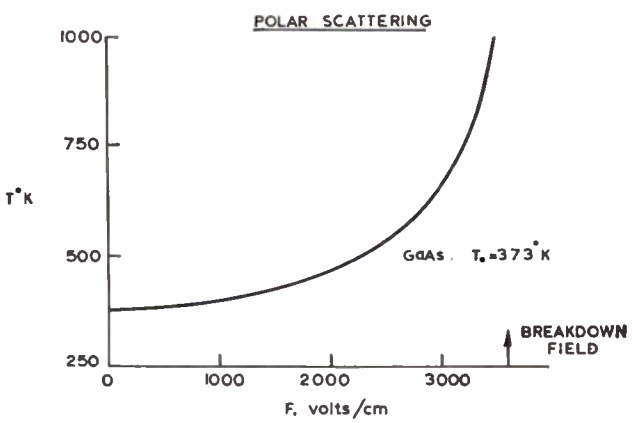


Fig. 6—*F-T* relation for GaAs at 373°K.

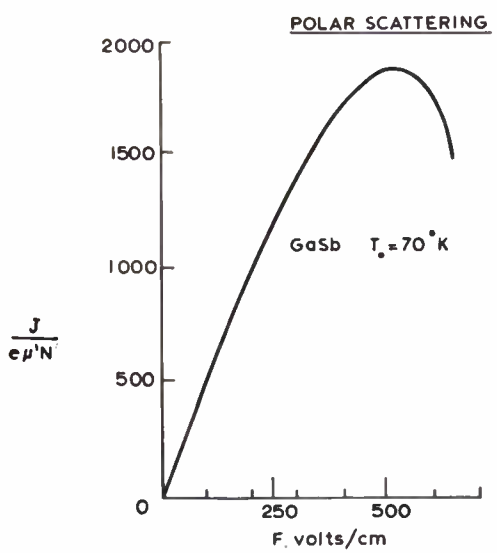


Fig. 4—Current density-field relation for GaSb at 70°K.

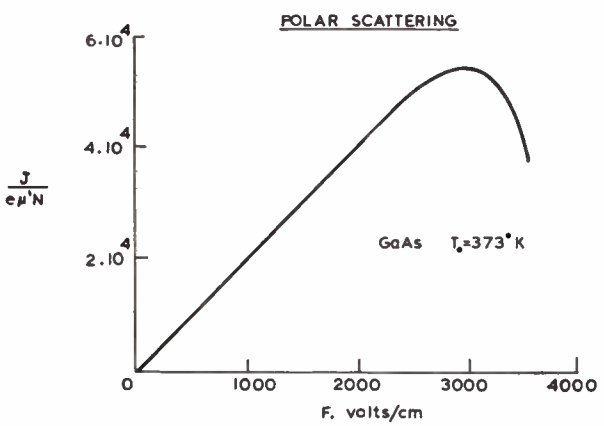


Fig. 7—Current density-field relation for GaAs at 373°K.

have a fairly high resistivity at 70°K, and creating in it a few electron-hole pairs by flooding with light. A more promising approach is to use a semiconductor with a high resistivity, such as semi-insulating GaAs.[6,7]

The conduction band structure of GaAs[8] is shown in Fig. 5. δE is 0.36 ev, so that the lattice temperature for optimum operation would be above room temperature. We may calculate the F-T relation using (13), taking $\theta = 410°K$, $\epsilon_0 = 12.5$, $\epsilon_\infty = 10.9$, $m^* = 0.072 m_0$, and $T_0 = 373°K$. We will later have to consider whether this equation is valid for semi-insulating GaAs. The results are given in Fig. 6. We may now deduce the current voltage curve of the GaAs TEA-bar, and we adopt the values $m^{*\prime} = 1.2 m_0$, $\mu = 10{,}000$ cm²/volt sec, $\mu' = 500$ cm²/volt sec, and $N_c'/N_c = 60$. We have neglected the multiplicity of [100] minima in calculating this ratio. The current-voltage relation is shown in Fig. 7, and there is a negative resistance region at fields above 3000 volts/cm.

IV. Validity of Assumptions

It is assumed in Section II that the carrier mobilities are field-independent, that the electron population as a whole can be characterized by a temperature T, and that the distribution of carriers between the two bands is given by a Boltzmann relationship involving this temperature T.

For materials where the dominant scattering mechanism is polar scattering the mobility is little dependent on field if the lattice temperature is near to the optical mode temperature. At low lattice temperatures the mobility falls markedly with field, and at high lattice temperatures there is a small increase in mobility with field. In the two cases we have considered earlier, with T_0 equal to 0.2θ for GaSb and 0.9θ for GaAs, our general conclusions are unaffected if we include in our calculations an allowance for the change of mobility.

The second assumption may not be valid for semi-insulating GaAs. Since the carrier concentration would be only 10^{10} per cm³, interelectronic collisions would not be the predominant mechanism in determining the energy and momentum distribution of the electrons. Eq. (13) is therefore not strictly applicable, and in this case it is doubtful if a meaningful temperature can be assigned to the electron population as a whole. This does not mean that our particular problem becomes insoluble, but we would have to change our approach.

When there are few interelectronic collisions in a polar semiconductor, it is likely that under the influence of a high electric field the electrons will divide into two groups X and Y, X with an effective temperature close to that of the lattice, and Y with a spread of energies extending from an energy rather greater than kT_0 up to an energy $k\theta$ and beyond. Group Y will probably not have a Maxwellian distribution of energy. If we assume that all electrons in Y will be scattered into the upper minima of the conduction band, since the density of states here is many times greater than that in the lowest minimum, we need not use the concept of electron temperature in our calculations. Instead we would need an expression for the relative populations of X and Y as a function of the field. Such an expression is not yet available in an explicit form.

Even if we assume that a meaningful temperature T can be assigned to the electron population as a whole, it is unlikely that we should obtain a correct value for T from the simple analysis of Section III. Here we deduced T from a theory which is applicable to conduction in a single band, and ignored interactions between the second band and the lattice. It is possible to include these interactions by modifying the relations used by Stratton. The expression he gives for conservation of energy is

$$JF = NeN_0F_0(2k\theta/\pi m^*)^{1/2}[\exp(G_0 - G) - 1]G^{1/2} \cdot \exp(G/2)K_0(G/2) \tag{14}$$

where $J = Ne\mu F$.

If we again assume that μ is independent of F, we obtain as the modified relation

$$F^2 = \frac{F_0N_0}{\mu}\left(\frac{2k\theta}{\pi m^*}\right)^{1/2}\left[1 + \frac{1}{Z}\left(\frac{m^{*\prime}}{m^*}\right)^{1/2}\right]\left[1 + \frac{\mu'}{Z\mu}\right]^{-1} \times [\exp(G_0 - G) - 1]G^{1/2}\exp(G/2)K_0(G/2). \tag{15}$$

Using the parameters given earlier for GaAs we may deduce F-T relations and the equivalent of a current-voltage curve, and these are shown in Figs. 8 and 9. A negative-resistance region is predicted again, but now a field of about 10,000 volts/cm is required. At this field the electron temperature is about 4 times the lattice temperature and our assumption of a field independent mobility must again be called into question. This point may be investigated by generalizing Stratton's second basic equation, that for the conservation of momentum, so that it applies to conduction in two bands. This calculation has been made by Rhoderick,[9] and his results are included in Figs. 8 and 9. Our assumption of a field independent mobility does not in fact lead to much error in predicting the general character of the behavior at high fields, but it does lead to an overestimate of the current which passes.

[6] J. W. Allen, "GaAs as a semi-insulator," *Nature*, vol. 187, pp. 403–405; July, 1960.
[7] C. H. Gooch, C. Hilsum, and B. R. Holeman, "Properties of semi-insulating GaAs," *J. Appl. Phys.* (Suppl.), vol. 32, pp. 2069–2073; October, 1961.
[8] H. Ehrenreich, "Band structure and electron transport of GaAs," *Phys. Rev.*, vol. 120, pp. 1951–1963; December, 1960.
[9] E. H. Rhoderick, private communication.

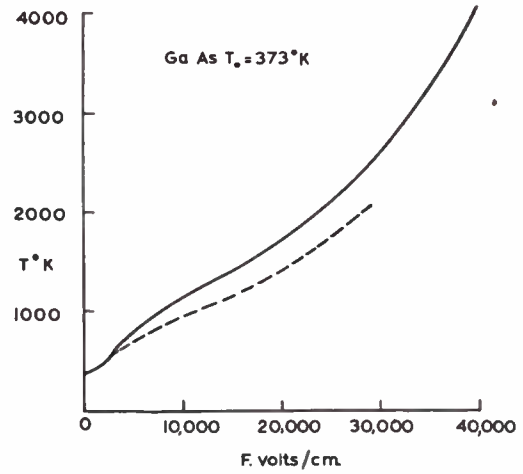


Fig. 8—*F-T* relation for GaAs at 373°K, using Stratton's theory modified for two-band conduction (solid line = mobility assumed field independent; broken line = effects of field dependent mobility included).

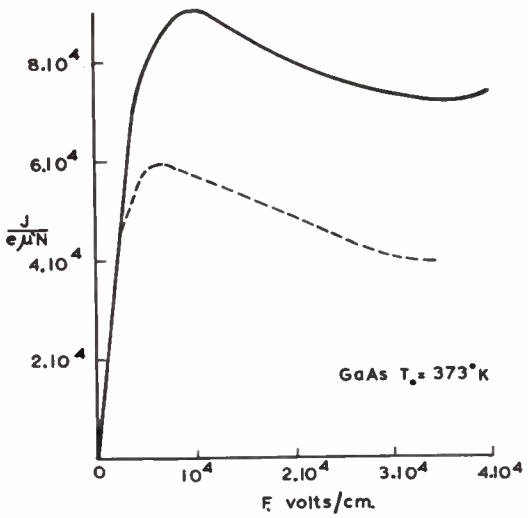


Fig. 9—Current density-field relation for GaAs at 373°K, using Stratton's theory modified for two-band conduction (solid line = mobility assumed field independent; broken line = effects of field dependent mobility).

We have not considered in this treatment the effect of impurity scattering or lattice acoustic scattering. In semi-insulating GaAs, the impurity concentration is probably greater than 10^{15} per cm^3, but the conventional theories of ionized impurity scattering may not be applicable since the donor and acceptor centers may be associated in pairs. Acoustic scattering in 3-5 compounds is negligible at low fields, but for electrons with an energy greater than $k\theta$, polar scattering is ineffective, and they may transfer energy to the lattice via the acoustic modes.

V. Practical Considerations

According to the simple theory of Section III, a TEA-bar of semi-insulating GaAs would be operated at 373°K at a field of 3200 volts/cm. The power consumption would be less than 100 watts/cm^3. This is a high value for continuous operation, but it should not be impracticable. It would not be advisable to operate at higher temperatures since σ would be higher, and the power consumption would be increased. If the modified theory of Section IV proves more appropriate the operating field will be 12,000 volts/cm, and the power consumption about 500 watts/cm^3, so that some form of intermittent operation would be necessary.

TEA-bars should oscillate at frequencies approaching 10^{12} cps, unless the design is such that stray inductance and capacity sets a lower limit. The efficiency as a generator would be low, an optimistic estimate being 1 per cent of the applied dc power. (The efficiency of a 4-mm klystron is about 0.2 per cent.)

VI. Improved Materials

The band structure of GaAs is not ideal for TEA operation at a temperature near 300°K. It would be preferable to use a material with a similar conduction band structure, but with δE only half as large. Such a material might be obtained by alloying GaAs with GaP. In GaP the lowest minima are those along the [100] directions, and the minima in the [111] directions are 0.1 ev higher. The minimum at the center of the zone is 0.25 ev higher still. As GaAs is added to GaP the central minimum drops relative to the [111] and [100] minima and in the semiconductor made from equal proportions of the two constituents, the minimum at $k=0$ is at the same energy as the [100] minima. We may therefore assume that in the alloy $GaAs_{0.7}P_{0.3}$, the lowest minimum would be that at $k=0$, and δE would be about 0.15 ev. Such a band structure would be very suitable for our purpose.

The material must, of course, be made semi-insulating, and as yet there is no evidence that this can be done.

Acknowledgment

The author is grateful to Dr. D. J. Oliver for checking some of the calculations, and to J. W. Allen, Dr. E. H. Rhoderick and Dr. O. Simpson for valuable discussions.

Current-Limited Rectifiers*

G. FRANKLIN MONTGOMERY†, SENIOR MEMBER, IRE

Summary—**Several ac rectifier circuits have the property, inherent in the rectification process, of limiting their output current to a specified maximum. They are technically simple and can be used as the foundation for power supplies that remain undamaged and absorb little power with a short-circuited load. In combination with a dc regulator, such a rectifier provides a constant-voltage supply with desirable overload characteristics. A modified bridge rectifier is especially useful.**

Introduction

CURRENT LIMITING is a useful feature in the design of ac rectifier power supplies. What is often wanted, for short-circuit protection or for other reasons, is a supply whose load voltage is constant for load currents less than some chosen value and small or zero for load currents greater than the chosen value. An ideal characteristic is shown in Fig. 1. Such a characteristic is provided, for example, by a source of constant direct current I_0 in parallel with a perfect shunt regulator of voltage E_L.

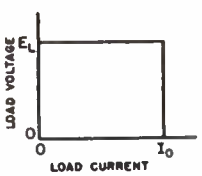


Fig. 1—Ideal current-limited load characteristic.

Several current-limiting methods are available. These include constant-current transformers or controlled saturable reactors that regulate the ac input to the rectifier, and series-, shunt-, or combination regulators that control the dc rectifier output. The dc regulators are augmented by one or more current-sensitive devices that disable the regulator when the load current exceeds a chosen value, thus permitting a rapid decrease in load voltage. A third possible choice is the use of controlled rectifiers.

For power supplies using dc regulation, there are circuits that limit the current with little or no increase in circuit complexity. These methods take advantage of inherent properties of the rectification process, and it is the object of this paper to discuss some of their basic characteristics.

Limiting Rectifier Circuits

The following circuit diagrams have been stripped to their essential elements to clarify behavior and to emphasize the ideal conditions of the analysis. Rectifiers are assumed to have zero voltage drop when conducting and zero reverse current when blocked. Inductors are assumed lossless. A zero-impedance battery is substituted for the dc load in accordance with the compensation theorem. The analytical effect of this substitution is to eliminate ripple voltage at the load, an effect that is approached in practice by using a large filter capacitance. The circuits are taken up in order of increasing complexity, although the electrical behavior of the simple half-wave rectifiers requires a more roundabout analysis than the rest. Half-wave circuit performance differs most from the optimum of Fig. 1. The difference, however, may have little practical importance for applications where economy and circuit simplicity are dominant.

The first circuit is shown in Fig. 2. If an inductor-input smoothing filter is used with a half-wave rectifier, the output voltage for a given output current is a decreasing function of the input inductance, and the maximum current that can be drawn from the supply is limited by the inductance.[1] Since we often strive for a constant output voltage, this property of the half-wave circuit is usually thought to be a disadvantage, and a method has been devised to correct it.[2] On the other hand, the inductor is a lossless means for limiting the short-circuit output current to a safe value. Within a given range of load current, constant load voltage can be maintained with a dc regulator.

Fig. 3 is a plot of the relation between normalized output voltage (E_B/E) and normalized output current ($\omega LI/E$), where E_B is the output voltage, E is the amplitude of the sinusoidal source voltage, I is the direct component of the rectifier output current, L is the inductance, and $(\omega/2\pi)=f$ is the source frequency. This relation is derived in the Appendix, where it is assumed that the inductor is lossless and that rectifier forward resistance and reverse conductance are both zero. Determining (E_B/E) as an explicit function of $(\omega LI/E)$ does not seem to be possible, and the given relation is parametric. A useful approximation, however, is

$$E_B/E = (0.9)\left[1 - \sqrt{\omega LI/E}\right], \tag{1}$$

and this function is shown as the dashed curve in Fig. 3.

These results also apply to the circuit of Fig. 4, which uses a modified full-wave (center-tap), half-bridge rectifier. The modification is somewhat clumsy in the re-

* Received by the IRE August 1, 1961; revised manuscript received, October 13, 1961.

† National Bureau of Standards, Washington, D. C.

[1] M. B. Stout, "Behavior of half-wave rectifiers, *Electronics*, vol. 12, pp. 32–34; September, 1939.

[2] G. F. Montgomery, "Improving rectifier circuits," *Electronics*, vol. 34, p. 88; April 7, 1961.

quirement of a separate inductor for each rectifier. However, in addition to eliminating asymmetrical current in the source, this circuit doubles both the output current and the ripple frequency, so that subsequent smoothing is accomplished more easily. The output current is $2I$, where I corresponds to the definitions for Fig. 2.

The third circuit is shown in Fig. 5. In this arrangement, the rectifier current is limited by the capacitance C, and inductance L is assumed to be effectively infinite. If (E_B/E) is the normalized output voltage and $(I/\omega CE)$ the normalized output current, there is a symmetrical correspondence between these variables and the variables of the first circuit.[3] Specifically, if (E_B/E) in Fig. 3 is replaced by $(I/\omega CE)$, and if $(\omega LI/E)$ is replaced by (E_B/E), then the curves describe the behavior of the circuit in Fig. 5. Approximation (1) also holds for this interchange of variables, becoming

$$E_B/E = [1 - (1.1)(I/\omega CE)]^2. \quad (2)$$

The inductance L has been assumed infinite in this case to simplify analysis; in practice, it should be chosen at least large enough to avoid resonance at the source frequency with what will ordinarily be a small C.

The fourth circuit, shown in Fig. 6, is a modification of the conventional voltage doubler. If the input capacitance C is intentionally made small, it will limit the rectifier current in a manner similar to the operation of Fig. 5. For ideal rectifiers as before, the relation between normalized output voltage (E_B/E) and normalized output current $(I/\omega CE)$ is explicit,[3] and

$$E_B/E = 2 - (2\pi)(I/\omega CE) = 2 - (I/fCE). \quad (3)$$

Since the equation is linear, it need not be plotted. The same equation holds for the "full-wave" circuit of Fig. 7.

The last circuit is shown in Fig. 8, in which the input capacitance C limits the current to the bridge rectifier. The output relation plotted in Fig. 9 is derived in the Appendix, where it is assumed that the filter inductance L is infinite. In practice, it is necessary only that this

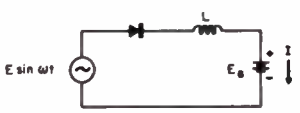


Fig. 2—Half-wave rectifier, inductance limited.

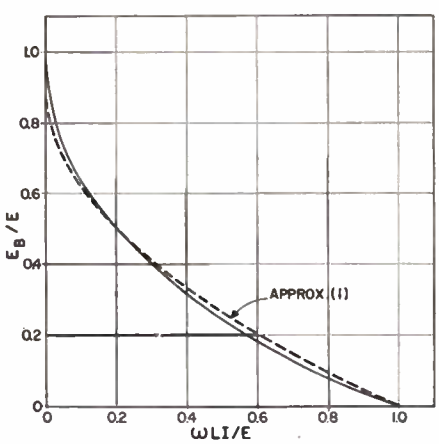


Fig. 3—Half-wave rectifier output characteristic.

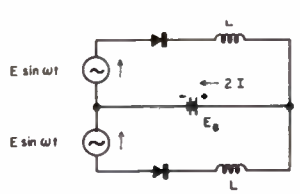


Fig. 4—Inductance-limited full-wave rectifier.

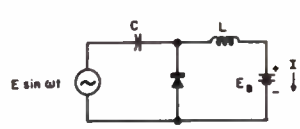


Fig. 5—Half-wave rectifier, capacitance limited.

[3] G. F. Montgomery, "The diode reactance modulator," *Trans. AIEE*, vol. 78 (*Commun. and Electronics*), pp. 980–983; January, 1959.

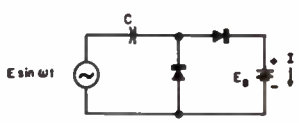


Fig. 6—Capacitance-limited voltage doubler.

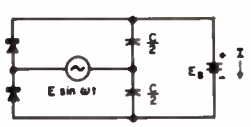


Fig. 7—Capacitance-limited full-wave voltage doubler.

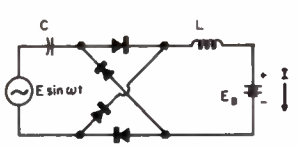


Fig. 8—Bridge rectifier, capacitance limited.

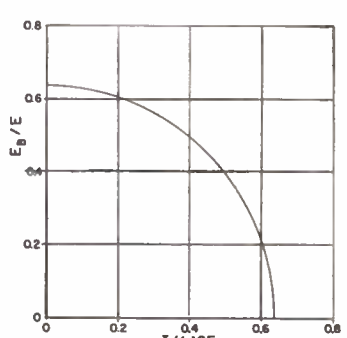


Fig. 9—Bridge rectifier output characteristic.

inductance be at least the critical inductance[4] for a minimum output current I. With a given inductance, and for output currents less than the critical current for that inductance, the normalized voltage departs from the curve shown and rises to $(E_B/E)=1$ at $I=0$, rather than to $(E_B/E)=2/\pi$. The curve in Fig. 9 is a circular arc,[5] for which

$$E_B/E = \sqrt{(2/\pi)^2 - (I/\omega CE)^2}. \quad (4)$$

Of the circuits considered—half-wave, voltage-doubler, and bridge—the bridge rectifier has the most attractive output characteristic, corresponding more nearly to the ideal of Fig. 1 than do the others. For a given load power, the short-circuit load current can be fixed easily at less than 1.5 times the maximum current at constant voltage.

Other modifications of conventional rectifier circuits are possible that have not been considered in detail, either because they appear to be less useful or because their behavior is predictable from the information given. For example, the bridge circuit of Fig. 8 can be used with a capacitor-input filter, in which case its output voltage is a linear function of the output current in analogy with the voltage-doubling rectifiers. It is worth noting that any rectifier can be current limited by inserting suitable reactances in series with its source. If the source is a transformer secondary winding, a reactance can be placed in series with the transformer primary, and no other modification of the circuit adjacent to the rectifier is needed. In many applications, however, the transformer supplying the rectifier also powers other constant-voltage circuits through auxiliary windings, and a primary series reactance cannot be used. The current-limiting rectifiers discussed here are useful in these circumstances.

Regulation

In Fig. 10, three current-limited rectifiers are shown with simple regulators to illustrate the general scheme. A gas voltage-regulator tube is used in the first example (a), the series resistance R being only large enough to suppress relaxation oscillation at low tube currents. Semiconductor breakdown (voltage-regulating) diodes presumably require no series resistor. A second filter inductor can be substituted for R. For larger currents, or for more precise regulation, a shunt vacuum-tube regulator can replace the VR tube. The series regulator used in the second example (b) is a type whose characteristics are well known. The bridge rectifier (c) is shown with a shunt regulator, and its typical load characteristic illustrates the small difference that can be achieved between short-circuit load current and maximum rated current.

For any power supply design, a choice must be made among series regulation, shunt regulation, or a combination of the two. The regulator must dissipate a power W_R which is a function of the load current. In some cases, it may be important to minimize W_R by design. To do so, both the function $E_B(I)$ and the time distribution of the load current must be known. A solution of the minimization problem is not sought here, but the following conclusions are useful as design guides. Let E_L be the (constant) load voltage and I_L the (variable) load current. Then,

1) For a shunt regulator, the regulator dissipation $W_R=0$ for a load current $I_L \geq I_m$, where $E_B(I_m)=E_L$; W_R increases monotonically with decreasing load current, becoming a maximum $W_R=E_L I_m$ for $I_L=0$.
2) For a series regulator, the regulator dissipation $W_R=0$ for a load current $I_L=0$; W_R increases with increasing load current, becoming a maximum for $I_L=I_p$, where $0<I_p\leq I_s$ and $E_B(I_s)=0$. The value of I_p in any particular case depends upon the properties of $E_B(I)$, the choice of E_L, and the maximum conductance of the series element of the regulator.

It may be advantageous to choose the parameters of the rectifier in such a way that maximum output (and input) power occurs for maximum rated load current. For this choice, the load power and the power drawn from the source both decrease as the load current in-

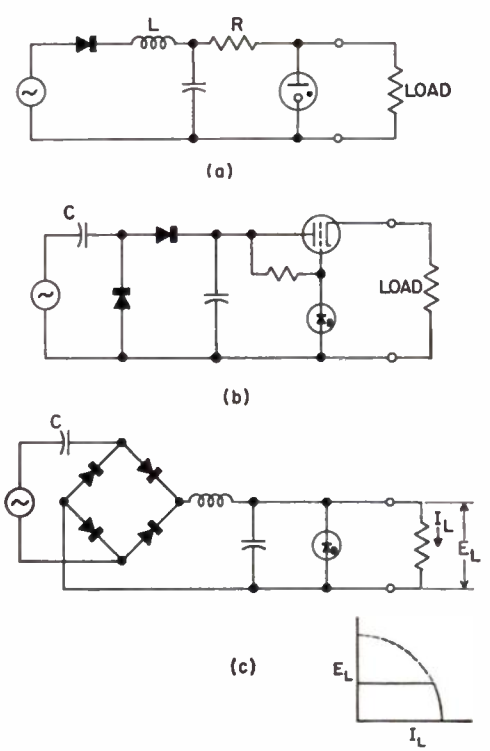


Fig. 10—Typical regulator circuits.

[4] F. E. Terman, "Radio Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., 4th ed., p. 715; 1955.

[5] This characteristic resembles the familiar variation of voltage across a variable resistor connected in series with a fixed capacitor and a sinusoidal voltage source. The analogy must not be carried too far, however. In Fig. 8 the source current is square, not sinusoidal. If inductance L is less than critical, the source current is neither sinusoidal nor rectangular, and the analogy disappears altogether.

creases above its rated maximum. The rectifier output power is $W_S = E_B I$. Using the derived equations for $E_B(I)$, the derivative dW_S/dI is found easily. Setting $(dW_S/dI) = 0$ yields the values of E_B and I for which W_S is a maximum, and these are listed in Table I. (The half-wave values are approximate.)

TABLE I
MAXIMUM OUTPUT POWER TABLE

Rectifier	Output voltage (E_B)	Output current (I)	Maximum power (W_S max)
Half-wave (Fig. 2)	$0.3\,E$	$0.44\,E/\omega L$	$0.13\,E^2/\omega L$
Half-wave (Fig. 5)	$0.44\,E$	$0.3\,\omega CE$	$0.13\,\omega CE^2$
Voltage doubler (Figs. 6, 7)	E	fCE	fCE^2
Bridge (Fig. 8)	$(\sqrt{2}/\pi)E$	$2\sqrt{2}\,fCE$	$(4/\pi)fCE^2$

For a shunt regulator, the tabular value of E_B is set equal to the regulated load voltage. For a series regulator, tabular E_B equals the regulated load voltage plus the drop in the regulator series element at maximum rated current. The amplitude of the source voltage is thus determined. The tabular value of I (equal to the maximum rated load current) is then used to determine the inductance or capacitance of the appropriate circuit element.

Conclusion

Six current-limited rectifier circuits that appear to be especially useful for short-circuit-proof, regulated power supplies have been examined. Current limiting occurs inherently in the rectification process for these circuits, and this behavior is brought out either by adding a reactive element to a conventional circuit or by modifying an element already present. The limiting obtained is essentially lossless; this property implies zero power consumption on short circuit and allows the load voltage to be dc regulated with minimal total dissipation.

Appendix

The Half-Wave Rectifier

In the circuit of Fig. 2, let the source voltage be $E\cos\omega t = E\cos\phi$, and let $i(\phi)$ be the rectifier current, $\theta = \phi_1 - \phi_0$ the angle of rectifier conduction, where ϕ_0 is the phase at which conduction begins and ϕ_1 is the phase at which conduction ends. Then,

$$\begin{cases} i = (1/\omega L)\displaystyle\int_{\phi_0}^{\phi} (E\cos\phi - E_B)d\phi, & \phi_0 \le \phi \le \phi_1 \\ i = 0, & \phi_1 \le \phi \le \phi_0 + 2\pi. \end{cases} \tag{5}$$

From (5), since $(di/d\phi) = 0$ at $\phi = \phi_0$,

$$(E_B/E) = \cos\phi_0 \tag{6}$$

and since $i(\phi_1) = 0$,

$$\sin\phi_1 - \sin\phi_0 = (E_B/E)(\phi_1 - \phi_0) = (\phi_1 - \phi_0)\cos\phi_0. \tag{7}$$

With the substitution $\phi_1 = \theta + \phi_0$, (7) becomes

$$\tan\phi_0 = -\frac{\theta - \sin\theta}{1 - \cos\theta}. \tag{8}$$

Now, by definition,

$$I = (1/2\pi)\int_{\phi_0}^{\phi_1} id\phi$$
$$= (1/2\pi)(E/\omega L)[\cos\phi_0 - \cos\phi_1 - (\phi_1 - \phi_0)\sin\phi_0 - (E_B/2E)(\phi_1 - \phi_0)^2], \tag{9}$$

which, with substitutions from (6), (7), and the definition of θ, becomes

$$(\omega LI/E) = (1/4\pi)(\cos\phi_0)(1 - \cos\theta - \sin\theta\tan\phi_0)^2. \tag{10}$$

Using θ as a parameter, Fig. 3 is plotted from (6), (8), and (10).

The Bridge Rectifier

In the circuit of Fig. 8, let the source voltage be $E\sin\omega t = E\sin\phi$, and let $i(\phi)$ be the capacitor current and $e_c(\phi)$ the capacitor voltage. Since the inductance L is infinite, $i = \pm I$, and the charge and discharge of the capacitor are linear. In order that the capacitor accumulate no charge, the charge and discharge intervals must be equal. Therefore,

$$\begin{cases} i = I, & \phi_0 \le \phi < \phi_0 + \pi \\ i = -I, & \phi_0 + \pi \le \phi < \phi_0 + 2\pi \end{cases} \tag{11}$$

where ϕ_0 is the phase at which first commutation occurs. Because of circuit symmetry, the average capacitor voltage is zero. It follows that the average is zero in either commutation interval. Then,

$$E_B = (1/\pi)\int_{\phi_0}^{\phi_0+\pi} (E\sin\phi - e_c)d\phi$$
$$= (1/\pi)\int_{\phi_0}^{\phi_0+\pi} E\sin\phi d\phi$$

or

$$(E_B/E) = (2/\pi)\cos\phi_0. \tag{12}$$

At first commutation, $E\sin\phi_0 - e_c(\phi_0) = 0$. Then,

$$e_c = E\sin\phi_0 + \int_{\phi_0}^{\phi} (I/\omega C)d\phi$$
$$= E\sin\phi_0 + (I/\omega C)(\phi - \phi_0), \quad \phi_0 \le \phi \le \phi_0 + \pi. \tag{13}$$

Since the capacitor charges linearly, $e_c(\phi_0 + \pi/2) = 0$, and from (13),

$$E\sin\phi_0 + (I/\omega C)(\pi/2) = 0$$

or

$$(I/\omega CE) = -(2/\pi)\sin\phi_0. \tag{14}$$

Eq. (4) is a combination of (12) and (14).

Beta Cutoff Frequencies of Junction Transistors*

J. LINDMAYER† AND C. Y. WRIGLEY†

Summary—**The 3-db cutoff frequency and the gain-bandwidth product are calculated for the common-emitter current gain. The calculation is made in a simple manner and in general form so as to apply to all transistors. The gain-bandwidth product is shown to be determined by the transit time across the base and the space-charge capacitances. The gain-bandwidth product reaches a maximum for drift transistors at a critical current. This current is given.**

Introduction

THE common-emitter current gain is part of any transistor specification, since transistors are used mostly in this configuration. It has been observed that for high frequencies the absolute value of this current gain falls at a rate of 6 db/octave. Thomas and Moll [1] have calculated the common-emitter current gain for exponentially-graded drift transistors. However, they considered only the intrinsic transistor; *i.e.*, the effects of the space-charge capacitances were ignored. Das and Boothroyd [2] have included the emitter space-charge capacitance, but ignored the collector capacitance. Both of the above treatments are based on the complicated complex expression derived for the common-base current gain by Krömer [3] for an exponential gradient. The common-emitter cutoff frequency shows a maximum at a critical emitter current; so far this current has not been calculated.

The argument presented here shows that the cutoff frequency can be obtained for any impurity profile in a simple manner, including the space-charge capacitances. The bias dependence will be discussed, and the critical current will be given for an exponential impurity distribution.

Calculation of the Cutoff Frequencies

Let us begin with the common-base admittance parameters in general terms. Assume that all four admittances (y_{11B}, y_{21B}, y_{12B} and y_{22B}) are known. Then we also have the common-emitter parameters, since:

$$\left.\begin{aligned} y_{11E} &= y_{11B} + y_{21B} + y_{12B} + y_{22B} \\ y_{21E} &= -y_{21B} - y_{22B} \\ y_{12E} &= y_{12B} - y_{22B} \\ y_{22E} &= y_{22B} \end{aligned}\right\}. \tag{1}$$

The relations in (1) are obtained by relating the common-base and common-emitter differential operators.

Let us consider the common-base admittances involved in (1). Fig. 1 shows the intrinsic transistor plus the space-charge capacitances and base resistance. We should now note that in the determination of β, the *base resistance plays no role.* Because of this fact, let us ignore the base resistance when starting with the common-base admittances. Using asterisks to denote the admittances of the intrinsic transistor, we can write the actual common-base admittances as:

$$\left.\begin{aligned} y_{11B} &= \overset{*}{y}_{11B} + j\omega C_E \\ y_{21B} &= \overset{*}{y}_{21B} \\ y_{12B} &= \overset{*}{y}_{12B} \\ y_{22B} &= \overset{*}{y}_{22B} + j\omega C_c \end{aligned}\right\}. \tag{2}$$

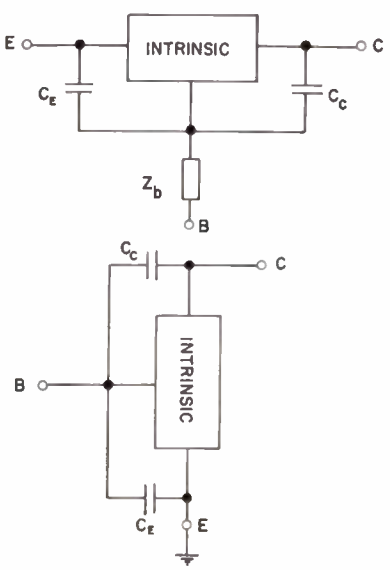


Fig. 1—The equivalent circuit contains the intrinsic transistor with the space-charge capacitances and base resistance/or impedance. When the transistor is rotated to common-emitter, the base impedance appears in series with the current generator applied for determination of beta.

The capacitances only change the input and output admittances. Let us next write the common-emitter current gain from (1) and (2) as:

$$\beta = \frac{y_{21E}}{y_{11E}} = -\frac{\overset{*}{y}_{21B} + \overset{*}{y}_{22B} + j\omega C_c}{\overset{*}{y}_{11B} + j\omega C_E + \overset{*}{y}_{21B} + \overset{*}{y}_{12B} + \overset{*}{y}_{22B} + j\omega C_c}. \tag{3}$$

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Now, consider that y^*_{12B} and y^*_{22B} are always much smaller than either y^*_{11B} or y^*_{21B}. This is due to the fact that the former are always proportional to dW/dV_c (the basewidth modulation by the collector voltage), which is small. Since they are additive in y_{11E} and y_{21E}, we can write β as:

$$\beta = -\frac{y^*_{21B} + j\omega C_c}{y^*_{11B} + y^*_{21B} + j\omega(C_E + C_c)}. \tag{4}$$

We can now see that if C_E and C_c are zero, $\beta=\alpha/(1-\alpha)$. However, if we had started from this as an assumption, it would have been impossible to obtain (4) with both space-charge capacitances included. $\beta=\alpha/(1-\alpha)$ is not a general relation.

According to (4), only two intrinsic admittances need to be known. If we denote the transit time of minority carriers across the base by τ_B, the admittances are very simple for frequencies much less than $1/\tau_B$. Let us write y^*_{11B} as:

$$y^*_{11B} = \frac{qI_E}{kT} + j\omega\frac{qQ_e}{kT}, \tag{5}$$

where I_E is the dc emitter current and kT/q is the thermal voltage. The real part is the well-known forward conductance, and Q_e is the part of the stored charge reclaimable through the emitter. We will not need the value of Q_e; all we need to know is that $Q_e<Q_s$, where Q_s is the stored charge.

The forward transfer admittance y^*_{21B} is defined under the same boundary condition:

$$y^*_{21B} = -\alpha_0\frac{qI_E}{kT} + j\omega\frac{q(Q_s - Q_e)}{kT}, \tag{6}$$

where α_0 is the LF common-base current gain. The real part of the collector current is $\alpha_0 I_E$, and it is negative since it opposes the four-pole current direction.[1] The imaginary part must contain Q_s-Q_e, because Q_e is the charge involved in the emitter current, and the remaining charge must be part of the collector current.

Instead of using the charges in (5) and (6), let us use two transit times. As is shown in Appendix I, the base transit time is $\tau_B=Q_s/\alpha_0 I_E$. In similar fashion, let us arbitrarily define $\tau_R=Q_e/\alpha_0 I_E$. All we need to know is that $\tau_R<\tau_B$, which holds since Q_e is a portion of Q_s. Making use of these two time constants, plus (5) and (6), we can write (4) as:

$$\beta = \frac{\alpha_0 - j\omega(\alpha_0\tau_B - \alpha_0\tau_R + C_c kT/qI_E)}{1-\alpha_0 + j\omega[\alpha_0\tau_B + (C_E + C_c)kT/qI_E]}. \tag{7}$$

It is more convenient to think in terms of β_0 (the LF value of β) instead of α_0. We may say with sufficient accuracy that $\beta_0=\alpha_0/(1-\alpha_0)$. Then, we can write the absolute value of (7) as:

$$\frac{|\beta|}{\beta_0} = \left[\frac{1+\omega^2[\tau_B-\tau_R+C_c(1+\beta_0)/\beta_0 kT/qI_E]^2}{1+\omega^2\beta_0^2[\tau_B+(C_E+C_c)(1+\beta_0)/\beta_0 kT/qI_E]^2}\right]^{1/2}. \tag{8}$$

This expression can be greatly simplified. Note that the frequency-dependent term in the numerator becomes significant only when ω is on the order of $1/\tau_B$. On the other hand, the frequency dependence of the denominator becomes important when the frequency is on the order of $1/\beta_0\tau_B$. This is the term which is responsible for the frequency dependence of $|\beta|$. Also, since $\beta_0\gg 1$, we can write $(1+\beta_0)/\beta_0\simeq 1$. We can write (8) as:

$$\frac{|\beta|}{\beta_0} = \frac{1}{\sqrt{1+\omega^2\beta_0^2[\tau_B + kT/qI_E(C_E+C_c)]^2}}. \tag{9}$$

In the next few steps, we will calculate the cutoff frequencies and show that (9) is sufficiently accurate for describing the frequency dependence of $|\beta|$. Also, we were justified in making our arbitrary definition of τ_R since it appeared in the negligible term.

Let us first calculate the 3-db cutoff frequency. Then, $|\beta|/\beta_0=1/\sqrt{2}$. From (9) we obtain:

$$\frac{1}{\omega_{c\beta}} = \beta_0\left[\tau_B + \frac{kT}{qI_E}(C_E + C_c)\right]. \tag{10}$$

The 3-db cutoff frequency is determined by three time constants. Apparently $\omega_{c\beta}$ is not at all comparable to $1/\tau_B$, and (10) is therefore very accurate.

The general expression in (9) results in a 6-db/octave fall for frequencies much higher than $\omega_{c\beta}$. In order to see the validity of (9) above $\omega_{c\beta}$, let us plot $|\beta|/\beta_0$ with $C_E=C_c=0$, as shown in Fig. 2. We can see from the plot that at $\omega=5\omega_{c\beta}$, the 6-db/octave fall is reached, as expected for such transitions. This frequency is still much smaller than $1/\tau_B$. Therefore, (9) is still accurate and gives the same frequency dependence as would the general solution for all frequencies. The conditions for validity are further eased if we consider that C_E+C_c is never negligible.

Along the 6-db/octave slope, the product of a $|\beta|$ and its corresponding frequency is a constant; this is the so-called "gain-bandwidth product." The constant has the dimension of a frequency, and is called f_T or ω_T. Apparently, this frequency is attained when $|\beta|=1$. Imposing this condition on (9), we find:

$$\frac{1}{\omega_T} = \tau_B + \frac{kT}{qI_E}(C_E + C_c). \tag{11}$$

This relation is extremely simple; yet it accurately supplies the gain-bandwidth product. We find another

[1] This is true for *p-n-p* transistors. However, a similar answer is obtained when *n-p-n* transistors are considered.

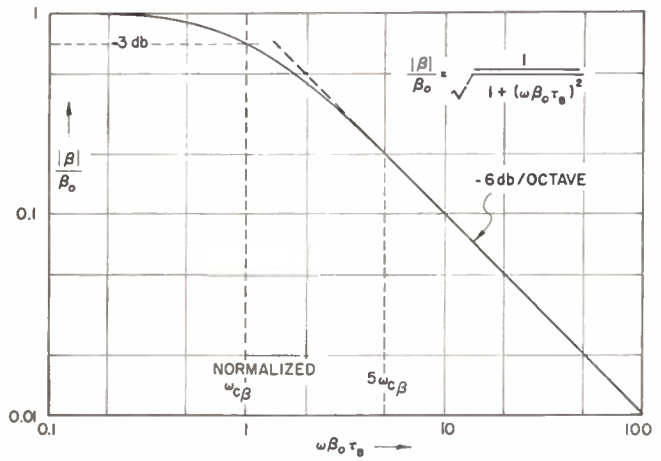


Fig. 2—Normalized common-emitter current gain as a function of frequency. Even with $C_E = C_c = 0$, the 6-db/octave fall is soon reached.

interesting relationship by comparing (10) and (11):

$$\omega_T = \beta_0\omega_{c\beta}. \tag{12}$$

The fact that Q_e had no effect on $\omega_{c\beta}$ or ω_T opened the way to finding these cutoff frequencies in general. According to Appendix II, τ_B can be expressed in general form as:

$$\tau_B = \frac{1}{D}\int_0^W \frac{\int_x^W N dx}{N}\, dx, \tag{13}$$

where N is the impurity distribution in the base from the emitter space-charge boundary to the collector space-charge boundary, and D is the diffusion constant of minority carriers.

Knowing (11) and (13), one can predict the gain-bandwidth product of *any* transistor.[2]

[2] Throughout the presentation, we have purposely ignored some minor time constants in order to retain a simple and continuous train of argument. Let us now consider them for the sake of completeness.

a) In our treatment, we have assumed that collected minority carriers instantaneously traverse the collector space-charge layer. However, the carriers will move at some finite speed [6] dictated by scattering and not by the field strength. This velocity is in the order of 10^7 cm/sec. Therefore, the space-charge transit time for a 0.1-mil layer is about 0.02 nsec. We see from Fig. 4 that τ_B is in the range of 0.6 nsec for today's high-speed transistors.

We should consider that the minority carriers must traverse the collector space-charge immediately after crossing the base region. Therefore, the transit times will be additive, as pointed out by Early, *et al.* [7]. From this, one can see that we only introduce a 3 per cent error by neglecting the space-charge transit time in our above example.

If, for some specific reason, the space-charge transit time should assume a significant magnitude, it can be taken into account by adding it to the base transit time:

$$\tau_B + \tau_B^* + \frac{\text{space-charge width}}{V_{\max}}.$$

b) We have also assumed that the collector bulk resistance is negligible. This assumption is obviously justified for alloyed and surface-barrier collectors, or other structures where the collector bulk is replaced by metal (ECDC). Some diffused collectors may have an appreciable bulk resistance which forms another time constant, r_cC_c. However, such transistors have other very serious weaknesses; *e.g.*, the base widens for high-current densities and the saturation voltage is high. Introduction of epitaxially grown layers has reduced the r_cC_c time constant to the order of 0.001 nsec, which is usually negligible.

Bias Dependence

Because of the simple result in (11), it is easy to predict the collector voltage dependence of ω_T. If a constant current is maintained, we would expect a dependence of ω_T on the voltage as shown by measurement of a drift transistor in Fig. 3 (next page). For very low collector voltages, C_c is very large for abrupt or nearly abrupt junctions because the collector space charge is compressed. Not only does the low voltage maintain a narrow space-charge layer, but it is made even narrower by the minority carrier current flowing across the collector space-charge layer. As the collector voltage is increased, the collector capacitance quickly assumes a relatively small value. Thereafter, τ_B decreases relatively slowly in drift transistors and ω_T slowly approaches qI_E/kTC_E.

Let us now consider the current dependence. This dependence is apparently more important, since the voltage dependence is quite mild for reasonable operating biases. If we differentiate (11) with respect to $1/I_E$, we find:

$$\frac{d1/\omega_T}{d1/I_E} = \frac{kT}{q}(C_E + C_c). \tag{14}$$

For low currents, a plot of $1/\omega_T$ vs $1/I_E$ will be a straight line. In this region, the current dependence of the gain-bandwidth product is due solely to the space-charge capacitances. As the current becomes large, the large number of minority carriers in the base will "wash out" the built-in field in drift transistors. Then, τ_B will increase while the second term in (11) will continue to decrease. These two effects result in a maximum of ω_T at a particular current. A measurement demonstrating these phenomena is given in Fig. 4.

It is desirable to know the critical current where the maximum occurs, since such a point represents the maximum gain-bandwidth product a given transistor can attain. This critical current cannot be calculated in a general form. However, for specific impurity functions, a solution can be computed. The authors have done this [4] for an exponential impurity distribution. The results of these computations can be applied directly to the present problem, with the following assumptions:

1) $N = N_E \exp(-x/x_0) = N_E \exp(-\eta x/W)$, where η was introduced for a convenient measure of the built-in field.
   $\eta = qWE_b/kT = W/x_0 = \ln N_E/N_W$, where E_b is the built-in field for low currents. N_E and N_W are the respective impurity concentrations at the emitter and collector space-charge boundaries.
2) C_c is either constant with respect to the current or negligible in value.
3) The basewidth W is constant with respect to the current.

Most transistors' impurity distributions can be approximated well by an exponential. The other assumptions can be satisfied, except for low collector voltages, if the doping adjacent to the collector junction is high

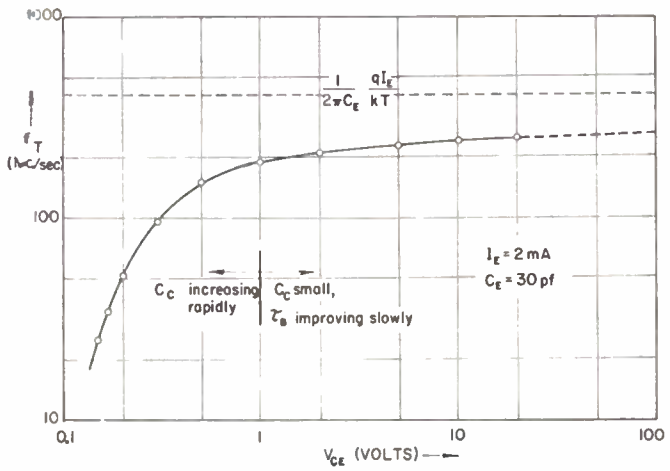


Fig. 3—Dependence of the gain-bandwidth product on collector voltage.

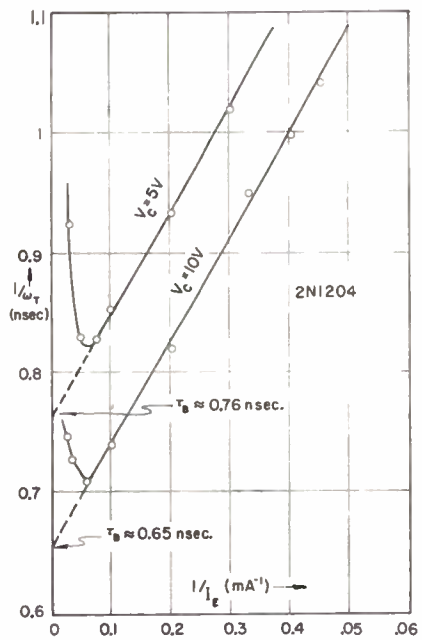


Fig. 4—The current dependence of the gain-bandwidth product in terms of angular frequency. The extrapolation then yields τ_B.

enough so that the space-charge boundary does not recede significantly [5].

With the above assumptions satisfied, the critical current at which the maximum in ω_T occurs is:

$$I_{cr} = \frac{N_E q A D}{W} GH\left(1 + \sqrt{1 + \frac{2}{GH^2}}\right) \qquad (15)$$

where

$$G = \frac{kTC_E}{q^2 N_E A W} \frac{\eta^3\sqrt{2}}{\sinh \eta - 8}$$

and

$$H = 4 + \frac{\exp 0.85\eta}{30}.$$

For $kTC_E/q^2N_E A W \leq 10^{-3}$, (15) simplifies to:

$$I_{cr} = \frac{N_E q A D}{W}\sqrt{2G}. \qquad (16)$$

Applying (15) to the example given in Fig. 4, we obtain satisfactory agreement with the critical current measured.

Since ω_T breaks away from the low-current slope quite sharply (as can be seen in Fig. 4), the maximum ω_T can be found by inserting the critical current into (11). The same critical current applies to the 3-db cutoff, $\omega_{c\beta}$.

Finally, it can be mentioned that there is a simple interrelation between the diffusion capacitance and the gain-bandwidth product, which is shown in Appendix III.

Conclusions

Calculation of β from $\alpha/(1-\alpha)$ is not correct, since this equation is not exact. By application of four-pole theory, it was shown that the gain-bandwidth product is a simple function of the base transit time and the space-charge capacitances. The solution is general, since the transit time can be calculated in general. It follows that any transistor's common-emitter current gain will follow a 6-db/octave fall. The 3-db cutoff and the gain-bandwidth product are interrelated by the LF gain, β_0.

$1/\omega_T$ is a linear function of $1/I_E$ for low currents. Plotting the current dependence of the gain-bandwidth product in this fashion is very convenient. The slope of such a plot supplies the sum of the two space-charge capacitances, and extrapolation to infinite current yields τ_B. In diffusion transistors, ω_T continuously improves with increasing current; in drift transistors it reaches a maximum. This critical current was given for an exponential impurity distribution. When the collector space charge varies with the current, a smaller critical current is obtained, lowering $\omega_{T\max}$.

It is interesting to note what would happen to the gain-bandwidth product if all dimensions of a transistor were reduced proportionately. The collector space-charge width can be reduced by lowering the applied voltage. The emitter space-charge width remains fairly constant. When C_e is neglected, both terms in (11) shrink with the square of the size for low currents

$$1/\omega_T = \text{const}_1 \text{ size}^2.$$

However, if we observe the maximum performance, we must consider that I_{cr} is proportional to the square root of the size. Therefore,

$$1/\omega_{T\max} = \text{const}_2 \text{ size}^2 + \text{const}_3 \text{ size}^{3/2}.$$

Thus, the emitter capacitance assumes increasing importance as the size is reduced.

Appendix I

Let us first calculate the transit time in terms of the minority carriers' velocity. The time needed to cross an infinitesimally small distance dx is

$$d\tau = \frac{dx}{V}, \tag{17}$$

where V is the velocity within dx. To obtain the transit time, we integrate from emitter to collector

$$\tau_B = \int_E^C \frac{1}{V}\,dx. \tag{18}$$

In order to obtain the velocity, consider for example the hole current in a p-n-p transistor:

$$I_p = -qAD\not{p}' + qA\mu E_b\not{p}, \tag{19}$$

where $\not{p}$ is the hole concentration. If we divide by $qA\not{p}$, we obtain

$$\frac{I_p}{qA\not{p}} = \text{velocity due to diffusion} + \text{velocity due to drift} = V. \tag{20}$$

Using (20) in (18), we obtain:

$$\tau_B = \int_E^C \frac{1}{V}\,dx = \frac{\int_E^C qA\not{p}\,dx}{I_p} = \frac{Q_s}{I_p} = \frac{Q_s}{\alpha_0 I_E}. \tag{21}$$

Appendix II

By requiring that the majority current is zero, and space-charge neutrality exists in the base region, the electric field is obtained as

$$E_b = -\frac{kT}{q}\frac{N'}{N}, \tag{22}$$

where N is the impurity distribution. Making use of (22) and the Einstein relation, the hole current may be rewritten as

$$I_p = -qAD\left(\not{p}' + \frac{N'}{N}\not{p}\right). \tag{23}$$

This is a linear differential equation for the hole concentration $\not{p}$ and has the solution

$$\not{p} = \frac{I_p}{qAD}\frac{\int_x^W N\,dx}{N}, \tag{24}$$

with the boundary condition $\not{p}=0$ at $x=W$. The stored charge is obtained by integrating (24) from $x=0$ to $x=W$:

$$Q_s = qA\int_0^W \not{p}\,dx = \frac{I_p}{D}\int_0^W \frac{\int_x^W N\,dx}{N}\,dx. \tag{25}$$

Since Q_s and τ_B are interrelated in (21), we may write:

$$\tau_B = \frac{1}{D}\int_0^W \frac{\int_x^W N\,dx}{N}\,dx. \tag{26}$$

Similar results are obtained when an n-p-n transistor is calculated. Eq. (26) holds only for currents less than I_{cr}.

Appendix III

The input admittance in common emitter can be obtained from (1) by making use of (2), (5), and (6):

$$y_{11E} \simeq y_{11B} + y_{21B} + j\omega C_c = \frac{qI_E}{kT}(1-\alpha_0) + j\omega\left(\frac{qQ_s}{kT} + C_E + C_c\right). \tag{27}$$

The capacitive portion is a diffusion capacitance qQ_s/kT plus the space-charge capacitances

$$C_m = \frac{qQ_s}{kT} + C_E + C_c = \alpha_0\frac{qI_E}{kT}\tau_B + C_E + C_c. \tag{28}$$

Dividing C_m by qI_E/kT, we obtain:

$$C_m\frac{kT}{qI_E} = \alpha_0\tau_B + \frac{kT}{qI_E}(C_E + C_c) \simeq 1/\omega_T. \tag{29}$$

The gain-bandwidth product can be obtained also by measuring the input capacitance at a specific current. For higher currents, however, this capacitance measurement is disturbed by the base resistance.

References

[1] D. Thomas and J. Moll, "Junction transistor short-circuit current gain and phase determination," Proc. IRE, vol. 46, pp. 1177–1184; June, 1958.

[2] M. Das and A. Boothroyd, "On the frequency dependence of the magnitude of common-emitter current gain of graded-base transistors," Proc. IRE (Correspondence), vol. 48, pp. 240–241; February, 1960.

[3] H. Krömer, "Zur der Theorie des Diffusion und des Drifttransistors," *Arch. elekt. Übertragung*, vol. 8, pp. 223–228; May, 1954.

[4] J. Lindmayer and C. Wrigley, "The high-injection-level operation of drift transistors," *Solid-State Electronics*, vol. 2, pp. 79–84; March, 1961.

[5] C. T. Kirk, "A Theory of Transistor Cut-Off Frequency Falloff at High Current Densities," M.I.T. Lincoln Lab., Lexington, Mass., Rept. No. 53G-0056; May 3, 1961.

[6] J. M. Early, "PNIP and NPIN junction transistor triodes," *Bell Sys. Tech. J.*, vol. 33, pp. 517–533; May, 1954.

[7] J. M. Early, *et al.*, "Engineering Services on Transistors," Bell Telephone Labs., Interim Tech. Rept. on Contract DA-36-039-sc-64618; July 1, 1956.

Correspondence

On the Axial Echo Area of the Cone-Sphere Shape*

The purpose of this correspondence is to report some initial results of a study of the backscattering from a cone-sphere. The target shape is the same as that considered by Keller in a recent paper,[1] where the radius and its first derivative are matched at the cone-sphere junction. In the present analysis, the axial echo area of the cone-sphere is derived by considering the time response of the target to an impulsive excitation.[2] In the article by Kennaugh and Cosgriff,[2] the low-frequency-derived conditions on the impulse response time function $f(t)$ are

$$\int_0^\infty f(t)dt = 0, \tag{1}$$

$$\int_0^\infty tf(t)dt = 0, \tag{2}$$

$$\int_0^\infty t^2 f(t)dt = K, \tag{3}$$

where $f(t)$ is the impulse response time function, t is the time, and K is proportional to the Rayleigh coefficient, *i.e.*, to the coefficient of ω^2 in the low-frequency expansion of the scattered field. These moment conditions were used by Kennaugh and Cosgriff[2] to correct the impulse response as predicted by physical optics by appending corrective responses. In the present case, the exact short pulse time response of a conducting sphere was known,[3,4] and as a first approximation the impulse response of the cone as predicted by physical optics was combined with these exact sphere calculations, assuming no interaction at the cone-sphere junction. The moment conditions in (1) and (2) were found to be satisfied without additional correction. The moment condition in (3) could not be imposed, since the Rayleigh coefficient of the cone-sphere shape is not known. As a first approximation then the axial-echo area of the cone-sphere (sphere radius R) was written as

$$\sigma = \pi R^2 \left| E_{\text{p.o.}}^{\text{cone}} + E_{\text{exact}}^{\text{sphere}} - E_{\text{p.o.}}^{\text{sphere}} \right|^2, \tag{4}$$

where $E_{\text{p.o.}}^{\text{cone}}$ is the normalized back-scattered field derived from the impulse response time function of the cone as predicted by physical optics, $E_{\text{exact}}^{\text{sphere}}$ is the normalized back-scattered field derived from the exact short pulse time response of the conducting sphere, and $E_{\text{p.o.}}^{\text{sphere}}$ is the normalized back-scattered field derived from the impulse response time function of that portion of the sphere blocked by the cone as predicted by physical optics. It was necessary to subtract the contribution from the blocked portion of the sphere since an analytical expression for the exact short pulse response of the sphere was not available.

In Fig. 1, the measured and calculated normalized axial echo areas of a 30° cone-sphere as a function of the sphere circumference in wavelengths are compared. Included in the figure are the echoing areas predicted by physical optics and by (35) of Keller.[1] It is noted that both these latter theories are seriously in error in the Rayleigh and low-resonance regions.

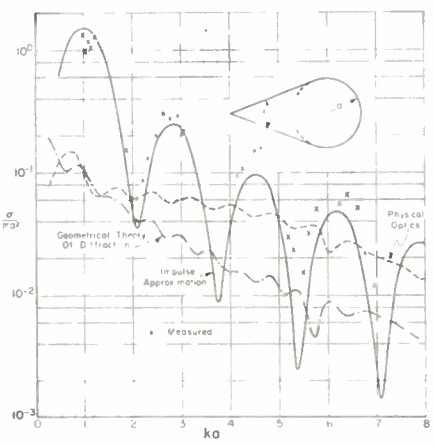


Fig. 1—Measured and calculated normalized axial echo area of a 30° cone-sphere as a function of the sphere circumference in wavelengths.

It is apparent from the impulse analysis described above that the axial echo area from the cone-sphere has two major contributions, the creeping wave around the back or shadowed portion of the sphere and the contribution from the discontinuity in the second derivative at the cone-sphere junction. Physical optics fails to predict the creeping wave, Keller's analysis neglected the junction discontinuity; as a consequence, both solutions are in error for low frequencies. Stated another way, both the above methods fail to satisfy the first moment condition in (2). It should be noted that a rigorous application of the geometrical theory of diffraction[5] does predict a diffracted ray at discontinuities in the second derivative. An analytical expression for the diffraction coefficient is difficult to obtain however. One further observation is made; according to the impulse analysis, within reasonable limits the cone angle has little effect on the magnitude of the echo area. Thus, for example, the axial echo area of a 15° cone-sphere would differ little from that shown in Fig. 1 except for the location of the maxima and minima due to the shift in the location of the cone-sphere junction. The measured data were recorded by R. W. St Clair using 6 cone-sphere targets at X-band frequencies. The experimental points in Fig. 1 are an average of several readings and are believed accurate to within ±0.5 db. A complete discussion of the analysis and results briefly described here will be included in a forthcoming paper.

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High-Frequency Available Power Gain of Diffused-Base Transistors*

The purpose of this note is to give the expression for the high-frequency available power gain of diffused-base transistors, having a drifting-field in the base region, used in the common-emitter configuration.

Throughout these calculations, the following assumptions are made:

1) The T-type equivalent circuit shown in Fig. 1 is adopted. The value of α is determined by the formula derived by Thomas and Moll.[1] Parasitic elements such as interterminal capacitances are neglected. Some of those, such as capacitances bridged across input or output terminals, may be matched out and will have no contribution to the available power gain. However, in general, parasitic elements affect available power gain, and should be considered for actual transistors. The collector-base diffusion capacitance and other elements associated with Early effect are not included in Fig. 1. Emitter and collector body resistances are neglected, hence, the frequency cutoff asso-

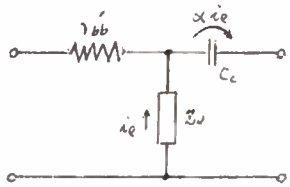


Fig. 1—Transistor equivalent circuit used.

* Received by the IRE, October 23, 1961. The research reported in this note was supported in part by the Aeronautical Systems Div., AF Systems Command, U. S. Air Force, Wright-Patterson AFB, Ohio.

[1] J. B. Keller, "Backscattering from a finite cone," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 175–182; March, 1960.

[2] E. M. Kennaugh and R. L. Cosgriff, "The use of impulse response in electromagnetic scattering problems," 1958 IRE National Convention Record, pt. 1, pp. 72–77.

[3] E. M. Kennaugh, "The scattering of short electromagnetic pulses by a conducting sphere," Proc. IRE (Correspondence), vol. 49, p. 380; January, 1961.

[4] Final Engineering Report, Antenna Lab., The Ohio State Univ. Res. Foundation, Columbus, Ohio, Rept. No. 1073-4, prepared under Contract AF 19(604)-6157, AF Cambridge Res. Labs., Office of Aerospace Res., Bedford, Mass.; January 1, 1961.

[5] J. B. Keller, "A geometrical theory of diffraction," *Proc. Symp. on Appl. Math.*, McGraw-Hill Book Co., New York, N. Y., vol. 8, pp. 27–52; 1958.

* Received by the IRE, October 17, 1961.

[1] D. E. Thomas and J. L. Moll, "Junction transistor short-circuit current gain and phase determination," Proc. IRE, vol. 46, pp. 1177–1184; June, 1958.

ciated with these resistances and circuit capacitances are not considered.

2) Low-frequency common-base short-circuit current gain α_{b0} is unity.

3) The feedback effect due to C_c and Z_e is neglected, and the common-emitter input impedance is dominantly determined by r_{bb}'. Usually, the emitter dc current will be large enough to satisfy this condition.

4) Emitter cutoff due to the time constant of Z_e is not considered.

Although these assumptions will not be valid for some actual transistors, the result obtained should give a figure of merit of a high-frequency transistor.

The H parameters of the common-emitter transistor circuit as shown in Fig. 1, can be calculated with the above assumptions, and is found to be,

$$h_{11e} = r_{bb}' \tag{1}$$

$$h_{12e} = 0 \tag{2}$$

$$h_{21e} = \frac{\alpha}{1-\alpha} \tag{3}$$

$$h_{22e} = \frac{j\omega C_c}{1-\alpha}. \tag{4}$$

According to Thomas and Moll, α may be expressed as

$$\alpha = \alpha_{b0} \frac{e^{j(K-1/K)f/f_\alpha}}{1+j\frac{f}{f_\alpha}}, \tag{5}$$

where K is the factor associated with drifting field of transistors, and related to the term m, the excess phase, as

$$m = \frac{1-K}{K}. \tag{6}$$

f_α is the frequency where the magnitude of α is 3 db below α_{b0}, and α_{b0} is the low-frequency common-base short-circuit current gain and assumed here as unity.

If

$$\frac{1-K}{K}\frac{f}{f_\alpha} \ll 1,$$

and α_{b0} is unity, α may be expressed as

$$\alpha = \frac{1-j\frac{1-K}{K}\frac{f}{f_\alpha}}{1+j\frac{f}{f_\alpha}}, \tag{7}$$

and

$$\frac{1}{1-\alpha} = \frac{1+j\frac{f}{f_\alpha}}{j\frac{1}{K}\frac{f}{f_\alpha}}, \tag{8}$$

then,

$$\frac{\alpha}{1-\alpha} = \frac{e^{j(K-1/K)f/f_\alpha}}{j\frac{1}{K}\frac{f}{f_\alpha}}. \tag{9}$$

Since h_{12e} is assumed as zero, the available power gain of a transistor having the above h parameters will be given by

$$G_{av} = \frac{|h_{21e}|^2}{4\,\mathrm{Re}(h_{11e})\,\mathrm{Re}(h_{22e})}, \tag{10}$$

where $\mathrm{Re}(h_{11e})$ and $\mathrm{Re}(h_{22e})$ represent the real parts of h_{11e} and h_{22e}, respectively.

Using (1)–(10), G_{av} is calculated as

$$G_{av} = \frac{Kf}{8\pi f^2 v_{bb}' C_c}. \tag{11}$$

As can be seen from (9), Kf_α is the frequency when

$$\left|\frac{\alpha}{1-\alpha}\right| = 1, \tag{12}$$

and if we express this frequency as f_T, G_{av} may be represented as

$$G_{av} = \frac{f_T}{8\pi f^2 v_{bb}' C_c}. \tag{13}$$

This last expression is very similar to that obtained by Pritchard,[2] except that f_α in his expression is replaced by f_T. His expression is valid when the minority carrier in base region moves only by diffusion, *i.e.*, the case $m=0.2$. The results obtained here can be applied to transistors having arbitrary value of excess phase m.

Figure of merit, or gain-bandwidth product, of high-frequency transistors can be written as follows:

$$\sqrt{G_{av}} f = \left(\frac{f_T}{8\pi v_{bb}' C_c}\right)^{1/2}. \tag{14}$$

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[2] R. L. Pritchard, "High-frequency power gain of junction transistors," Proc. IRE, vol. 43, pp. 1075–1085; September, 1955.

On the Extended Definition of Linearity*

Experience in the use of the extended definition of linearity which was recently published[1] has revealed several respects in which the definition can be improved. A reformulated version of the definition which differs in detail but not in substance from the original is presented below.

Preliminary Definitions

A. Definitions of Complete Characterization, State, Input and Output

A (nonrandom)[2] system B is *completely characterized with respect to input u, output y, and state s if:*

1) B is associated with three time-functions denoted by u, y and s, with $u(t)$, $y(t)$ and $s(t)$ representing the values of u, y and s at time t. The union of ranges of $u(t)$ for all values of t is denoted by U, with the symbols Y and S having similar meaning in relation to $y(t)$ and $s(t)$. (In effect, U, Y and S represent, respectively, the totality of values which $u(t)$, $y(t)$ and $s(t)$ can assume as t varies from $-\infty$ to ∞. Unless indicated to the contrary, $u(t)$, $y(t)$ and $s(t)$ will be assumed to be vectors. A particular value of $s(t)$ will be denoted by q.)

2) The functions u, y and s have the following properties:

a) u is *controllable*, *i.e.*, the values of $u(t)$ can be chosen at will from U.
b) y is *observable*, *i.e.*, the values of $y(t)$ can be observed (measured).
c) s is *resettable*, *i.e.*, the values of s can be changed (reset) externally at arbitrary times t_0, t_0', t_0'', $\cdots$. However, s is not controllable during the intervals (t_0, t_0'), (t_1, t_0''), $\cdots$, and it may or may not be observable.

3) Given the value of s at some initial time t_0, the output at any subsequent time t_1 (before s is reset) can be expressed as a function of $s(t_0)$ and the values of the input between the instants t_0 and t_1. Symbolically

$$y(t_1) = B[s(t_0);\, u_{[t_0,t_1]},\, t_0,\, t_1] \tag{1}$$

for all t_0, t_1, $s(t_0)$ and $u_{[t_0,t]}$, where $u_{[t_0,t_1]}$ denotes the segment of the function u over the closed interval $[t_0, t_1]$, and B is a function of the arguments indicated. In effect, (1) states that the value of the output at any time t_1 is completely determined by the value of s at some initial time t_0 and the values of the input between t_0 and t_1.

If 1)–3) hold, then u, y and s will be said to *qualify* as the *input*, *output* and *state functions* of B, respectively, and B will be said to be *completely characterized*. It is easy to deduce that (1) implies and is implied by the following two relations:

$$s(t_1) = f[s(t_0);\, u_{[t_0,t_1]},\, t_0,\, t_1] \tag{2}$$

$$y(t_1) = g[s(t_1),\, u(t_1),\, t_1] \tag{3}$$

in which f and g are single-valued functions of their arguments. Note that (2) is consistent with 2)-c).

B. Definition of Zero (Neutral, Null) State

A state θ is a *zero state* of B if the response of B to zero input with B initially in state θ is identically zero, *i.e.*: $B[\theta; 0] \equiv 0$. ($B[q; u]$ is abbreviated notation for the response of B to u with B initially in state q.) In general, a system may have no zero state or it may have one zero state or it may have two or more nonequivalent zero states.

In the previous note, linearity was defined in terms of the ground state of B. In most physical systems, the ground state and the zero state are coincident. There are some simple systems, however, (*e.g.*, LC networks) which do not have a ground state in the sense defined in the previous note. For this reason, it is preferable to define linearity in terms of the zero state rather than the ground state. Such a definition is given below.

* Received by the IRE, October 16, 1961. This work was supported in part by the National Science Foundation.

[1] L. A. Zadeh, "An extended definition of linearity," Proc. IRE (Correspondence), vol. 49, pp. 1452–1453; September, 1961.

[2] The restriction to nonrandom systems does not play an essential role. The definition in question can be extended to stochastic (probabilistic) systems in a straightforward fashion.

C. Definition of Linearity

A system B is *linear* if

1) B has at least one zero state.

2) If B has just one zero state θ, then the following relation holds for all states q, all real constants k, all $u_{[t_0,t_1]}$ and $v_{[t_0,t_1]}$ and all t_0, t_1 (with $t_1 \geq t_0$):

$$k[B[q; u_{[t_0,t_1]}, t_0, t_1] - B[q; v_{[t_0,t_1]}, t_0, t_1]] = B[\theta; k(u_{[t_0,t_1]} - v_{[t_0,t_1]}), t_0, t_1]. \quad (4)$$

3) If B has two or more nonequivalent zero states θ_i, $i = 1, 2, \cdots$, then in (4) θ is replaced by $\theta_{i(q)}$, meaning that to each state q there corresponds a zero state $\theta_{i(q)}$ depending on q such that (4) holds for all q, all real constants k, all $u_{[t_0,t_1]}$ and $v_{[t_0,t_1]}$ and all t_0, t_1 (with $t_1 \geq t_0$).

Comment

In most cases, B has just one zero state and it is not necessary to consider case 3). An example of a system in which 3) rather than 2) has to be used is a switchable linear network, that is, an RLC network containing one or more switches whose operation is independent of the voltages and currents in the network.

Concluding Remark

The main difference between (4) and the definition given earlier lies in the replacement of the ground state with the zero state. Usually, but not always, the ground state is identical with the zero state, so that the two definitions are equivalent. In any case, the consequences of extended linearity which were established previously are valid regardless of whether the ground state or the zero state is used in the definition.

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Simulation of Third-Order Systems with One Operational Amplifier*

In analog computation, need often arises for the simulation of third-order systems and this is generally done with the aid of two or more operational amplifiers. The purpose of this note is to indicate a method whereby third-order systems can be simulated with the aid of only one operational amplifier and two-terminal impedances consisting of only resistors and capacitors.

A network for the simulation of third-order systems is shown in Fig. 1, and its transfer function can be shown to be

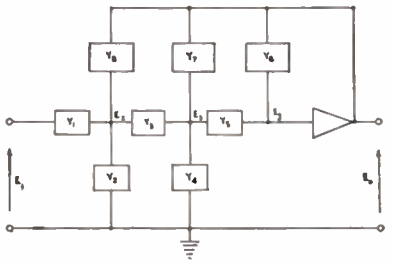
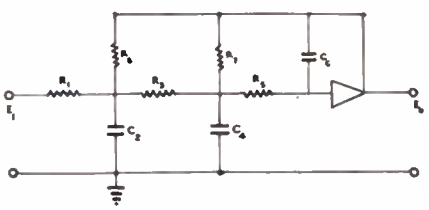
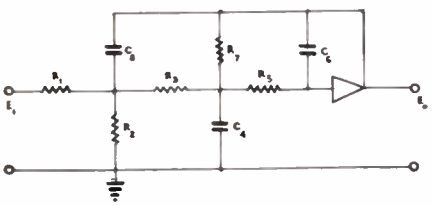


Fig. 1—Block diagram of an operational amplifier and its associated input feedback networks for simulating third-order systems.



(a)



(b)

(a) $b_0 = \dfrac{\alpha^2}{(1+3\alpha)}$,

$$a_1 = \frac{\alpha}{(1+3\alpha)} RC_2 + (1+\alpha)RC_6,$$

$$a_2 = \frac{\alpha(1+2\alpha)}{(1+3\alpha)} R^2C_6(C_2+C_4),$$

$$a_3 = \frac{\alpha^2}{(1+3\alpha)} R^3C_2C_4C_6,$$

$R_1 = R_3 = R_5 = R$ and $R_7 = R_8 = \alpha R$,

$$a_1 > \frac{a_3(1+2\alpha)}{a_2(1+3\alpha)}$$

(b) $b_0 = \dfrac{a_0}{3}$,

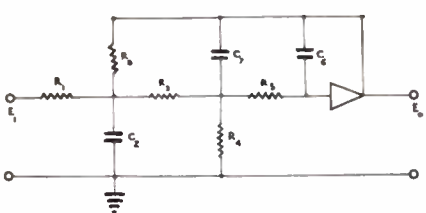
$$a_1 = \tfrac{1}{3}[(5a_0+1)RC_6 + (a_0+1)RC_8],$$

$$a_2 = \tfrac{1}{3}RC_6[3a_0RC_4 + (2a_0+1)RC_8],$$

$$a_3 = \frac{a_0}{3} R^3C_4C_6C_8,$$

$R_1 = R_2 = R_3 = R_5 = R$, $R_7 = a_0R$,

$$a_1 > \frac{a_3}{a_2}(a_0+1).$$



(c)

(c) $b_0 = a_0$,

$$a_1 = [(5a_0+3)RC_6 + (2a_0+1)RC_7],$$

$$a_2 = 3a_0R^2C_2C_6 + a_0R^2C_2C_7 + (2a_0+1)R^2C_6C_7,$$

$$a_3 = a_0R^3C_2C_6C_7,$$

$R_1 = R_3 = R_4 = R_5 = R$ and $R_8 = a_0R$,

$a_1a_2 > a_0a_3$ $\quad J^3 - 27J^2 < 0$

Fig. 2—Networks for the simulation of $\dfrac{E_0}{E_1} = -\dfrac{b_0}{a_3s^3 + a_2s^2 + a_1s + 1}$.

$$\frac{E_0}{E_1} = -\frac{Y_1Y_3Y_5}{Y_6(Y_1+Y_2+Y_3)(Y_3+Y_4+Y_5+Y_7) + Y_3Y_6(Y_4+Y_5+Y_7) + Y_5Y_7(Y_1+Y_2+Y_3+Y_8) + Y_3Y_5Y_8}. \quad (1)$$

* Received by the IRE, August 25, 1961; revised manuscript received, September 8, 1961.

A third-order system is characterised by a transfer function of the form

$$F(s) = -\frac{(b_2s^2 + b_1s + b_0)}{(a_3s^3 + a_2s^2 + a_1s + 1)}. \quad (2)$$

An inspection of (1) will indicate that simulation of the system as characterized by (2) is possible, provided the admittances (Y's) are properly chosen. The discussion in this note, however, will be confined to a case where $b_2=b_1=0$, and hence to systems with transfer functions of the form

$$F(s) = -\frac{b_0}{(a_3s^3 + a_2s^2 + a_1s + 1)}, \quad (3)$$

where the constants b_0, a_1, a_2 and a_3 are positive and real.

The other cases where $b_1 \neq 0$ and $b_2 \neq 0$ will be discussed in separate notes which will be communicated later.

Simulation of the system characterized by (3) with the network of Fig. 1 would require at least three capacitors. A possible choice [see Fig. 2(a)] is to make

$$\left.\begin{aligned} Y_2 &= SC_2 \\ Y_4 &= SC_4 \\ Y_6 &= SC_6 \\ Y_1 = Y_3 = Y_5 &= \frac{1}{R} \end{aligned}\right\}, \quad (4)$$

and

$$Y_7 = Y_8 = \frac{1}{\alpha R}$$

Substituting (4) into (1) and simplifying,

$$\frac{E_0}{E_i} = -\frac{\frac{\alpha^2}{(1+3\alpha)}}{\frac{\alpha^2}{(1+3\alpha)}R^3C_2C_4C_6s^3 + \left\{\frac{\alpha(1+2\alpha)}{(1+3\alpha)}R^2C_6(C_2+C_4)\right\}s^2 + \left\{\frac{\alpha R}{(1+3\alpha)}C_2+(1+\alpha)RC_6\right\}s+1}. \quad (5)$$

Eqs. (3) and (5) will be identical if

$$b_0 = \frac{\alpha^2}{(1+3\alpha)} \quad (6)$$

$$a_1 = \frac{\alpha}{(1+3\alpha)}T_1 + (1+\alpha)T_3 \quad (7)$$

$$a_2 = \frac{\alpha(1+2\alpha)}{(1+3\alpha)}T_3(T_1+T_2) \quad (8)$$

$$a_3 = \frac{\alpha^2}{(1+3\alpha)}T_1T_2T_3, \quad (9)$$

where

$$\left.\begin{aligned} T_1 &= RC_2 \\ T_2 &= RC_4 \\ T_3 &= RC_6 \end{aligned}\right\}. \quad (10)$$

It is now required to determine from (6)–(10) the values of α, T_1, T_2 and T_3, that is, R, C_1, C_2 and C_6, in terms of the known constants b_0, a_1, a_2, and a_3.

Solution of (6) gives

$$\alpha = \frac{3b_0 + \sqrt{9b_0^2 + 4b_0}}{2} \quad (11)$$

as the negative root is inadmissible.

Eliminating T_2 and T_3 from (7)–(9) gives

$$T_1^3 - \frac{a_1(1+3\alpha)}{\alpha}T_1^2 + \frac{a_2(1+\alpha)(1+3\alpha)^2}{\alpha^2(1+2\alpha)}T_1 - \frac{a_3(1+\alpha)(1+3\alpha)^2}{\alpha^3} = 0. \quad (12)$$

Eq. (12) is a cubic and its roots can be easily determined by techniques that are very well known. Now, (12) has at least one real root and this root will be positive if

$$a_1 > \frac{a_3(1+2\alpha)}{a_2(1+3\alpha)}. \quad (13)$$

It can be proved that if the condition of (13) is satisfied, then the corresponding values of T_2 and T_3 will also be real and positive and these can be obtained by substituting the value of T_1 in (7)–(9). Having thus determined α, T_1, T_2, T_3 and choosing a convenient value for any one of the capacitors (say C_2), it is easy to calculate the component values R, C_4, C_6 with the aid of (10).

Three possible circuits and the conditions that each one must satisfy for simulating the system as defined by (3) are shown in Fig. 2.

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Approximation of Tunnel-Diode Characteristics*

The purpose of this letter is to describe an approximation to determine the characteristics of tunnel diodes by the currents and voltages belonging to the peak and valley points of the characteristics (U_M and U_m).

Upon applying a forward voltage, two opposing currents are flowing through the tunnel diode:

1) The field-emission or Zener current is caused by electrons passing from the valence band to the conductance band.

2) The tunnel or Esaki current is caused by electrons tunneling from the conductance band to the valence band.

Let us suppose that the number of quantum states in the conductance band is $N_c(E)$ and that in the valence band $N_v(E)$. The probability of occupancy for a shell of the energy E is given by the $f(E)$ Fermi-Dirac distribution. The Fermi level E_v on the p side is below the top of the valence band, and that on the n side E_c will be located above the bottom of the conductance band. The probability of penetrating the gap by an electron may be assumed to be equal for both directions.

The Zener current I_z depends on the number of the occupied quantum states of energy E in the valence band of the p layer, *i.e.*, on the number of electrons of energy E that is on $N_v \cdot f_v$ and, further, on the number of the empty states of energy E in the conductance band of the n layer, *i.e.*, $N_c(1-f_c)$.

The same considerations are valid for the tunnel current I_t. In this way we obtain the relation deduced by Esaki,

$$I = I_z + I_t = \text{const} \int^{E_x} N_c(E)N_v(E) \cdot [f_c(E) - f_v(E)]dE. \quad (1)$$

When supposing parabolic bands, the number of the quantum states per unit energy are

$$N_v(E) = A\sqrt{E}, \quad (2)$$

$$N_c(E) = A\sqrt{E_x - E}. \quad (3)$$

The Fermi distributions are

$$f_v(E) = \frac{1}{1 + \exp\frac{E - E_v}{kT}} \quad (4)$$

and

$$f_c(E) = \frac{1}{1 + \exp\frac{E + E_c - E_x}{kT}}. \quad (5)$$

It is apparent from the band model (Fig. 1) that the relation between E_x, the Fermi levels and the external voltage is

$$E_x = E_c + E_v - eU. \quad (6)$$

Supposing equal impurities on both sides and introducing a fictitious potential U_m,

$$E_c = E_v = \frac{eU_m}{2}, \quad (7)$$

using the symbol

$$x = 2\frac{E}{E_x} - 1 \quad (8)$$

for normalizing the integration limits, and using (2) and (3), we obtain

$$N_vN_c = \frac{1}{2}A^2 \cdot \frac{e}{kT}(U_m - U)\sqrt{1 - x^2}. \quad (9)$$

* Received by the IRE, October 6, 1961. Work was performed at the Dept. of Vacuum Tubes, Technical University of Budapest, Budapest, Hungary.

From (4) and (5),

$$f_c - f_v = \frac{sh\,\frac{1}{2}\,\frac{eU}{kT}}{ch\left[\frac{1}{2}\,\frac{e}{kT}(U_m - U)x\right] + ch\,\frac{1}{2}\,\frac{eU}{kT}}. \quad (10)$$

Using (1), (9), and (10) we obtain for the equation of the characteristic:

$$I = \text{const}\,\frac{1}{4}\,A^2\left(\frac{e}{kT}\right)^2(U_m - U)^2 sh\,\frac{1}{2}\,\frac{eU}{kT}\int_{-1}^{+1}\frac{\sqrt{1-x^2}}{ch\left[\frac{1}{2}\,\frac{e}{kT}(U_m - U)x\right] + ch\,\frac{1}{2}\,\frac{eU}{kT}}dx. \quad (11)$$

If we integrate in the surrounding of $|x| = 1$, then—in consequence of the decrease of the numerator and the increase of the denominator—this part of the integration will hardly influence the result. Inside the actual integration interval the term with x in the denominator is negligible compared to the constant term. Thus, with good approximation,

$$I = \text{const}\left(\frac{e}{kT}\right)^2(U_m - U)^2 th\,\frac{1}{2}\,\frac{eU}{kT} \cdot \int_{-1}^{+1}\sqrt{1-x^2}\,dx. \quad (12)$$

Therefore, after reducing the constants, the approximated characteristic is

$$I = a(U_m - U)^2 th\,\frac{1}{2}\,\frac{eU}{kT}. \quad (13)$$

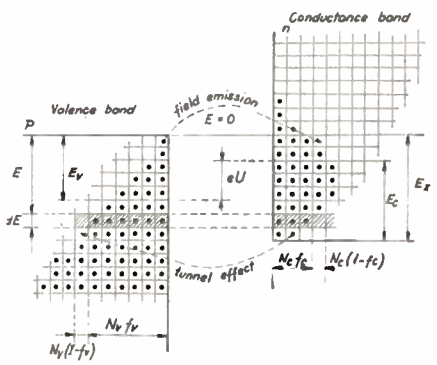


Fig. 1.

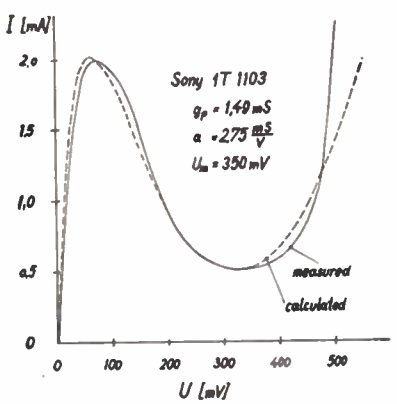


Fig. 2.

According to this formula the current has a minimum value at U_m, where the current decreases to zero. If we suppose in first approximation that the current in this point originates from ohmic surface effects, the equation of the characteristics of the nonidealized tunnel diode will be

$$I = a(U_m - U)^2 th\,\frac{1}{2}\,\frac{eU}{kT} + g_s U, \quad (14)$$

where

$$g_s = \frac{I_m}{U_m} \quad (15)$$

and

$$a = \frac{I_M - g_s U_M}{(U_m - U_M)^2 th\,\frac{1}{2}\,\frac{eU_M}{kT}}. \quad (16)$$

In Fig. 2 are compared the measured and the calculated characteristics of a tunnel diode, type Sony 1T1103.

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Linear Randomly-Varying Systems*

Zadeh's lucid and valuable analysis of significant developments in time-varying network theory[1] suggests that the following additional remarks may be of some interest concerning some still unpublished work[2] on randomly-varying systems—more specifically on a theory of linear transformations or operators in which the operator as well as the operand may be stochastic. Various statistical measures and possibly nth-order distribution functions are to be calculated for outputs of statistically varying systems in terms of similar measures of the input and distribution laws for the system operator. Although the approach is different, it is clear that Zadeh's work is closely related at some points. Thus by defining an appropriate transform of Zadeh's $R(jw, \tau)$

$$K_H(s, f) = \int_{-\infty}^{\infty} d\tau e^{2\pi i\tau(f-s)}R(s, \tau),$$

where $s = i\omega$, we can write

$$\phi_{Hz}(f) = \int_{-\infty}^{\infty} ds K_H(s, f)\phi_x(s)$$

as a relation between the spectral density $\phi_x(s)$ for the input or original process $x(t)$ and the spectral density of the output or transformed process $y(t)$ or $H\{x(t)\}$, where H is a stochastic operator representing the randomly time-varying system. We are making the often reasonable assumption that the properties of the medium H do not depend on the input signal x but may involve parameters $\alpha, \beta, \cdots$, of which one or more are random.

The kernel K_H is a spectral representation of the Green's function of the stochastic operator H and may be called a "stochastic Green's function." The form of this function depends on the particular statistical measure chosen. Often the correlation is more useful (and further, the resulting form is easily generalized to nonstationary operators and processes). The stochastic Green's function $G_H(\beta, \zeta)$ gives the autocorrelation R_{Hz} of the output in terms of the autocorrelation R_x of the input thus:

$$R_{Hz}(\beta) = \int_{-\infty}^{\infty} d\zeta G_H(\beta, \zeta)R_x(\zeta).$$

The form of these kernels in terms of "weighting functions" $h(t, \tau)$ for each realization H_i of the ensemble representing the stochastic operator H is

$$G_H(\beta, \zeta) = \int_{-\infty}^{\infty} d\tau \langle h(t, \tau)h^*(t + \beta, t + \zeta)\rangle$$

$$K_H(s, f) = \int_{-\infty}^{\infty} d\tau_1 \int_{-\infty}^{\infty} d\tau_2 \int_{-\infty}^{\infty} d\tau \cdot e^{2\pi i\tau f}e^{-2\pi i(\tau_2 - \tau_1)s}\langle h(t, \tau_1)h^*(t + \tau, \tau_2)\rangle.$$

Generalizations for mappings from one random function space to another in terms of any chosen "statistical measure" and existence conditions for the appropriate stochastic Green's functions have been developed and will be discussed elsewhere. The solution at least in principle of the general statistical problem of finding the output distribution in terms of the input distribution is made by expanding the output multivariate characteristic function in the correlations of the output, each of which is written in terms of input correlations and stochastic Green's function calculable from knowledge of the input distribution and the distributions in the operator, and then transforming (Levy's Inversion Theorem) to get the output distribution. Particularly of interest in the connection of Zadeh's paper is perhaps the fact these kernels have been calculated for some very difficult problems: *e.g.*, randomly sampled processes with any desired probability law for the intervals at which samples are taken of a random process, and for a problem in wave propagation in a random medium.

In the differential-equation approach, it may not always be necessary to restrict random coefficients to small excursions about an average, *i.e.*, to cases where a perturbation approach is necessary. If we consider a differential equation of the form $Ly = x$ where the differential operator L incorporates random coefficients and x is a random process, we would like to write an

* Received by the IRE, October 26, 1961; revised manuscript received, November 16, 1961.

[1] L. A. Zadeh, "Time-varying networks, I" Proc. IRE, vol. 49, pp. 1488–1503; October, 1961.

[2] G. Adomian, "Linear Stochastic Operators," Ph.D. dissertation, University of California, Los Angeles; August, 1961.

expression for example for the spectral density of $y(t)$, *i.e.*, $\phi_y(f)$, in terms of $\phi_x(f)$ and a stochastic Green's function K_H where H represents the inverse of L. Clearly if L can be inverted we have the problem already solved, but if it cannot then we would like to find an expression for K_H directly in terms of the statistics of L, *e.g.*, in terms of the covariance of the coefficients in L. Such expressions have been shown to exist in general. A trivial case would show that y can be given in terms of x and a kernel depending only on averages of L where $L=\langle L\rangle+\Delta L$, when ΔL is small. However, substantial progress has been made for cases where ΔL is not small and no restrictions exist on relative rates of variation in L and in y. These methods cannot be discussed coherently in the available space but are being published elsewhere.

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A Completely Consistent Definition of Temperature Sensitivity Including Negative Conductance Devices*

Probably the most widely-used criterion of receiver sensitivity is the noise figure, which expresses the degradation in signal-to-noise ratio caused by the noise generated by the receiver itself, *when the receiver is connected to a source whose "noise temperature" is 290°K*. Another criterion, which the author believes to be far more useful than noise figure, is the *temperature sensitivity* or system noise-temperature of the receiver.[1] This quantity refers the total noise in the system to the receiver input and expresses it in dimensions of degrees Kelvin.

It will presently be shown that the temperature-sensitivity concept requires no distinction as to whether the source conductance or output conductance of a network is positive or negative.[2] The admissibility of negative conductance is made possible by the use of the quantities *exchangeable power* and *exchangeable gain*.[3]

The exchangeable power of a source consisting of a sinusoidal current-generator in parallel with an admittance whose real part is G_g is given by

$$P_{eg} = \frac{i^2}{4G_g} \quad (G_g \neq 0), \tag{1}$$

where i^2 is the mean-square current of the generator and G_g is different from, but not necessarily greater than, zero. If a linear, two-terminal-pair network is connected to the source, an exchangeable power is defined for the output terminals of the network. The *exchangeable gain* of the network, when connected to the source, is defined by

$$A_e = \frac{P_{eo}}{P_{eg}}, \tag{2}$$

where P_{eo} is the exchangeable power at the output terminal pair of the network.

The available noise power of a passive conductance over a small band of frequencies df cycles wide is $KTdf$, where T is the thermal temperature of the conductance in degrees Kelvin and K is Boltzmann's Constant.[4] The conductance G may be represented by a noiseless conductance of G mhos in parallel with a current generator whose mean-square value is given by

$$i^2 = 4KTGdf. \tag{3}$$

Suppose the conductance is nonpassive, and possibly negative. The noisiness of such a conductance and the thermal temperature of the conductance are generally unrelated, since other mechanisms besides thermal noise contribute to the total noise. However, we *define* an *equivalent noise temperature* T_{eq}, such that the mean-square current of the noise-current generator is given by

$$i^2 = 4KT_{eq}Gdf \tag{4}$$

regardless of the nature of G. The sign of T_{eq} is positive if G is positive; negative if G is negative. Although T_{eq} is expressed in degrees Kelvin, it may be totally unrelated to the actual thermal temperature.

The exchangeable noise power of the conductance is given by

$$P_e = \frac{i^2}{4G} = KT_{eq}df. \tag{5}$$

P_e and T_{eq} are positive if G is positive; negative if G is negative.

The exchangeable noise power at the output terminals of a linear, two-terminal-pair network, when connected to this source of noise is, over a narrow band of frequencies df cycles wide,

$$dP_{eo} = KT_{eq}A_e df + dP_{eoi}, \tag{6}$$

where A_e is the exchangeable gain of the network when connected to the source, and dP_{eoi} is the excess noise power, exchangeable at the output terminals of the network, resulting from sources of noise internal to the network. The two terms of the right-hand side of (6) are either both positive or both negative according as the conductance presented by the output terminals of the network is positive or negative.

The *temperature sensitivity* of the network *with its source* is defined by

$$T_s = \frac{dP_{eo}}{KA_e df} = T_{eq} + T_r, \tag{7}$$

where

$$T_r = \frac{dP_{eoi}}{KA_e df}. \tag{8}$$

The quantity T_r is the *excess noise temperature* of the network. It can be shown that the sign of T_r is always the same as the sign of T_{eq}, which, in turn, is the same as that of the source conductance. The effect of the internal noise sources of the network is always to increase the absolute value of T_s above the absolute value of T_{eq}.

The excess noise temperature of two networks in cascade is given by

$$T_{r12} = T_{r1} + \frac{T_{r2}}{A_{e1}}, \tag{9}$$

where T_{r1} and A_{e1}, respectively, are the excess noise temperature and exchangeable gain of the first network, and T_{r2} is the excess noise temperature of the second network. The sign of T_{r12} is always the same as the sign of the source conductance, regardless of the sign of the output conductance of either network.

The temperature sensitivities and excess noise temperatures discussed above are spot quantities, referring to a particular frequency. The concepts of average excess noise temperature, average temperature sensitivity, and noise bandwidth are developed according to the same principles as those applied to the noise-figure definition.[5] Weighted averages are employed, where the weighting function is the exchangeable gain of the network (with its source) as a function of frequency.

Temperature sensitivity is an absolute, quantitative measure of receiver sensitivity because it specifies the amount of exchangeable signal power at a given frequency required at the antenna terminals for a particular output signal-to-noise ratio prior to demodulation. Temperature sensitivity is not dependent upon some arbitrary standard, whereas noise figure is based on a "room temperature" (290°K) antenna. Furthermore, the excess noise temperature T_r of the receiver is a function of the noise generated by the receiver alone. Noise from the antenna at image frequencies is included as a separate term in the temperature-sensitivity expression. Excess noise figure, on the other hand, includes image noise from the antenna. For these reasons, it is the opinion of the author that everything which is describable in terms of noise figure can be described better in terms of noise temperature.

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* Received by the IRE, March 21, 1961; revised manuscript received, October 9, 1961.

[1] C. T. McCoy, "Present and future capabilities of microwave crystal receivers," PROC. IRE, vol. 46, pp. 61–66; January, 1958.

[2] The term "conductance" refers to the real part of an admittance, which may or may not have an imaginary part.

[3] H. A. Haus and R. B. Adler, "An extension of the noise figure definition," PROC. IRE (Correspondence, vol. 45, pp. 690–691; May, 1957.

[4] It is assumed here that the temperature T is high enough or the frequency f is low enough that $KT \gg hf$, where h is Planck's Constant.

[5] Standards on electron devices: "Methods of measuring noise" (53 IRE 7 S1), PROC. IRE, vol. 41, pp 890–896; July, 1953.

Experimental Aperture Medium Coupling Loss*

Schott[1] has proved by experiments that the plane wave-power gain of the received antenna cannot be fully realized if the incoming radiation is of a more complex nature such as due to an aggregate of scattering elements randomly located.

Bullington[2] thought that the reduction of received antenna gain is due to the nonuniform phase front of the radio signal at the receiver, caused by the presence of multipath fading.

Gordon[3] noted that for large beamwidths the full plane wave-power gain of the antenna is realized; whereas for small bandwidths, the system antenna power gain (transmitting plus receiving) increases much more slowly. He explained the phenomenon as due to less scattering volume being seen by the smaller beam at the receiver.

Booker and de Bettencourt[4] were the first to give a more quantitative treatment and described this behavior as "aperture-medium coupling loss" of the antenna system (consisting of two antennas—transmitting and receiving), under the assumption of isotropic atmospheric turbulence and identical transmitting and receiving antennas. The loss depends on both beamwidth and path distance. As pointed out by the authors, their method will give values of coupling loss that are probably too large because the variations of the gradient of refractive index with height have been neglected in their calculations.

Staras[5] extended Booker and de Bettencourt's[4] treatment to nonconical and nonidentical antennas with the assumption of anisotropic atmospheric turbulence. He concluded that the coupling loss is not shared equally by the two antennas even if they are identical.

Hartmann and Wilkerson[6] continued the investigation, using the same assumptions of Booker and de Bettencourt[4] but in addition, taking into account the exponential decrease of the gradient of refractive index with height.

Because of the different assumptions used, the results given by these theoretical evaluations are very much diversified. A comparison of these calculated results is shown in Fig. 1. Since the aperture medium coupling loss is statistically in nature, the median value is of most importance. The calculated values as shown may be considered as the median values.

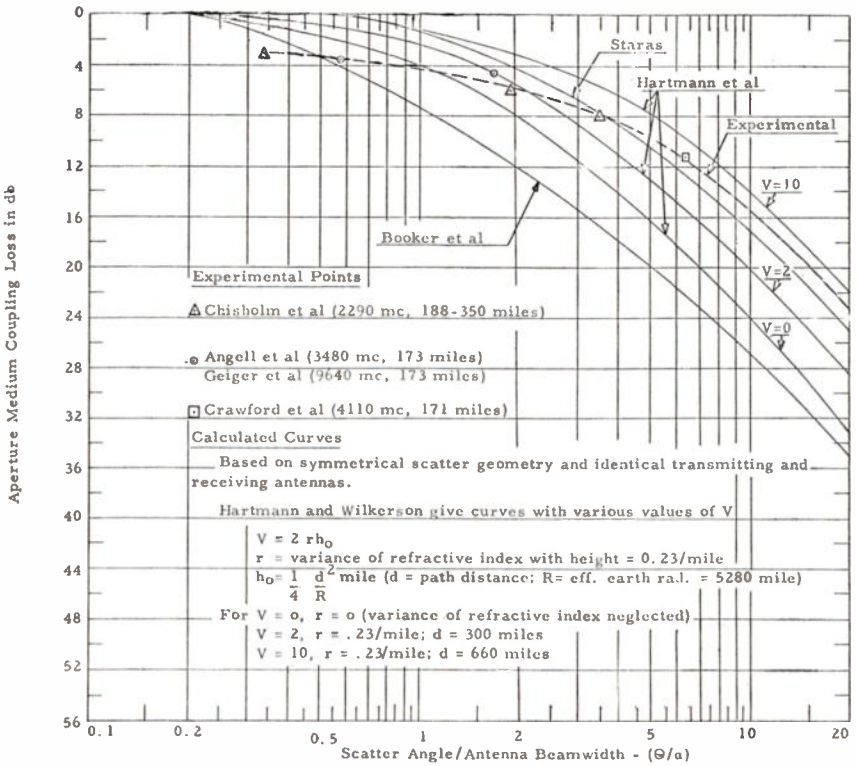


Fig. 1—Aperture medium coupling loss.

TABLE I

EXPERIMENTAL APERTURE MEDIUM COUPLING LOSS DATA

Reference Author	Location	Frequency (Mc)	Distance (mi)	Angular Distance θ	Antenna Diameter	Antenna Beam α	Antenna Gain db	θ/α Ratio	L_c 1 large antenna	L_c 2 large antennas
Chisholm *et al.*[7]	Crawford Hill-Round Hill	2290	188	2.1°	5 ft— 5 ft	6.1°—6.1°	29—29	0.34	—	3 db
		2290	188	2.1°	28 ft—28 ft	1.1°—1.1°	44—44	1.90	—	6 db
		2290	350	3.8°	28 ft—28 ft	1.1°—1.1°	44—44	3.50	—	8 db
Crawford *et al.*[8]	Crawford Hill-Pharsala	4110	171	1.9°	60 ft— 8 ft	0.3°—2.3°	56—38	6.30	5.7 db	11.4 db
Angell *et al.*[9]	Start Pt.-Wembley, Eng.	3480	173	1.5°	8 ft—horn	2.6°	36	0.57	1.8 db	2.6 db
Geiger, *et al.*[10]	Start Pt.-Wembley, Eng.	9640	173	1.5°	8 ft— 8 ft	0.9°—0.9°	46—46	1.66	—	4.5 db

There are available some long-term experimental data reported by Chisholm,[7] Crawford[8] Angell,[9] and Geiger[10] as shown in Table I. A curve based on the experimental data is also plotted in Fig. 1. It is seen that none of the calculated values follows this experimental curve very closely. In general, the calculated values are optimistic at low scatter angle/antenna beamwidth ratios but pessimistic at high ratios.

Since experimental data covers scatter angle/antenna beamwidth ratio up to 6.3 only, the points on the experimental curve beyond this ratio are obtained by extrapolation. It is suggested that this experimental curve be used to estimate the aperture medium coupling loss, which is considered as a good compromise among all theoretical curves.

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* Received by the IRE, October 16, 1961.

[1] F. W. Schott, "On the response of a directive antenna to incoherent radiation," PROC. IRE, vol. 39, pp. 677–680; June, 1951.

[2] K. Bullington, "Characteristics of beyond-the-horizon radio transmission," PROC. IRE, vol. 43, pp. 1175–1180; October, 1955.

[3] W. E. Gordon, "A simple picture of tropospheric radio scattering," IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-4, pp. 97–101; March, 1956.

[4] H. G. Booker and J. T. de Bettencourt, "Theory of radio transmission by tropospheric scattering using very narrow beams," PROC. IRE, vol. 43, pp. 281–290; March, 1955.

[5] H. Staras, "Antenna-to-medium coupling loss," IRE TRANS. ON ANTENNAS AND PROPAGATION: vol. AP-5, pp. 228–231; April, 1957.

[6] W. J. Hartmann and R. E. Wilkerson, "Path antenna gain in an exponential atmosphere," *J. Res. NBS (Radio Propagation)*, vol. 63D, pp. 273–286; November-December, 1959.

[7] Joint Technical Advisory Committee, "Long-range tropospheric transmission," pt. II of "Radio transmission by ionospheric and tropospheric scatter," PROC. IRE, vol. 48, pp. 30–44; January, 1960. See especially p. 36, Fig. 11-7.

[8] A. B. Crawford, *et al.*, "Studies in tropospheric propagation beyond the horizon," *Bell Sys. Tech. J.*, vol. 38, pp. 1067–1178; September, 1959.

[9] B. C. Angell, *et al.*, "Propagation measurements at 3840 Mc/s over 173-mile path," *Proc. IEE*, Suppl., pt. B, vol. 105, pp. 128–142; January, 1958.

[10] G. V. Geiger, *et al.*, "Propagation measurements at 3480 and 9640 Mc/s beyond the radio horizon," *Proc. IEE*, pt. B, vol. 107, pp. 531–546; November, 1960.

Llewellyn-Peterson Equations for Crossed-Field Beams*

A relation like the Llewellyn-Peterson equations was derived for a beam in a region of crossed-electric and magnetic fields by Van Duzer,[1] who added the force arising from the magnetic field to the Llewellyn analysis. The resulting transformation exhibits singularities which were interpreted as a possible new growth mechanism. These singularities are believed to be a result of the method used in the derivation of the relationship rather than a characteristic of the stream. It is possible to eliminate the singularities and obtain a new and less restrictive transformation by modifying the coordinate system. A brief summary of the transformation and its results[2] will be described.

To obtain any of the transformations,

1) obtain the complete electron trajectory in terms of the initial conditions and time;
2) add to the initial conditions a small ac perturbation and allow a small change in transit time to result;
3) let the resulting ac change in position be zero and solve for the ac change in transit time;
4) substitute the change in transit time in 2) to obtain the transformation.

In the one-dimensional model used by Llewellyn it is clear that the change in position used in 3) must be measured in the direction of motion. In a crossed-field region the motion is two-dimensional. We must therefore re-evaluate the change in position. A typical trajectory is shown in Fig. 1. The dotted curves show the maximum excursion caused by the perturbation.

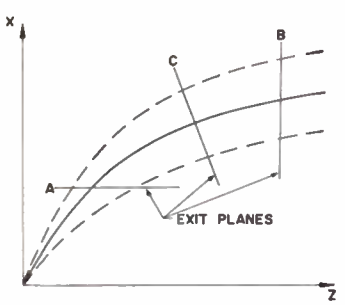


Fig. 1.

In using the x direction in 3) (plane A of the figure), the change in transit time grows as a result of the curvature of the trajectory, since the difference between the ac and dc trajectory increases. In the limit, when the ac trajectory does not cross the plane, the change in transit time is infinite. To partially eliminate this difficulty we could use the z direction in 3). This is shown in B of Fig. 1.

If we determine 3) from a condition involving the motion in both the x and the z directions we may orient the exit plane perpendicular to the average motion at the exit plane. This is shown in C of Fig. 1.

We do this by letting the slope of the trajectory be the negative reciprocal of the slope of the exit plane. The resulting transformation is given below.

$$\begin{bmatrix} V(\omega) \\ v_{xb}(\omega) \\ q_{xb}(\omega) \\ v_{zb}(\omega) \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} & a_{13} & a_{14} \\ a_{21} & a_{22} & a_{23} & a_{24} \\ a_{31} & a_{32} & a_{33} & a_{34} \\ a_{41} & a_{42} & a_{43} & a_{44} \end{bmatrix} \begin{bmatrix} I_x(\omega) \\ v_{xa}(\omega) \\ q_{xa}(\omega) \\ v_{za}(\omega) \end{bmatrix},$$

where

$$a_{22} = \left[\cos\theta - \frac{a_x}{\omega_c u_T} \cdot \left(\frac{u_x}{u_T}\sin\theta - \frac{u_z}{u_T}(1-\cos\theta)\right)\right] e^{-j\omega\tau}$$

$$a_{23} = \frac{1}{K}\left[-\sin\theta + \frac{a_x}{\omega_c u_T} \cdot \left(\frac{u_x}{u_T}(1-\cos\theta) + \frac{u_z}{u_T}(\theta-\sin\theta)\right)\right] e^{-j\omega\tau}$$

$$a_{24} = \left[-\sin\theta + \frac{a_x}{\omega_c u_T} \cdot \left(\frac{u_x}{u_T}(1-\cos\theta) - \frac{u_z}{u_T}\sin\theta\right)\right] e^{-j\omega\tau}$$

$$a_{32} = K\left[\sin\theta + \frac{\partial a_x/\partial\tau}{\omega_c^2 u_T} \cdot \left(\frac{u_x}{u_T}\sin\theta + \frac{u_z}{u_T}(1-\cos\theta)\right)\right] e^{-j\omega\tau}$$

$$a_{33} = \left[\cos\theta - \frac{\partial a_x/\partial\tau}{\omega_c^2 u_T} \cdot \left(\frac{u_x}{u_T}(1-\cos\theta) + \frac{u_z}{u_T}(\theta-\sin\theta)\right)\right] e^{-j\omega\tau}$$

$$a_{34} = K\left[\cos\theta - 1 + \frac{\partial a_x/\partial\tau}{\omega_c^2 u_T} \cdot \left(\frac{u_x}{u_T}(\cos\theta-1) + \frac{u_z}{u_T}\sin\theta\right)\right] e^{-j\omega\tau}$$

$$a_{42} = \left[\sin\theta - \frac{u_x^2}{u_T^2}\sin\theta - \frac{u_x u_z}{u_T^2}(1-\cos\theta)\right] e^{-j\omega\tau}$$

$$a_{43} = \frac{1}{K}\left[\cos\theta - 1 + \frac{u_x^2}{u_T^2}(1-\cos\theta) + \frac{u_x u_z}{u_T^2}(\theta-\sin\theta)\right] e^{-j\omega\tau}$$

$$a_{44} = \left[\cos\theta + \frac{u_x^2}{u_T^2}(1-\cos\theta) - \frac{u_x u_z}{u_T^2}\sin\theta\right] e^{-j\omega\tau}.$$

$K = j\omega\epsilon_0 B$, $u_T^2 = u_x^2 + u_z^2$, $\theta = \omega_c\tau$, $\omega_c = \eta B$.

$V(\omega)$ is the ac voltage between the two planes, parallel to the z axis, which bound the trajectory. $I(\omega)$ is the total current density between the two planes. $v_x(\omega)$ and $v_z(\omega)$ are the time-varying components of the x-directed and z-directed velocities, respectively, and $q_x(\omega)$ is the ac component of the x-directed convection current density. The subscripts a and b refer to the entrance and exit planes, respectively. B = the magnetic field. ω = radian frequency. $\eta = |e|/m$, the electron charge-to-mass ratio. u_x and u_z are the dc velocities in the x and z directions, and a_x is the dc acceleration in the x direction, all measured at the b plane. The coefficients a_{11}, a_{12}, a_{13}, and a_{14} may be obtained from Van Duzer.[1] The coefficients a_{21}, a_{31} and a_{41} have not been obtained using the method described. In cases where $I(\omega)$ is small these coefficients are not needed.

Early in their development, Pierce[3] calculated the noise figure of traveling-wave tubes, using the Llewellyn-Peterson Equations and a few simple assumptions. Although approximate, the calculation was important in understanding the noise processes. Using the same method the noise figure of a forward-wave amplifier has been calculated.

$$F = \left[(4-\pi)\left(\frac{T_c}{T}\right)\left(\frac{\omega_c}{\omega}\right)^2\right]\left[\frac{V_0}{KI_0}\right]\left[\left(\frac{1+\tanh\frac{\omega d}{u}\coth\frac{\omega a}{u}}{\operatorname{sech}\frac{\omega d}{u}}\right)^2\right] \times \left[\left|-j\frac{S(\delta_2+\delta_3)}{2\delta_2\delta_3} + \frac{S^2}{2\delta_2\delta_3} + \frac{1}{2} - g\right|^2 + \left(\frac{1}{\coth^2\frac{\omega a}{u}}\left(\left|-j\frac{S(\delta_2+\delta_3)}{2\delta_2\delta_3} - \frac{S^2}{2\delta_2\delta_3} - \frac{1}{2} - g\right|^2\right)\right)\right],$$

where T_c is the cathode temperature, T is the circuit temperature and all other parameters have been defined by Gould.[4]

The important assumptions are:

1) AC electric field and ac current are assumed in one direction only.
2) The velocity fluctuations at the cathode in both the x and z directions are the Rack velocity fluctuations.
3) The electron trajectories in the gun are like those in an infinite space-charge-limited crossed-field diode.
4) The small signal theory of Gould[4] is assumed to apply to the interaction region.

Measurements made on a backward-wave amplifier are within an order of magnitude of those predicted by the corresponding expressions developed for such tubes.

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[1] T. Van Duzer, "Transformation of fluctuations along accelerating crossed-field beams," IRE Trans. on Electron Devices, vol. ED-8, pp. 78–86; January, 1961.

[2] Electronics Res. Lab., Univ. of California, Berkeley, Quarterly Progress Rept. No. 4, U. S. Army Signal Corps Contract No. DA 36-039 SC-74981; November 1, 1960–January 31, 1961.

[3] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.

[4] R. Gould, "Space-charge effects in beam type magnetrons," *J. Appl. Phys.*, vol. 25, pp. 599–605; May, 1957.

Measurements of the Laser Output*

The output of the ruby laser has been the subject of much speculation. For this reason we have made quantitative measurements of this energy with a simulated blackbody absorber. This absorber was in the form of a hollow cone with a ½-in diameter aperture. The inner surface of the cone was painted black and another small inverted cone was placed directly behind the aperture to avoid reflections back into space. The system was then suspended by a thin rod of balsa wood in a container having a sapphire window. The air was pumped out of the container until a pressure of better than 10^{-5} mm mercury was reached. Fig. 1 shows this blackbody absorber.

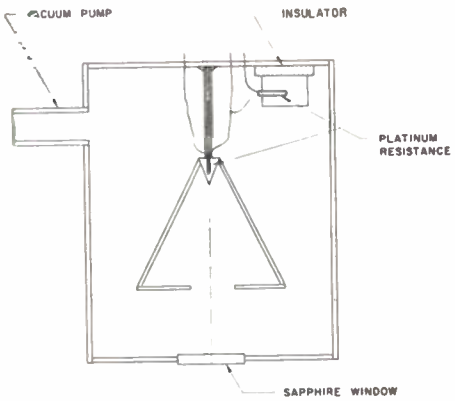


Fig. 1—Schematic drawing of the blackbody.

A bridge circuit was designed to measure the change in resistance of a platinum resistor imbedded in the blackbody, and thus, the temperature rise due to absorption of energy from a laser pulse could be detected. With the known mass and specific heat capacity of the blackbody, the rise in temperature could then be used to calculate the amount of energy absorbed. The accuracy of the system was such that a minimum energy of 10^{-2} joules could be measured. The system was also calibrated with a 600°C blackbody emitter.

Using this system we have measured the output energy of two commercial ruby lasers. The first used an Edgerton, Germeshausen and Grier FX-38 flash tube for its excitation and both the ruby and the flash lamp were housed in a cylindrical cavity. The ruby was 2½ in long (0.250-in diameter), had a chromium doping of 0.05 per cent and was operated at room temperature. Both sides of the ruby had multilayer dielectric coatings. Fig. 2 is a plot of laser output as a function of the input energy to the pump source. It is almost a straight line in the range of the experiment.

The second commercial laser had a 3-in long ruby with a 0.1 per cent chromium concentration. This ruby was cooled with nitrogen vapor and was silvered only on one side. This laser, which employed a helical flash tube for excitation with an input of 3600 joules, gave an output of 0.17 joule.

* Received by the IRE, December 1, 1961.

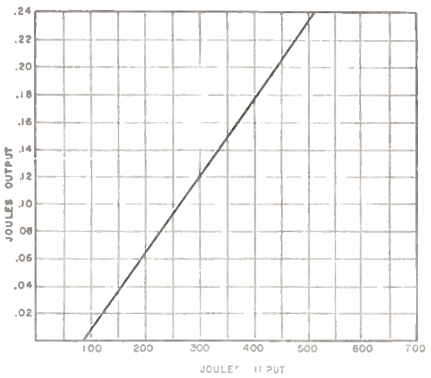


Fig. 2.

In another experiment the blackbody absorber was used to determine the output of a 1 meter long helium-neon gas laser. This measurement indicated an output of 3 mw.

In some instances the ability of the laser beam to punch holes through metallic foils has been used to arrive at an estimate of the output power. This method, however, has many inherent difficulties because the amount of energy reflected by the foil during the melting and evaporation and the amount passed through the hole after evaporation cannot be known accurately.

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High-Gain Dynamic Microwave Photomultiplier*

Current interest in optical communication techniques has created the need for photodetectors capable of responding to light with modulation components in the microwave region. It is the purpose of this note to report on recent experimental results with a photomultiplier which is capable of microwave response and to briefly describe the multiplication method which is called dynamic crossed-field electron multiplication (DCFEM).

This multiplication method is realized by providing a region of two spatially uniform fields—a microwave electric field and a crossed steady magnetic field—bounded by two electrodes, one having a high secondary emission ratio δ (active electrode), and the other a relatively low δ (inactive electrode) as shown sketched in Fig. 1. An electron which is present at the surface of the active electrode during a portion of the microwave period gains energy from the electric field and has a path which is curved by the magnetic field so that it is returned to the active electrode, producing secondary electrons upon arrival. This trajectory, labelled "normal" in Fig. 1, is repeated by the secondary electrons; and after n such multiplication steps, requiring precisely n microwave periods, δ^n electrons may be collected and delivered to an external circuit.

An analysis of DCFEM[1] which includes the effects of initial secondary electron velocities shows that for certain ranges of magnetic field, electrons leaving the active electrode over a phase interval as large as $\pi/2$ are compressed into successively smaller phase intervals on succeeding steps, providing a strong control over electron transit time. The output consists of a train of pulses with durations small compared to the microwave period and having a repetition frequency equal to the microwave frequency.

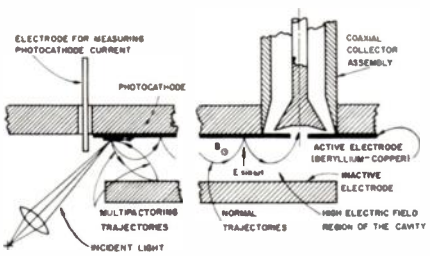


Fig. 1—Sketch of the essential parts of the experimental photomultiplier.

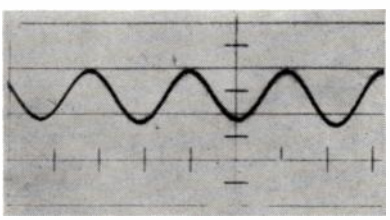
The principles of DCFEM may be readily applied to realize a photomultiplier by locating a photoemissive surface near one end of the active electrode. If this photocathode is irradiated by a light signal amplitude-modulated at frequencies not exceeding half the microwave driving frequency the envelope of the collector pulse train will correspond to the light modulation envelope and video amplification is obtained. A second and perhaps more interesting mode of detection is realized if the light is modulated at frequencies exceeding the driving microwave frequency. In this case the envelope of the output pulse train corresponds to the frequency differences between the light modulation components and the microwave driving frequency, thus providing superheterodyne detection.

In the experimental photomultiplier sketched in Fig. 1, the two electrodes are incorporated in the high electric field region of a rectangular re-entrant cavity resonant at 3 Gcps. The spacing between the electrodes is 3 mm and the electric-field intensity is in the range 10^5 to 10^6 volts/meter, providing 8 or more multiplication steps. The active electrode is Beryllium-Copper, a small area of which (20 mm²) is covered with a Cesium-Antimonide photocathode. An external electromagnet supplies a uniform field of 500 Gauss. The light incident upon the photocathode is sinusoidally modulated at 300 cps by passing light from a

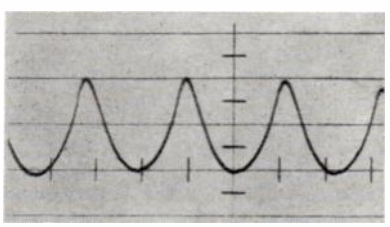
* Received by the IRE, November 16, 1961; revised manuscript received, December 4, 1961. This work is sponsored by AFOSR, ARDC, under Contract No. AF49(638)-556.

[1] O. L. Gaddy and D. F. Holshouser, "Analysis of Dynamic Crossed-Field Electron Multiplication," Elec. Engrg. Res. Lab., Engrg. Experiment Station, Univ. of Illinois, Urbana, Tech. Note No. 1, Contract No. AF49(638)-556; November, 1961.

linearly polarized source through a spinning polaroid disk, maintaining a peak-to-peak photocathode current of 10^{-10} amperes. Fig. 2 shows two oscillographs of the resulting multiplier output current for two different electric field conditions. Fig. 2(a) shows the output current for the maximum electric field intensity which gave linear multiplication. The output current in this case is 10 μa giving a linear gain of 10^5. Fig. 2(b) shows the output current when the electric field intensity is increased further, resulting in a saturation phenomenon. A form of mutipactoring occurs at the higher electric fields with probable electron trajectories as shown in Fig. 1. Several techniques for reducing this effect and attaining higher output current levels are being considered.



(a)



(b)

Fig. 2—(a) Output current for maximum electric field which gives linear operation. Time scale: 1.6 msec/cm; current scale: 10 μa/cm. (b) Output current for increased electric field intensity. Time scale: 1.6 msec/cm; current scale: 10 μa/cm with increasing current downward.

Having obtained a high linear gain, the next step in the experimental program is the detection of microwave modulated light[2] utilizing the superheterodyne method discussed above. Although this multiplication scheme is not the first dynamic multiplier that has been proposed,[3,4] it does possess the advantages of tight control of transit time and a controlled number of multiplication steps that other multiplication schemes lack. It is believed that this scheme has promise as a low-level, low-noise detector of high-frequency modulated light, both in optical communication and as a basic research tool.

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[2] D. F. Holshouser, H. Von Foerster, and G. L. Clark, "Microwave modulation of light using the Kerr effect," *J. Opt. Soc. Am.*, vol. 51, pp. 1360–1365; December, 1961.
[3] P. T. Farnsworth, "Television by electron image scanning," *J. Franklin Inst.*, vol. 218, pp. 411–444; October, 1934.
[4] G. B. Banks, U. S. Patent No. 2,113,214; April, 1938.

Origin of Noise in a Ferromagnetic Parametric Amplifier*

Measurements on a ferromagnetic amplifier which have been reported elsewhere[1] indicated a surprisingly high noise figure for the amplifier. The results were surprising because it had been commonly assumed that the ferromagnetic amplifier would share the low noise figure characteristic of most other parametric amplifiers. Further measurements on the amplifier are reported here which indicate that the noise is not thermal in origin.

The amplifier on which the measurements were made is described in an earlier publication.[2] Briefly, it consists of a small sphere of ferromagnetic material mounted in a cavity resonant at the pump frequency with a coaxial transmission line coupled to the material by means of a loop of wire around the sphere. In this amplifier resonances are provided at signal and idler frequencies by modes of the ferromagnetic material which can have low-order spatial variations (magnetostatic modes[3]) and, thus, can couple to the loop rather easily. Time-varying reactive coupling is provided between the sample modes by the cavity RF magnetic field which is directed parallel to an applied dc magnetic field. The signal and idler resonances are tuned to the correct frequencies ($f_s+f_i=f_p$) by adjustment of the dc magnetic field. One-half watt of pump power at a frequency of approximately 11,000 Mc is used in the measurements and a dc magnetic field of around 1800 oersteds is required.

Sample resonances also exist which have very high-order spatial variations (spin waves) so that they do not radiate and cannot be loaded by the loop. Furthermore, the spin waves are parametrically pumped by the rf magnetic field[4] and have sufficiently low losses so that they are oscillating at the pump-power levels required for amplifier operation. In reporting the previous measurements it was noted that these oscillating spin waves might be the source of noise in this amplifier, rather than some sort of thermal process. It was suggested that measurements of the noise figure as a function of temperature would provide a means for differentiating between the two sources of noise. If the main source of noise is a thermal process, then the noise power should decrease linearly with temperature. While no analysis has been carried out which would predict the variation with temperature of noise power contributed by oscillating spin waves, it is likely that this noise power would be more or less constant, or even increasing, as temperature is reduced.

Measurements of the noise figure have been made over a range of sample temperatures down to 50°K using standard techniques for the noise figure measurements. The results are shown in Table I where the measured noise figure has been converted to a single-sideband noise figure. Results are also given for the voltage gain-bandwidth product. Accurate measurement of the noise figure was difficult because of the narrow bandwidth of the amplifier so that the results may be in error by as much as 1 db. However, the data indicate that the noise figure is essentially constant as a function of temperature. This proves rather conclusively that the noise is not thermal in origin.

TABLE I
Variation of Amplifier Characteristics with Temperature

Temperature of Sphere (°K)	Center Frequency (cps)	Voltage Gain ×Bandwidth (cps)	Noise Figure (db)
370	5575×10^6	5×10^6	13
272	5484×10^6	5×10^6	10.5
120	5600×10^6	1×10^6	12
50	5462×10^6	1×10^6	11.5

In this type of parametric amplifier, where magnetostatic modes of a sample of ferromagnetic material are used for both signal and idler resonances, it is inevitable that when conditions are adjusted for amplification, spin waves will be pumped to oscillation. These oscillating spin waves increase the energy of the magnetic lattice and, as shown by the data, result in a high effective spin temperature for the signal and idler resonances.

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* Received by the IRE, November 30, 1961.
[1] R. T. Denton, "Theoretical and experimental characteristics of a ferromagnetic amplifier using longitudinal pumping," *J. Appl. Phys.* (Suppl.), vol. 32, pp. 300–307; March, 1961.
[2] *Ibid.*, p. 303.
[3] L. R. Walker, "Magnetostatic modes in ferromagnetic resonance," *Phys. Rev.*, vol. 105, pp. 390–399; January, 1957.
[4] E. Schlömann, J. J. Green, and U. Milano, "Recent developments in ferromagnetic resonance at high power levels," *J. Appl. Phys.* (Suppl.), vol. 31, pp. 386–395, May, 1960; F. R. Morgenthaler, "Survey of ferromagnetic resonance in small ferrimagnetic ellipsoids," *Ibid.*, pp. 95–97.

An Extremely-Low-Noise 6-Gc Nondegenerate Parametric Amplifier*

A 6-Gc nondegenerate parametric amplifier was built and was operated successfully both at room temperature and at liquid nitrogen temperature. The amplifier was pumped at 23 Gc. The effective input noise temperature of the amplifier, including the insertion loss of the room temperature circulator and the mixer noise, was less than 180°K at room temperature and was less than 60°K at liquid nitrogen temperature; of this more than 20°K is attributed to the circulator and the mixer.

The development of this amplifier is not yet complete; the entire amplifier including

* Received by the IRE, November 9, 1961; revised manuscript received, November 21, 1961.

the circulator will finally be contained in a closed dewar and will be operated continuously at liquid nitrogen temperature with very little maintenance. This is a report on preliminary experiments which were carried out in a laboratory dewar and with a room temperature circulator. More detailed information will be published at a later date.

The main features of the amplifier are as follows:

Input and output frequency	5.85 Gc
Pump frequency and power	23 Gc, 5 mw
Varactor diode	Gallium arsenide point contact diode. $C_0 \approx 0.42$ pf at 300°K (includes about 0.07 pf of package capacitance)
Dynamic quality factor $\tilde{Q}$ of the amplifier at 5.85 Gc (including the circuit loss)	≈6 at 300°K, and about 20 per cent higher at 77°K
Bias voltage	−1.2 volt
Normalized input impedance R_s/R_0	≈1/8.5 at 300°K and 1/10.5 at 77°K
Normalized idler impedance R_s/R_L	≈2
Gain	20 db
Estimated noise temperature (Amplifier alone)	150°K at 300°K 38°K at 77°K.

A descriptive drawing of the amplifier is shown in Fig. 1. The amplifier construction resembles that of the lower sideband up-converter described in a previous paper.[1] Since the junction capacitance of the gallium-arsenide diode decreases about 20 per cent from room temperature to liquid-nitrogen temperature (the series resistance, however, decreases only about 4 per cent), a considerable readjustment of the amplifier is needed from room-temperature operation to liquid-nitrogen-temperature operation. This adjustment is complicated by the fact that, in order to minimize the unnecessary loss in the amplifier circuit, only a simple teflon trimmer is used in the idler circuit and none in the input circuit, and hence, the main adjustment has to be made at room temperature. After a thorough investigation of the temperature characteristic of the amplifier, the input circuit was tuned (*i.e.*, adjusted to have net zero reactance) at −2.3 volts of the diode bias; the idler circuit, at −4 volts. When the temperature of the amplifier was lowered to liquid nitrogen temperature, the amplifier was almost in the optimum condition at −1.2 volts of bias voltage. The final tuning was made by the teflon trimmer in the idler circuit.

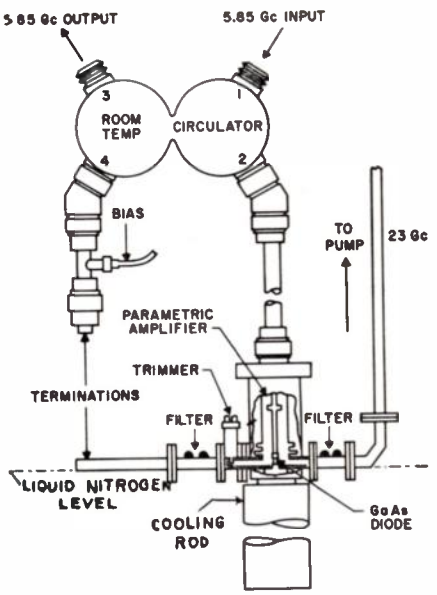


Fig. 1—Refrigerated 6-Gc one-port parametric amplifier. Only the cooling rod is immersed into liquid nitrogen.

The bias voltage is applied through the fourth port of the circulator; this eliminates the unnecessary loss and instability often caused by the bias circuit. The circulator[2] and the stainless-steel coaxial line were very carefully made so as to reduce the total insertion loss of these components to below 0.15 db and the VSWR to better than 1.07 at 5.85 Gc.

The amplifier was first operated at room temperature, and an over-all noise figure of better than 2.1 db (effective input noise temperature of 180°K) was measured with 20-db gain and 25-Mc bandwidth. Since the insertion loss of the input circuit was about 0.15 db and the single-sideband noise figure of the mixer and IF amplifier was 7.5 db, the measured result is very close to the expected value. The bandwidth is limited mainly by the idler circuit; the amplifier can be tuned over 260 Mc simply by tuning either the pump frequency or the idler circuit. By a careful design of the idler circuit, the instantaneous bandwidth will be increased.

To operate the amplifier at liquid nitrogen temperature, as discussed before, the amplifier was completely readjusted at room temperature in order that it would tune at −1.2 volts of bias when its temperature was lowered to 77°K. At this temperature, the performance of the amplifier could then be optimized with a slight adjustment of the teflon trimmer in the idler circuit. The measured over-all noise temperature was less than 60°K; of this more than 20°K was contributed by the circulator and the mixer. Since the performance of the amplifier was temperature sensitive, its temperature had to be maintained constant to within 5°K. The pump power used was 5.5 mw at 23 Gc to achieve a gain of 20 db and a system noise temperature of 60°K.

Thus far, three room temperature amplifiers and two liquid nitrogen ones have been built; in each case the performance was close to that described above. In the final model the entire amplifier, including the circulator, will be contained in a closed dewar and will be operated continuously at liquid-nitrogen temperature. A further improvement in the over-all noise temperature for the final model is expected.

The author wishes to express his appreciation to S. Plauski, who did many of the measurements and amplifier adjustments, N. C. Vanderwal, who supplied the gallium arsenide diodes, and R. G. Voss, who did the mechanical designs.

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An Anomalous Effect in the Characteristics of GaP Point Contact Diodes*

We have observed on single crystals (~0.2 cm×0.2 cm×0.1 cm) of *N* type gallium phosphide (resistivity ~0.2 Ω cm), using a tungsten point for one contact and silver as the ohmic contact, typical current-voltage traces[1] taken at a frequency of 60 cps as shown in Fig. 1. These traces indicate the appearance of an apparent "hysteresis" effect in the forward direction. The leading trace (especially at point *a*, curve 1) of the figure shows a marked deviation from the return trace (curve 2, Fig. 1). Above point *b* on curve 1, Fig. 1, the slope of the current-voltage curve changes until point *c* where both the initial and return trace (curve 2, Fig. 1) come together. The return trace does not follow the leading trace as shown. For the same current, the voltage on the return trace is observed to be larger than on the initial trace. The reverse characteristics are also not exactly reproducible, but the effect



Fig. 1—Oscilloscope trace taken at 60 cps of current-voltage characteristics of GaP showing "hysteresis" effect with and without light. Horizontal scale: 1 v/cm; vertical scale: 12.3 μa/cm.

[1] M. Uenohara and H. Seidel, "961-Mc lower-sideband up-converter for satellite-tracking radar," *Bell Sys. Tech. J.*, vol. 40, pp. 1183–1205; July, 1961.

[2] Raytheon circulator CCL-4 (room temperature model) and CCL-5 (liquid nitrogen).

* Received by the IRE, November 10, 1961; revised manuscript received, November 24, 1961.

[1] All measurements reported in this communication were made at a temperature of 23°C.

is not nearly as great. Upon irradiation of the junction with (visible) white light, the effect in the forward direction is considerably enhanced as shown in Fig. 1, curve 3, and in the region equivalent to *b*, the voltage is observed to increase at a faster rate than the current. The return trace again coincides closely with that found without light.

DC current-voltage characteristics are shown in Fig. 2. The characteristics were measured both in the increasing and decreasing direction and the curves were reproducible showing no anomaly.

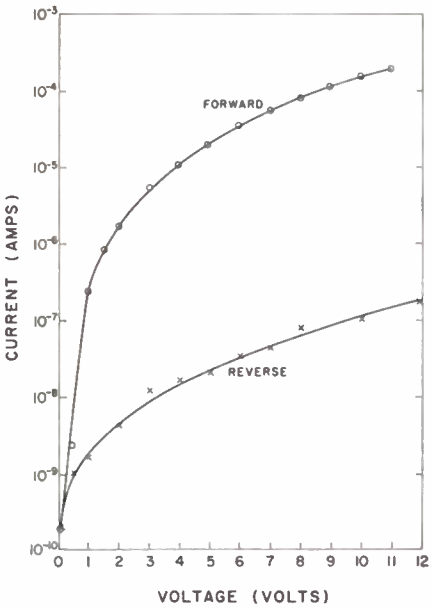


Fig. 2—Typical dc current-voltage characteristics of GaP point contact diode.

A second point contact with dc bias (positive) was then placed about 0.1 cm away from the first contact; the "hystersis" effect could be "quenched" with a 10 μ a bias current. Even the use of light did not return the effect.

The effect shown in Fig. 1 was also noted for other frequencies in the audio range.

A possible explanation of our experiments may be related to the likelihood of trapping centers being formed in the space-charge region. If these trapping centers were negatively charged, the charged traps could reduce the built-in voltage, permitting a larger current flow (injection of holes) at the lower voltage level (as for example point *a*, curve 1, Fig. 1) than expected from a diffusion current. Some of these injected carriers (holes) begin to fill the traps increasing the built-in voltage to its normal value and decreasing the rate of current contribution until point *b*. Above point *b*, the normal diffusion current takes over. The rate of recombination or generation of traps would be expected to be frequency dependent.

The total current in the forward direction might then be thought to be composed of the current due to traps and the normal diffusion current. The contribution of each predominates in different regions of the curve; at point *c* the normal diffusion current predominates and normal *I-V* characteristics prevail. Sah, Noyce and Shockley[2] have dealt with manifestations of these effects. Further, Loebner and Poor[3] have postulated that an emission band observed at 9000 Å in GaP may be associated with hole injection into *N* type GaP followed by recombination at a copper center.

The return trace (Fig. 1) indicates that the traps are being emptied when the voltage is in the negative part of the cycle. The rate of emptying traps is much faster than the rate of filling the traps. The *I-V* characteristic has the appearance that one would normally expect for a diode in the forward direction.

When light is applied, more ionized traps are created, depleting the space charge at a faster rate, and the increase in current, as shown by curve 2, (even begins at a lower voltage). The return trace coincides with the case for no light.

With the use of a positive dc bias, it is believed that the injection of holes is great enough to fill the traps at a rate faster than the 60 cps rate and, as a result, the effect is not noted on the trace.

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[2] C. T. Sah, R. N. Noyce, and W. Shockley, "Carrier generation and recombination in *p-n* junctions and *p-n* junction characteristics," Proc. IRE, vol. 45, pp. 1228–1243; September, 1957.
[3] E. E. Loebner and E. W. Poor, Jr., "Electroluminescence in single crystals of GaP," *Bull. Am. Phys. Soc.*, vol. 4, p. 45; January, 1959.

The Reflex Klystron as an Amplifier at 73 kMc*

The purpose of this note is to report millimeter-wave amplification at 73 kMc by a QKK 838 reflex klystron. To the author's knowledge, the operating frequency is the highest ever reported for this type of amplifier.[1,2]

A schematic diagram of the QKK 838 amplifier is shown in Fig. 1(a). The reflex klystron was connected to the main waveguide through an *H*-plane crossguide. The end of the crossguide arm opposite the tube was terminated by a shorting plunger. This shorting plunger adjusts both the load admittance and the coupling of the tube to the main waveguide. The load admittance of the tube was also adjusted by EH tuners connected at the input and output of the amplifier circuit.

When the circuit admittance and the electronic admittance of the tube are properly adjusted, regenerative amplification results. The gain of this amplifier was measured by comparing the amplifier output with the output of a straight waveguide section which was substituted for the amplifier circuit by waveguide switches. The gain of the amplifier is plotted as a function of the repeller voltage in Fig. 1(b). A maximum gain of 17.5 db at 73.76 kMc was obtained.

Fig. 1(c) is an oscilloscope picture showing the variation of the amplifier output with repeller voltage. The method used was essentially the same as that used by Quate, *et al.*[1] The repeller voltage of the amplifier klystron was swept by a sawtooth wave of approximately 120 cps. In this case the maximum gain of the amplifier was 9 db. This relatively low gain is due mainly to saturation of the amplifier due to the relatively large input signals which were required for an oscilloscope display. This amplifier showed strong saturation effects for high-input signals. The maximum gain dropped to 1.5 db when the input power level was raised 46 db above the case for Fig. 1(b). However, this amplifier showed very good linearity for small signals. The relative output level of the amplifier is plotted against the input level in Fig. 2. The linear relation of the input versus output, together with the increase in output, indicate that the QKK 838 is actually amplifying and is not just oscillating.

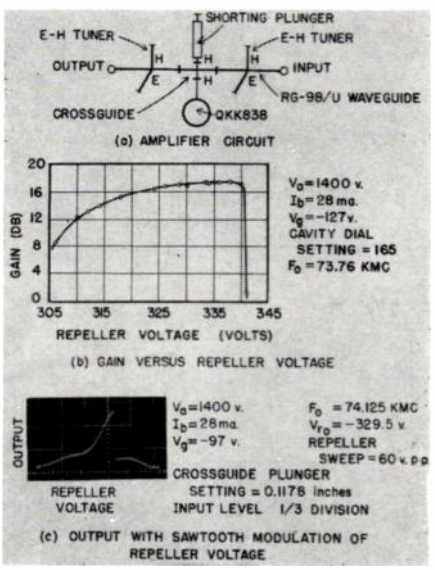


Fig. 1—Gain of QKK838 amplifier.

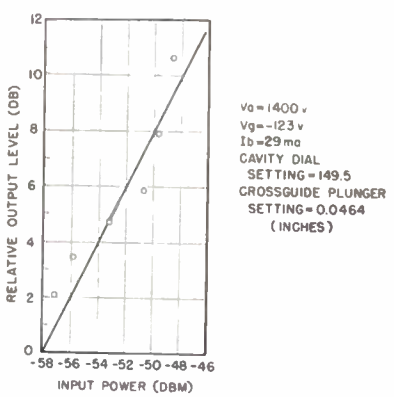


Fig.—QKK838 amplifier linearity.

* Received by the IRE, December 4, 1961.
[1] C. F. Quate, R. Kompfner, and D. A. Chisholm, "The reflex klystron as a negative resistance-type amplifier," IRE Trans. on Electron Devices, vol. ED-5, pp. 173–179; July, 1958.
[2] K. Ishii, "On use of reflex klystrons for microwave and millimeter waves," *Proc. National Electronics Conf.*, vol. 16, pp. 744–752; October, 1960.

The bandwidth of this amplifier was dependent on the gain. If the circuit was adjusted for low gain, a relatively wide bandwidth was obtained. For example, when the amplifier was adjusted for a gain of 5.5 db at a center frequency of 73.7 kMc, the 3-db bandwidth was 170 Mc. The reflex klystron is inherently a narrow-band device. No wideband characteristics can be expected, as with traveling-wave amplifiers or backward-wave amplifiers. However, wider bandwidths can still be expected than is possible with masers at the same operating frequencies.[3]

Though no effort to stabilize the amplifier was made, the QKK 838 amplifier was just stable enough to take the data. For a while, the output of the amplifier did drift. Thus, it is felt that for stable operation controlled environmental conditions and an automatic gain control circuit to adjust the repeller voltage for stable output would be needed. Also, isolators, circulators, or damping attenuators are extremely important for the stabilization of this type of amplifier.

Though no effort to minimize the noise figure of the QKK 838 amplifier was made, the noise figure of this amplifier was *not* promising. A noise figure of 40.3 db was obtained by the small signal method. Reflex klystron amplifier noise figures depend on the type of tube and the circuit configuration.[4] For example, if a grid-less reflex klystron such as a VA 99 were used with a carefully stabilized circuit of a different configuration, the noise figure might be reduced.

Thus, at present, the QKK 838 amplifier is suitable for amplification of signals which are as small as -58 dbm, but for extremely small or large signals, the tube will not produce usable gain.

The experimental results reported were obtained with the QKK 838 serial No. 17, which was the only amplifier tube available at the author's laboratory. If ten or more tubes were tested, they might show different results, for there will be slight individual differences in the characteristics of such millimeter-wave klystrons. An aging effect was observed on the tube used during this experiment. At the time, the gain of 17.5 db was observed, the tube could be used to generate 42 milliwatts at 74 kMc when operated as an oscillator. After 6 months, the output was reduced to only 200 microwatts. Under this aged condition, the gain of the amplifier was at most 3 db.

Though the reflex klystron tube in this extremely high-frequency range is very expensive, it is still economical in comparison with other microwave amplifiers such as the traveling-wave amplifier, the backward-wave amplifier, the maser, and the parametric amplifier with pumping oscillator. At present, according to the author's investigation, millimeter-wave Esaki-diode amplifiers cannot compete with the reflex klystron economically. Esaki diodes for the lower frequency ranges are inexpensive. At higher frequencies, however, Esaki diodes are more expensive than reflex klystrons. At X band, for example, Esaki diodes are approximately 10 times the cost of reflex klystrons. For millimeter-waves, Esaki diodes have shown promising results in laboratory tests.[5] However, a millimeter wave Esaki diode is not now available on the market. At this time, reflex klystron is capable of a higher output power than the Esaki diode in this extremely high-frequency range. The reported results indicate that if a QKK 867 or 100V10 reflex klystron were used, amplification at 100 kMc will be possible.

The authors thank Raytheon Company for contributing the klystron tubes and A. L. Brault, S. Krupnik, Jr., J. A. Stefancin, and J. E. Billo for their assistance.

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[3] H. Heffner, "Masers and parametric amplifiers," *Microwave J.*, vol. 2, pp. 33–40; March, 1959.
[4] K. Ishii, "Noise figure of reflex klystron amplifiers," IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 291–294; May, 1960.
[5] C. A. Burrus and R. Trambarulo, "A millimeter-wave Esaki diode amplifier," Proc. IRE (Correspondence), vol. 49, pp. 1075–1076; June, 1961.

Visual Matrix Multiplication by Flow Graphs*

Consider the following relationship involving the general circuit matrix:

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} A'B' \\ C'D' \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix}. \tag{1}$$

This can be represented by the flow graph of Fig. 1. The right-hand terminals on Fig. 1 represent the elements in the right-hand vector of (1) and the left-hand terminals represent the elements in the left-hand vector. The flow graph between these terminals represents the matrix of (1).

If we had three cascaded networks we would have the relationship:

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} A''' & B''' \\ C''' & D''' \end{pmatrix} \begin{pmatrix} A'' & B'' \\ C'' & D'' \end{pmatrix} \cdot \begin{pmatrix} A' & B' \\ C' & D' \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \tag{2}$$

wherein the primes identify each matrix with a particular network. We would like to find the multiplication product of the three matrixes in (2). This multiplication product will be a single matrix that will represent the three cascaded networks. That is:

$$\begin{pmatrix} A & B \\ C & D \end{pmatrix} = \begin{pmatrix} A''' & B''' \\ C''' & D''' \end{pmatrix} \begin{pmatrix} A'' & B'' \\ C'' & D'' \end{pmatrix} \cdot \begin{pmatrix} A' & B' \\ C' & D' \end{pmatrix} \tag{3}$$

When networks are cascaded, the input voltage and current of one component network is the output voltage and current of the network preceding it. Since the flow graph of Fig. 1 converts the output voltage and current to the input voltage and current for the network it represents, we may cascade such flow graphs of cascaded networks to get the over-all flow graph of the system. Therefore, the flow graph of (2) would be represented by Fig. 2. It will be observed that this flow graph is unidirectional, *i.e.*, has no loops, so that its evaluation will be relatively simple. From this flow graph we can determine the elements in the left-hand matrix of (3).

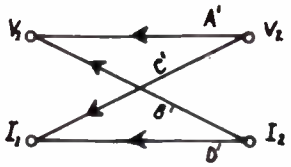


Fig. 1—Flow graph of (1).

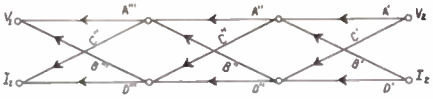


Fig. 2—Flow graph of (2).

Element A will be equal to the sum of all the path products from V_2 to V_1; similarly, B will be equal to the sum of the path products from I_2 to V_1, C will be equal to the sum of the path products from V_2 to I_1 and D will be equal to the sum of the path products from I_2 to I_1. By visual inspection of Fig. 2 we find:

$$\begin{aligned} A &= A'A''A''' + A'C''B''' + C'B''A''' \\ &\quad + C'D''B''' \\ B &= D'D''B''' + D'B''A''' + B'A''A''' \\ &\quad + B'C''B''' \\ C &= C'D''D''' + C'B''C''' + A'A''C''' \\ &\quad + A'C''D''' \\ D &= D'D''D''' + D'B''C''' + B'C''D''' \\ &\quad + B'A''C'''. \end{aligned} \tag{4}$$

Thus we have a visual means of matrix multiplication. As one gains facility with this method it should prove a time saver and a source of fewer errors compared to formal matrix multiplication.

As an aid in using this method note that when none of the multiplied matrixes has a zero element the number of paths will equal m^n where m is the degree of the matrix and n is the number of multiplications. In our case $m^n = 4$. Thus there will be four path products in the right-hand side of each of Eq.'s (4).

Another matrix often multiplied to solve network problems is the wave matrix.[1] Its flow graph is given in Fig. 3, wherein I_1 and

[1] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Co., Inc., New York, N. Y., p. 551; 1948.

* Received by the IRE, December 1, 1961.

R_1 are waves traveling towards the load and generator, respectively, at the input terminals, and I_2 and R_2 are waves traveling towards the load and generator, respectively, at the output terminals. τ is the transmission coefficient, and Γ_1 and Γ_2 are the reflection coefficients referred to the input and output terminals, respectively.

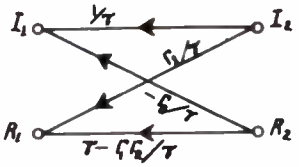


Fig. 3—Flow graph of wave matrix equation.

Should a matrix have a zero element such as the general circuit matrix for a shunt Y or series Z, or the wave matrix for a line length, merely delete the pertinent path from the flow graph.

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One-Tunnel-Diode Flip-Flop*

Kaenel[1] presents an arm-chair analysis of a flip and flop having one each: resistor and inductor tunnel diode. He cites the reliability and stability one would expect from such a simple configuration. The purpose of this correspondence is to report on the operation of a similar circuit and to present oscillographs clearly showing the triggering and switching modes.

Fig. 1 shows the circuit diagram of the flip-flop connected through a dc blocking capacitor and isolating resistor to a variable pulse generator. The characteristic curve of the tunnel diode with the resistive load line superimposed upon it in Fig. 2 shows that there are two stable points, A and C (disregard lines with arrows). Point B is in the negative resistance portion of the curve and the diode cannot be stably biased there unless the slope of the load line is steeper than the tunnel-diode characteristic.

For bidirectional triggering, the incoming positive pulse must be converted into a positive going pulse followed by a negative going pulse appearing across the tunnel diode. This could be accomplished simply by differentiating the pulse with an RC network. Alternately, the authors' circuit utilizes the capacity across the tunnel diode due to junction capacitance and oscilloscope input capacitance. The pulse from the signal generator charges the tunnel-diode capacitance which causes damped oscillation to begin between the circuit inductor and tunnel diode capacitance. Thus the positive pulse from the pulse generator causes a doublet or combination of positive and negative pulses to appear across the tunnel diode as if differentiation had occurred. For simplicity of presentation, this doublet will be called the "differentiated pulse."

Consider the circuit to be initially at point A. The positive portion of the differentiated input pulse forces the operating point of the diode over I_p. This immediately causes the tunnel diode to switch to its high-voltage state since the region of negative resistance is unstable. The negative portion of the differentiated pulse will then try to lower the voltage across the diode but the back voltage of the inductor will keep it from reaching E_v and switching again. After this, the circuit will slowly proceed to point C according to the L/R time constant, which is

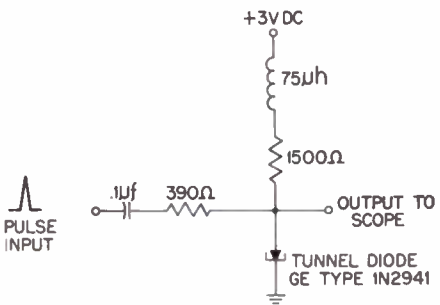


Fig. 1—Circuit diagram of tunnel-diode flip-flop.

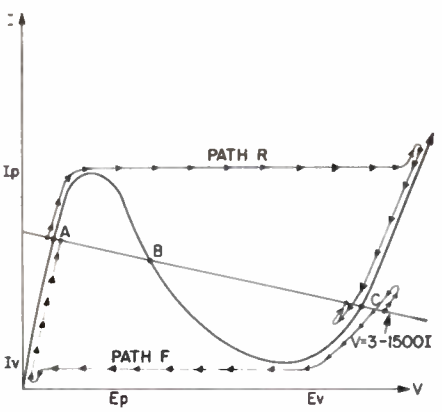


Fig. 2—Tunnel-diode characteristic and resistive load line.

far longer than the pulse duration. Path R in Fig. 2 shows the path of operation of the tunnel diode when triggered to a high-voltage state.

The next pulse finds the diode in its high-voltage state. The positive portion of the differentiated pulse raises the voltage temporarily while the following negative portion drives it below E_v and switching again occurs. Switching from the high-voltage to low-voltage state by the negative portion of the differentiated pulse is now possible because the back voltage from the inductor is tending to keep the current at a lower level than during the other switching mode. Path F in Fig. 2 shows the path of operation when triggered to a low-voltage state.

Fig. 3 shows the square wave output when the flip-flop is driven by a pulse train. The pulses are visible as dots above the corners of the wave. A Tektronic Model 532 Oscilloscope with 53/54 plug-in was used for all pictures (0.2 v/cm, 50 μsec/cm). Fig. 4 shows the rise portion of the square wave with the triggering pulse superimposed on top (0.2 v/cm, 2 μsec/cm). The oscilloscope was set to chopped output so that the time sequence of the two wave pictures is correct. Fig. 5 shows the falling portion of the square wave with the triggering pulse again clearly superimposed.

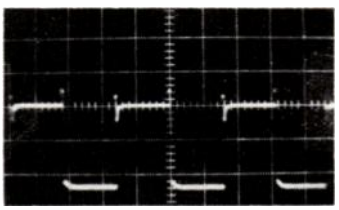


Fig. 3—Square-wave output.

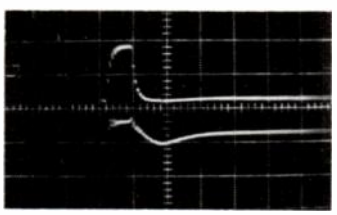


Fig. 4—Square-wave rise.

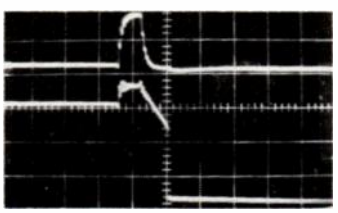


Fig. 5—Square-wave fall.

It can be seen from the photographs that the rise of the square wave is triggered by the beginning of the positive pulse although the oscilloscope is not fast enough to display the actual rise. The pulse also causes slight damped oscillations to begin which probably are due to circuit "ringing" rather than tunnel-diode oscillation *per se*. The negative dip due to the differentiated pulse is clearly shown and is followed by the exponential rise to the final value with a time constant of L/R. In Fig. 5 which shows the triggering of the fall portion of the square wave, it is plain that the differentiated pulse first causes an increased voltage across the diode but that the negative portion drives the tunnel diode down to E_v where extremely rapid switching occurs. This mode of switching differs from the one suggested by Kaenel since his trigger pulse was negative going before positive going, while the differentiated pulse used in this arrangement is exactly opposite.

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* Received by the IRE, September 18, 1961; revised manuscript received, November 8, 1961.

[1] R. A. Kaenel, "One-tunnel-diode flip-flop," PROC. IRE, vol. 49, p. 622; March, 1961.

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Continuous Operation of the $CaWO_4:Nd^{+3}$ Optical Maser*

Continuous operation of a solid-state optical maser has been demonstrated. The system investigated is $CaWO_4$ containing trivalent neodymium, and generation of a continuous coherent electromagnetic wave at 1.065 μ is obtained.

The energy level diagram of the trivalent neodymium ion in $CaWO_4$, derived by combining the infrared fluorescence spectrum[1] with preliminary absorption data, is shown in Fig. 1. The gross features are very similar to Carlson and Dieke's detailed analysis[2] of the energy levels of Nd^{+3} in $LaCl_3$. The principal pumping region for infrared fluorescence lies in the visible around 17,000 cm^{-1}. Stimulated emission can be obtained in several lines in $CaWO_4:Nd^{+3}$. For the continuous operation experiments to be described, we shall be concerned only with the most intense line at 1.065 microns, the line with the lowest threshold for maser action. The terminal state lies some 2000 cm^{-1} above the ground state (Fig. 1).

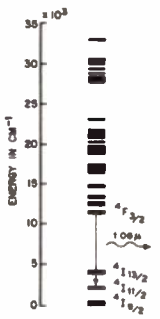


Fig. 1—Energy level diagram of Nd^{+3} in $CaWO_4$.

A schematic of the apparatus constructed to conduct continuous operation experiments is shown in Fig. 2. It consists of an elliptical cylinder, all of whose walls are well polished reflecting surfaces. At the two foci of the ellipse are mounted, respectively, a water cooled General Electric AH6 linear high pressure quartz mercury lamp and the maser crystal.

The method of cooling was evolved in discussions with R. J. Collins. It consists of heat transfer between the crystal and a bath of flowing liquid oxygen that has been precooled to liquid nitrogen temperature in order to eliminate bubbling. A circulating liquid filter of sodium nitrite solution surrounds the dewar to eliminate ultraviolet radiation below 4000 Å. The filter has a negligible effect on the threshold for stimulated emission but is a vital factor for extended continuous operation. An RCA 7102 photomultiplier tube with a 1-mm silicon window is employed for detection of stimulated emission, in conjunction with a dual-beam oscilloscope.

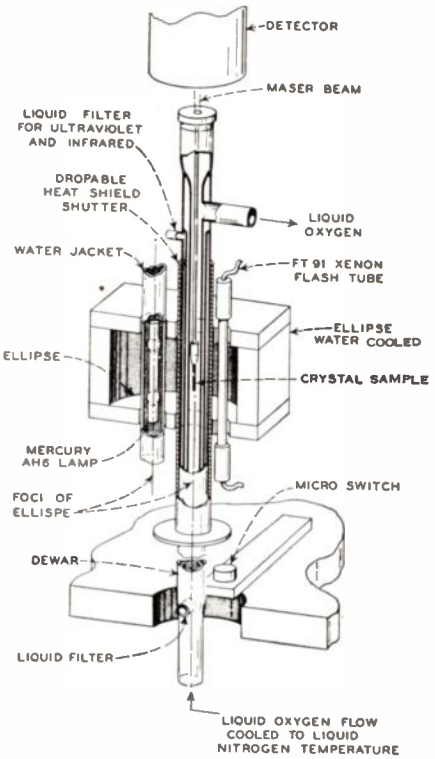


Fig. 2—Schematic of experimental equipment.

The form of optical wavelength resonator used is a rod of length large compared to its diameter with approximately confocal[3,4] spherical ends. These surfaces are made highly reflecting with silver coatings, one end being left partially transmitting (~1 per cent).

A pattern of stimulated emission in continuous operation is shown in Fig. 3. The variation of pump light intensity with time is displayed in the lower trace, maser output in the upper, on a time scale of 5 sec/division. The traces are triggered simultaneously. After 7 sec (to show the zero line of fluorescence and lamp intensity) the lamp is turned on and the crystal is exposed for 12 sec to illumination produced by 1300-w dc power into the lamp. This power level is slightly above the threshold for stimulated emission and maser oscillation occurs. Note that about 5 sec is required for thermal equilibrium to be established. The power level is then lowered to 940 w for 8 sec, yielding fluorescence output only (nearly coinciding with the zero line). The power is then raised to 1600 w, well above threshold, and continuous operation is maintained for about 12 sec. Note that the dc level of maser output is now well above the fluorescence level. Finally, the lamp is turned off. The trace appears to indicate that maser output in continuous operation is not very constant. However, the trace is misleading since the "spikes" are caused by 60-cps ripple on the dc, amounting to about 20 per cent. This is verified by Fig. 4 where the sweep speed has been increased to 5 msec/division (power level 1500 w). Note the zero line of maser output and lamp intensity. The trace clearly shows that maser output follows the 60-cps ripple on the dc, and that "spiking" is not present in continuous operation.

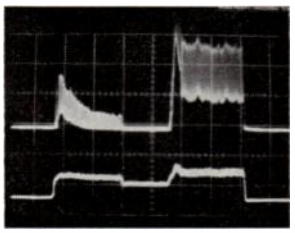


Fig. 3—Continuous operation of the $CaWO_4:Nd^{+3}$ optical maser. Lamp intensity is displayed in the lower trace, maser output in the upper trace. Time scale—5 sec/division. After 7 sec (to show zero line) the lamp is turned on and the maser crystal is exposed for 12 sec to 1300-w dc power into the AH6, a power level slightly above the threshold for stimulated emission. The power level is then lowered to 940 w for 8 sec, yielding fluorescence output only (nearly coinciding with the zero line). The power is then raised well above threshold to 1600 w, continuous operation maintained for about 12 sec, and, finally, the lamp is turned off.

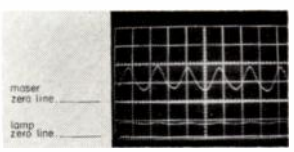


Fig. 4—Detail of continuous operation on a time scale of 5 msec/division. Power input to the lamp 1500 w (lower trace). Note that maser output (upper trace) follows the 60-cps ripple on the dc.

At the 1600-w level of pump power in Fig. 3 the maser output power transmitted through the lightly silvered end of the crystal is about 1-mw ac superimposed on a dc output of about 1 mw. Multiple dielectric layer reflectors in place of silver on the ends of the crystal should significantly raise this figure, as well as lower the pump power required for continuous maser operation.

A more detailed account of the techniques developed for this work and the results obtained has been submitted for publication.

Acknowledgment

The authors are grateful for superb cooperation from many members of the Laboratory. Specifically the able technical assistance of R. A. Thomas, A. E. Di Giovanni, and A. M. Broyer was invaluable. The machining, hand polishing, and aluminizing of the ellipse was ably performed by C. J. O'Neill, J. P. Rudert, and E. M. Kelly. Absorption data was provided by Miss D. M. Dodd. Fruitful and stimulating discussions with C. G. B. Garrett, R. J. Collins, and J. P. Gordon contributed greatly to the success of this work.

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* Received by the IRE, December 14, 1961.

[1] L. F. Johnson and K. Nassau, "Infrared fluorescence and stimulated emission of Nd^{+3} in $CaWO_4$," Proc. IRE, vol. 49, pp. 1704–1706; November, 1961.

[2] E. H. Carlson and G. H. Dieke, "Fluorescence spectrum and low levels of $NdCl_3$," *J. Chem. Phys.*, vol. 29, pp. 229–230; July, 1958. See also E. H. Carlson, "Absorption and Fluorescence Spectra of $NdCl_3$ and States of the Neodymium Ion," Johns Hopkins University, Baltimore, Md., Spectroscopic Rept. No. 16; March, 1960.

[3] G. D. Boyd and J. P. Gordon, "Confocal multimode resonator for millimeter through optical wavelength masers," *Bell Sys. Tech. J.*, vol. 40, pp. 489–508; March, 1961.

[4] A. G. Fox and Tingye Li, "Resonant modes in a maser interferometer," *Bell Sys. Tech. J.*, vol. 40, pp. 453–488; March, 1961.

Defining the Coherence of a Signal*

Before we can discuss Golay's view[1] that "either radiation is coherent, or it is not," we need a definition. The one suggested is not good ("CW radiation which can interfere with indefinitely delayed portions of itself") for reasons connected with the terms "CW," "interfere" and "indefinitely delayed." Furthermore, Golay's views extend to cases where "the duration of the signal is not indefinite."

By attempting to frame a definition we bring out some of the factors. To avoid difficulties connected with spatial phenomena of radiation fields, let us limit attention to a signal voltage $v(t)$ derived from the radiation field by an antenna. Let $v(t)$ be the real part of $F(t)\exp(j\omega t)$, where $F(t)$ is a complex, time-dependent phasor representing $v(t)$. In order to take this step we must choose ω. If we choose it too high, or too low, then pha $F(t)$ will show a net progression in time in one sense or the other, but by a careful choice it may be possible to prevent this.

If we are dealing with a statistically stationary signal, we can contemplate the following quantitative definition of a (complex) degree of coherence $\Gamma(\tau)$, such that

$$\Gamma(\tau) = \frac{\langle F(t)F^*(t+\tau)\rangle}{\langle F(t)F^*(t)\rangle},$$

where asterisks denote complex conjugates and

$$\langle \cdots \rangle \equiv \frac{1}{2T}\lim_{T\to\infty}\int_{-T}^{T}\cdots dt.$$

It will be seen that the net progression of $F(t)$ associated with the arbitrariness in the choice of ω is cancelled out by the identical but opposite retrogression of $F^*(t+\tau)$, and so the result does not depend on the choice of ω. The proposed quantity is a function of the time delay τ, in general diminishing with increasing τ, and might or might not ultimately die out. If the band were narrowed, the degree of coherence would persist longer.

Now if we are not dealing with a statistically stationary signal, the expression given above can be rescued if the time averages assume well-defined values over short averaging times. Denoting such a time average by the notation

$$\langle\langle \cdots \rangle\rangle \equiv \frac{1}{2T}\lim\int_{t-T}^{t+T}\cdots dt$$

and including a factor[2] to compensate for change in strength, we have the complex, epoch-dependent degree of coherence $\Gamma_t(\tau)$, where

$$\Gamma_t(\tau) = \frac{\langle\langle F(t)F^*(t+\tau)\rangle\rangle}{\langle\langle F(t)F^*(t)\rangle\rangle}\,\frac{1}{2}\cdot\left\{\sqrt{\frac{\langle\langle F(t)F^*(t)\rangle\rangle}{\langle\langle F(t+\tau)F^*(t+\tau)\rangle\rangle}}+\sqrt{\frac{\langle\langle F(t+\tau)F^*(t+\tau)\rangle\rangle}{\langle\langle F(t)F^*(t)\rangle\rangle}}\right\}.$$

This expression will depend in general on the epoch t, but has attractive properties. For example, $|\Gamma_t(\tau)|$ can be invariant with epoch under secular changes in the attenuation and phase shift through the medium between the source and the antenna. In general, however, dependence on t and τ would have to be envisaged in discussing the coherence of a nonstationary signal, and one could bring to bear the technique of handling dynamic spectra.

In connection with the "molerun" go no-go test for coherence, consider an ordinary 60-cycle voltage generator which is driven by a gasoline engine and is equipped with means for keeping the envelope voltage in the general neighbourhood of 160 v and the frequency in the neighbourhood of 60 cycles. The complex phasor $F(t)$ would generate a circular ring distribution. Over delays τ of a few cycles, the degree of coherence would be appreciable, but by the time a circularly symmetrical ring distribution had been formed, the degree of coherence would have fallen to a small value.

Any oscillator equipped with means for maintaining phase consistency would show a larger degree of coherence for longer τ, and so for any yes-no definition of coherence based on underlying phase-consistency or ability to interfere, we would hardly wish the gasoline-driven generator, which merely exhibits amplitude consistency, to fall in the "coherent" category. In Golay's test the ring distribution is subjected to further circular smearing by the use of a reference signal that is perfectly monochromatic but offset in frequency. Under this test both gasoline generators and ideal sine wave generators give ring distributions. The ring distribution, therefore, does not qualify as a criterion for coherent signals.

Conversely, the "molehill" test is not a satisfactory indicator of incoherence.

It appears from this discussion that the degree of stationarity of a signal should enter into the framing of a definition of its coherence and that mention of the lapse time τ should be made in statements about the coherence of a signal.

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* Received by the IRE, September 13, 1961.

[1] M. J. E. Golay, "Note on the probable character of intelligent radio signals from other planetary systems," Proc. IRE (Correspondence), vol. 49, p. 959; May, 1961.

[2] R. N. Bracewell, "Radio interferometry of discrete sources," Proc. IRE, vol. 46, pp. 97–105; January, 1958.

Radio Signals from Other Planets*

The 1420-Mc line of hydrogen can hardly be said to be located in a relatively noisy region of the cosmic radio spectrum.[1] The average sky brightness temperature amounts to a few degrees Kelvin. There are some very limited regions in the direction of the center of the galaxy, and to each side along the galactic equator, where the brightness temperature is higher, reaching 50°K or more depending on the size of the antenna. At double the wavelength, these temperatures would be 6 or 7 times higher. In addition to this continuum emission there is line emission, generally fairly close to the nominal hydrogen-line frequency, but also concentrated fairly heavily towards the galactic plane.

However, the planets from which signals are expected must not be so remote as to lie in the galactic plane,[2] and therefore the brightness temperatures in the colder parts of the sky are more appropriate. At present the noise contributions of the receiver and ambient medium are thus more important than the cosmic noise. Consequently no case has been made for avoiding 1420 Mc. There are, however, reasons[2] for not using 1420 Mc or other microwave frequencies, if the distance to the nearest community more advanced than our own is 100 light years rather than the 10 light years investigated by Project Ozma.

Under circumstances where signaling by radio waves seemed reasonable, it would not seem worthwhile to correct out the various Doppler shifts. A stable transmitter on a remote planet would give us a signal from which we might hope to deduce the length of the day and the year, the radius and inclination of the planetary orbit, the radius of the planet, the latitude of the transmitter, and even other parameters. As a first harvest during the interval before a signal could be acknowledged, this would be much more exciting news than a string of prime numbers. The latter would only prove that the designers of high-power transmitters can also count, and is not appropriate to signals in the precontact phase.

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[2] R. N. Bracewell, "Communications from superior galactic communities," *Nature*, vol. 186, pp. 670–671; May 28, 1960.

* Received by the IRE, September 13, 1961.

[1] M. J. E. Golay, "Note on the probable character of intelligent radio signals from other planetary systems," Proc. IRE (Correspondence), vol. 49, p. 959; May, 1961.

Approximate Method of Calculating Electromechanical Coupling Factor II*

In the previous letter[1] by the same title, the author reported a new formula for calculating a coupling factor k from the resonance frequency f_R and the anti-resonance frequency f_A of an appropriate transducer, which is, however, to be applied for the "transverse piezoelectric effect," represented by the length-extensional mode of a bar, the contour-shear mode of a square plate and the radial mode of a disk. Here a different formula will be reported [see (11)] for the "longitudinal piezoelectric effect,"

* Received by the IRE, September 7, 1961.

[1] M. Marutake, "Approximate method of calculating electromechanical coupling factor," Proc. IRE (Correspondence), vol. 49, p. 967; May, 1961.

represented by the thickness-extensional mode of a bar and a plate, and the thickness-shear mode of a plate.

There are essential differences between the two effects. For the longitudinal effect, the wave velocity v is determined by the elastic constant for the constant displacement (s^D or c^D), and the admittance reaches a "minimum," called anti-resonance, when the thickness is an odd multiple of half wavelengths. These properties will be shown in the following calculation of the "impedance" Z:

$$Z = \frac{1}{j\omega C^F} + \sum_{n=1}^{\infty} \frac{j\omega L_n}{1 - \omega^2 L_n C_n}, \qquad (1)$$

where C^F is the "free" capacitance. As an alternative the general equivalent circuit, which is correct in all frequency ranges, is represented by connecting *in series*, C^F, (C_1 and L_1 *in parallel*), (C_2 and L_2 *in parallel*), $\cdots$, (C_∞ and L_∞ *in parallel*).

That (1) is an exact solution is proved as follows: Generally, Z is solved rigorously from the theory of vibration as

$$Z = \frac{1}{j\omega C^{\prime}} \left\{ 1 - k^2 \frac{1}{f(\omega)} \right\}$$

$$= \frac{1}{j\omega C^F} \left\{ 1 + \frac{k^2}{1 - k^2} \left(1 - \frac{1}{f(\omega)} \right) \right\}, \qquad (2)$$

where $C^{\prime}$ is the "clamped capacitance," *i.e.*,

$$C^{\prime} = (1 - k^2) C^F, \qquad (3)$$

and

$$f(\omega_{An}) = 0 \qquad (n = 1, 2, 3, \cdots), \qquad (4)$$

where ω_{An} is the antiresonance angular frequency of the nth overtone mode. By Mittag-Leffler's theorem,

$$1 - 1/f(\omega) = \sum_{n=1}^{\infty} p_n \left\{ 1 - \frac{1}{1 - (\omega^2/\omega_{An}^2)} \right\}$$

$$= \sum_{n=1}^{\infty} \frac{-p_n(\omega^2/\omega_{An}^2)}{1 - (\omega^2/\omega_{An}^2)}, \qquad (5)$$

where p_n is a constant, and particularly

$$p_1 = p. \qquad (6)$$

Substituting (5) in (2), we obtain

$$\left. \begin{aligned} C_n &= \frac{1}{p_n} \frac{1 - k^2}{k^2} C^F, \\ L_n &= 1/(\omega_{An}^2 C_n). \end{aligned} \right\} \qquad (7)$$

Now, at the frequency near the anti-resonance of the fundamental mode,

$$\omega L_n \ll \frac{1}{\omega C_n} \ll \frac{1}{\omega C_1} \simeq \omega L_1$$

$$(n = 2, 3, 4, \cdots), \qquad (8)$$

and we can approximate as

$$Z = \frac{1}{j\omega C^F} + \frac{j\omega L_1}{1 - \omega^2 L_1 C_1}. \qquad (9)$$

As an alternative, the equivalent circuit is represented by connecting *in series* C^F and (C_1 and L_1 *in parallel*). Then

$$C^F/C_1 = 1/r = (f_A^2 - f_R^2)/f_R^2, \qquad (10)$$

and from (7) and (6)

$$k^2/(1 - k^2) = 1/(pr)$$

$$= (1/p)(f_A^2 - f_R^2)/f_R^2. \qquad (11)$$

This is a new approximate formula for the longitudinal effect, and it is accidentally identical with the formula which has been used for the "transverse effect" until now.

In Table I, the calculated values from (11) are compared with true values calculated from following (12),[2] in the case of the thickness-extensional mode of a bar where $k_{33} = 4\pi d_{33}^2/(\epsilon_{33}^T \cdot s_{33}^E)$ and $p = 8/\pi^2$.

$$k_{33}^2 = \frac{\pi}{2} \left(\frac{f_R}{f_A} \right) \tan \left(\frac{\pi}{2} \frac{\Delta f}{f_A} \right), \qquad (12)$$

where $\Delta f = f_A - f_R$.

TABLE I
CALCULATED VALUES OF k_{33}

Calculated from / $\Delta f/f_R$	Eq. (11)	Eq. (12)
0.01	0.156	0.156
0.1	0.454	0.453
0.3	0.678	0.677
0.5	0.779	0.778
1.0	0.887	0.886
5.0	0.989	0.989
∞	1.000	1.000

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[2] D. A. Berlincourt, C. Cmolik, and H. Jaffe, "Piezoelectric properties of polycrystalline lead titanate zirconate compositions," PROC. IRE, vol. 48, pp. 220–229; February, 1960.

Results Obtained with Tunnel-Diode Superregenerative Receivers*

A tunnel-diode regenerative amplifier may be superregenerated by making the forward-biasing voltage on the diode a periodic function of time. This has been done with good results at 100, 325, and 930 Mc.

Tabulated below are the results obtained. MDS means *minimum detectable signal* as determined by listening tests when 30 per cent amplitude modulation at 1000 cps is used. An RF signal generator having an internal impedance of 50 Ω was used.

Operating Frequency Mc	MDS μv	MDS dbm	Quench Frequency kc	Gain db
100	1	−107	35	not measured
325	0.4	−115	50	54
930	0.3	−117	50	32

Fig. 1 shows diagrammatically the receiver configuration employed. Shown in Figs. 2(a)–(c) are the quench waveforms used. The pips are due to build-up of the RF signal.

* Received by the IRE, December 12, 1961.

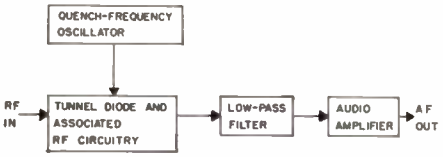


Fig. 1—Tunnel-diode receiver.

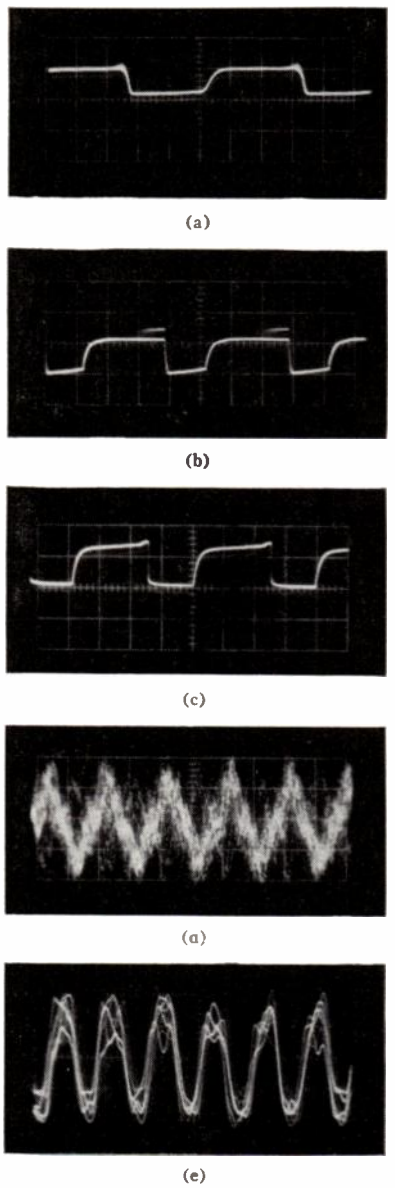


Fig. 2—(a)–(c) Quench waveforms, dc level is indicated by horizontal line closest to bottom of waveform, 5 μsec/cm, 50 mv/cm; (a) 100 Mc; (b) 325 Mc; (c) 930 Mc. (d)–(e) Detected 2-μv AM signal, 500 μ sec/cm; (d) 325 Mc; (e) 930 Mc.

A 2-μv signal was applied to the 325-Mc receiver. 30 per cent amplitude modulation of the carrier at 1000 cps was used. The detected signal is shown in Fig. 2(d). Most of the blur in the photograph is caused by the presence of quench frequency components, due to insufficient filtering. The bandwidth of the audio amplifier was approximately 3 kc.

A 2-μv signal similar to the one used for the 325-Mc receiver was applied to the 930-Mc receiver. A photograph of the detected

signal for the 930-Mc receiver is shown in Fig. 2(e). A 3-kc audio amplifier bandwidth was used. In contrast to Fig. 2(d), Fig. 2(e) shows no evidence of quench-frequency components in the output. Better filtering was used.

The RF portion of the 100-Mc receiver uses a lumped-parameter resonator. Both the 325- and 930-Mc receivers use coaxial-cavity resonators. All of the receivers are easy to work with. Results are readily reproducible.

Bandwidths have not been precisely determined. They are, however, quite narrow.

Using a three foot whip antenna together with the 100-Mc receiver, it was possible to receive commercial FM broadcasts. Good sensitivity and selectivity were obtained. Fidelity was fair.

From results obtained to date, it appears that tunnel-diode superregenerative receivers will find wide application in radar and communications systems.

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Magnetoresistance Multiplier with Higher Gain*

The Hall effect in semiconductors has been used for multiplication and, consequently, for measuring an electric power.[1-3] Now, it is proved that the magnetoresistance effect in semiconductors can also be used as a multiplier with much higher gain and simpler construction than the corresponding Hall effect.

In general, the resistance of a semiconductor element shows an increase at first proportional to the square of and then linear with the applied magnetic field.

The principle of the magnetoresistance multiplier described here is based on the fact that the time-average of the voltage drop across a magnetoresistive semiconductor is always proportional to the net product of a current passed through it and a magnetic field or a magnetizing current passed through a magnetizing coil, when the semiconductor is properly biased magnetically in the linear region of the magnetoresistance characteristics.

Experiments were carried out for its application to the power measurements at both ac 50 cps and dc, with the use of a Halltron magnetic circuit MC-1 made by Ohio Semiconductor Inc., Columbus, Ohio.

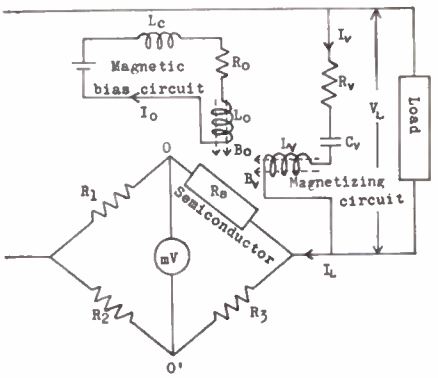
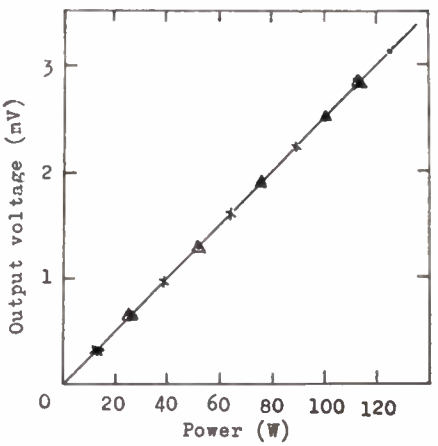


Fig. 1—Magnetoresistance wattmeter.



Power factor ●— 100 per cent
×— 70 per cent leading
△— 70 per cent lagging

Fig. 2—Characteristics of a magnetoresistance wattmeter at ac.

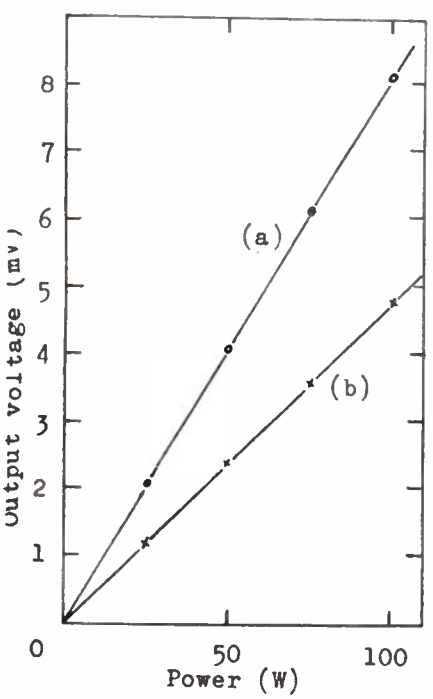


Fig. 3—Comparison between magnetoresistance wattmeter and Hall-effect wattmeter at dc. (a) Magnetoresistance wattmeter. (b) Hall-effect wattmeter.

The Halltron semiconductor element (presumably of InAs polycrystal) was used as a magnetoresistance element, with its Hall terminals shorted. The schematic diagram of the arrangement is shown in Fig. 1. A choke L_C and a resistor R_0 in the bias circuit are to prevent an ac current induction by the magnetizing current I_V. The bias current I_0 was set at 24 ma in reference to the magnetoresistance characteristics. A capacitor C_V and a resistor R_V in the magnetizing circuit were chosen so as to give a magnetic flux B_V in phase with the load voltage V_L and a very high impedance in the circuit compared with the load. A bridge circuit is necessary only at dc operation, and requirements for the arms are $R_1 \gg R_S$, $R_1 \gg R_2$, and $R_{S0}R_2 = R_1R_3$. In this case, C_V should be omitted.

Under the proper condition of the phase adjustment at ac operation, the resistance of the semiconductor can be expressed by

$$R_S = R_{S0} + AV_0 \sin \omega t,$$

where R_{S0} is the resistance of the semiconductor under the dc magnetic bias only, $V_0 \sin \omega t$ is the load voltage, and A is a proportionality constant. Then, the voltage drop across the semiconductor is

$$v = (R_{S0} + AV_0 \sin \omega t) \cdot BI_0 \sin (\omega t + \phi),$$

where $I_0 \sin (\omega t + \phi)$ is the load current, $\cos \phi$ is a power factor of the load and B is a proportionality constant. And its time average is

$$\bar{v} = AB \frac{V_0 I_0}{2} \cos \phi.$$

Therefore, the active power can be measured by the time-average voltage across the semiconductor element.

Fig. 2 shows the experimental results on various kinds of load at ac. A comparison with the corresponding Hall-effect device at dc (Hall voltage measured across the Hall terminals of the Halltron under the same condition) is shown in Fig. 3. The instability in the bias current does not affect the measurement at ac operation.

The advantages of a magnetoresistance multiplier over a Hall-effect one may be as follows.

1) Much simpler construction of the semiconductor element. Only two terminals are required instead of four.
2) Consequently no need for ac or dc balancing between the terminals.
3) Less accuracy needed in the geometric arrangement against the magnetic field.
4) No trouble due to the rectifier action because of the large area of the contacts.
5) No trouble due to the local heating effect in an electromagnetic field.
6) Much higher gain in the output.

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* Received by the IRE, September 27, 1961.

[1] H. E. M. Barlow, "The Application of the Hall effect in a semiconductor to the measurement of power in an electromagnetic field," *Proc. IEE*, vol. 102B, pp. 179–185; March, 1954.

[2] H. E. M. Barlow, "The design of semiconductor wattmeters for power-frequency and audio-frequency applications," *Proc. IEE*, vol. 102B, pp. 186–191; March 1954.

[3] H. E. M. Barlow and S. Katoaka, "The Hall effect and its application to power measurement at 10 G c/s," *Proc. IEE*, vol. 105B, pp. 53–60; January, 1958.

The November, 1961, issue carried Correspondence entitled "The Delta Function and the Fourier Integral," which raised certain questions about the delta function. The number of readers who responded was substantial. Since it was not possible to publish all the letters submitted, which presented different approaches to the subject, two were chosen as being representative. In publishing these letters below, together with the author's reply, we wish to express our appreciation to all those who responded.—The Editor

The Delta Function*

From the standpoint of applications, the relevant property of the delta "function" is the so-called "sifting" property,

$$\int_{-\infty}^{\infty} \delta(t-\tau)f(t)dt = f(\tau), \qquad (1)$$

where $f(t)$ is an arbitrary continuous function defined on $(-\infty, \infty)$. Unfortunately, there exists no locally integrable point function possessing this desirable property so that the delta function might more properly be rechristened the "delta fiction." Notwithstanding this very real objection, the usual formal manipulations involving the δ function are a convenient shorthand for the applied worker and yield correct results in a number of cases sufficient to embarrass critics of the method. Moreover, respectability can be gained by an appeal to the theory of distributions or some other generalization of the function concept; in this regard, the expository article by Erdélyi [2] gives an excellent introduction (for the engineer) to the various rigorous methods for handling the δ-function and also contains a good list of references to the mathematical literature on the subject.

In a recent letter [1] McMurtrey raises two questions concerning the use of delta functions in Fourier integrals. One question involves the apparent ambiguity in the value of the integral

$$\frac{1}{2\pi}\int_{-\infty}^{\infty} \delta(\omega)e^{i\omega t}d\omega$$

under the change of variable $\omega = 2\pi f$, while the other is concerned with the appropriateness of the "identity,"

$$\delta(t) = \int_{-\infty}^{\infty} \cos 2\pi ftdf.$$

The first question is most easily resolved by introducing the unit step function $u(t)$ defined by

$$u(t) = \begin{cases} 1 & \text{for } t > 0 \\ 1/2 & \text{for } t = 0 \\ 0 & \text{for } t < 0 \end{cases}$$

and writing the questioned integral (eq. (6) of [1]) in Stieltjes form as

$$\frac{1}{2\pi}\int_{-\infty}^{\infty} e^{i\omega t}du(\omega).$$

This integral exists in the rigorous sense and has the value $1/2\pi$. Moreover the change of variable $\omega = 2\pi f$ may now be effected legitimately ([3] p. 19) yielding

$$\frac{1}{2\pi}\int_{-\infty}^{\infty} e^{2\pi i f t}d[u(2\pi f)].$$

But $u(2\pi f) \equiv u(f)$ so that the latter integral has the value $1/2\pi$ as before. In terms of the δ-function, this suggests that the proper heuristic formula should be

$$\int_{-\infty}^{\infty} f(t)\delta(\alpha t - \tau)dt = \frac{1}{\alpha}f\left(\frac{\tau}{\alpha}\right), \qquad (2)$$

where $f(t)$ is continuous and α is any fixed real number. In particular, for $\tau = 0$, (2) becomes

$$\int_{-\infty}^{\infty} f(t)\delta(\alpha t)dt = \frac{1}{\alpha}f(0), \qquad (3)$$

so that $\delta(\alpha t)$ and $\delta(t)$ do not have equivalent properties with regard to the evaluation of integrals unless $\alpha = 1$.

Secondly, the proposed formal equality

$$\delta(t) = \int_{-\infty}^{\infty} \cos 2\pi ftdf \qquad (4)$$

can be given a pseudorationalization by the following argument.

Let

$$\phi_n(t) = \frac{\sin 2\pi nt}{\pi t} \qquad \text{for } n = 1, 2, \cdots.$$

Then, it is easily verified that

$$\int_{-\infty}^{\infty} \phi_n(t)dt = 1$$

independently of n; furthermore, from the theory of Fourier integrals ([4] p. 115), if $f(t)$ is continuous and absolutely integrable on $(-\infty, \infty)$, then

$$\lim_{n\to\infty} \int_{-\infty}^{\infty} f(t)\phi_n(t)dt$$

$$= \lim_{n\to\infty} \int_{-\infty}^{\infty} f(t)\frac{\sin 2\pi nt}{\pi t}\,dt = f(0).$$

Thus, the functions

$$\{\phi_m(t)\}_{m=1}^{\infty}$$

may be thought of as approaching $\delta(t)$ in some sense as $n \to \infty$; or, heuristically,

$$\delta(t) = \lim_{n\to\infty} \frac{\sin 2\pi nt}{\pi t}. \qquad (5)$$

But

$$\int_{-n}^{n} \cos 2\pi ftdf = \frac{\sin 2\pi nt}{\pi t},$$

so that we have (again formally, since the indicated limit does not exist)

$$\delta(t) = \lim_{n\to\infty} \int_{-n}^{n} \cos 2\pi ftdf. \qquad (6)$$

Formulas such as (4) have been used to good effect in the engineering literature (*e.g.*, [5], p. 201); such formulas can also be given a rigorous interpretation within the framework of the theory of distributions [2].

Note that the emphasis here has been on the retention of property (1) while giving up such inadequate pointwise descriptions as saying that $\delta(t)$ is zero everywhere except at the origin where it has an infinite value. It is essentially this observation which suggests regarding δ as a functional, and an approach based on this idea can be made rigorous (see [2] for discussion and details).

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References

[1] A. W. McMurtrey, "The delta function and the Fourier integral," Proc. IRE (Correspondence), vol. 49, pp. 1702–1703; November, 1961.
[2] A. Erdélyi, "From delta functions to distributions," in "Modern Mathematics for the Engineer," E. F. Beckenbach, Ed., McGraw-Hill Book Co., Inc., New York, N. Y., pp. 5–50; 1961.
[3] D. V. Widder, "The Laplace Transform," Princeton University Press, Princeton, N. J.; 1946.
[4] F. G. Tricomi, "Vorlesungen Über Orthogonalreihen," Springer-Verlag, Munich, Germany; 1955.
[5] J. A. Aseltine, "Transform Method in Linear System Analysis," McGraw-Hill Book Co., Inc., New York, N. Y.; 1958.

* Received by the IRE, November 20, 1961.

The Delta Funtion and the Fourier Integral*

In a recent note,[1] McMurtrey poses two questions about the δ-function in the form of "paradoxes." They are:

1) Is the value of

$$f(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} \delta(\omega)e^{i\omega t}d\omega \text{ one or } \frac{1}{2\pi}?$$

2) Is

$$\delta(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty} \cos \omega t d\omega$$

a legitimate equation?

The answers are easily obtained (and without resorting to the formalism of the theory of distributions) as long as one sticks to a prescribed set of ground rules regarding the use of the δ-function. A natural source for such a set of rules is Dirac,[2] who, after all, invented the δ-function. Dirac defines $\delta(x)$ by the set of relations

$$\delta(x) = \infty \quad \text{at} \quad x = 0 \qquad (1a)$$

$$\delta(x) = 0 \quad \text{for} \quad x \neq 0 \qquad (1b)$$

$$\int_{-\infty}^{\infty} \delta(x)dx = 1 \qquad (1c)$$

$$\int_{-\infty}^{\infty} f(x)\delta(x-a)dx = f(a) \qquad (1d)$$

(for any $f(x)$ for which $f(a)$ is defined).

* Received by the IRE, December 6, 1961.

[1] A. W. McMurtrey, "The delta function and the Fourier integral," Proc. IRE (Correspondence), vol. 49, pp. 1702–1703; November, 1961.

[2] P. A. M. Dirac, "The Principles of Quantum Mechanics," The Clarendon Press, Oxford, England, 4th ed.; 1958.

Hence, by the set of equations (1) above, the relation

$$f(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\delta(\omega)e^{i\omega t}d\omega. \tag{2}$$

leads unambiguously to the result that $f(t)=1/2\pi$ and cannot be interpreted to yield $f(t)=1$. McMurtrey's error in his equation (9) resides in the fact that he neglects to integrate with respect to the δ-function argument, and, according to (1c) and (1d), this is an implicit requirement. It should be pointed out that McMurtrey incorrectly replaces the argument of the δ-function with a number before integrating. For example,

$$\int_{-\infty}^{\infty}\delta(0)dt$$

is not defined since $\delta(0)$ is infinite.

McMurtrey's second "paradox" concerns the relation

$$\delta(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\cos\omega t d\omega. \tag{3}$$

He points out that if one makes the substitution $t=1$ in (3), one obtains the relation

$$\delta(1) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\cos\omega d\omega \tag{4}$$

which, by the definition of the δ-function, should converge to zero. He points out that (4) does not converge to any number, but neglects to mention the criterion of convergence to which he refers. In the classical sense, he is correct. However, the δ-function is not a classical function. Indeed, if one insists on using the word "function" in connection with it at all, it should be tacitly understood that one means a *symbolic* function, as has been pointed out by Friedman.[3] It is in this light that (3) and (4) must be considered. That is, once one allows the use of δ-functions, one must stick to the ground rules. Therefore, the definition of the δ function becomes overriding in (3) and (4) and

$$\frac{1}{2\pi}\int_{-\infty}^{\infty}\cos\omega d\omega$$

is *defined* as being zero due to (1b), even though it does not converge to zero in the classical sense. The validity of (3) then resides in its formal attainment coupled with adherence to the rules set up in (1). As an example, consider the following:

$$I = \int_{t=-\infty}^{\infty} f(t)\left[\frac{1}{2\pi}\int_{\omega=-\infty}^{\infty}\cos\omega t d\omega\right]dt$$

$$= \int_{\omega=-\infty}^{\infty}\left[\int_{t=-\infty}^{\infty} f(t)\cos\omega t dt\right]\frac{d\omega}{2\pi}. \tag{5}$$

Using the identity

$$\cos\omega t = \frac{e^{i\omega t}+e^{-i\omega t}}{2} \tag{6}$$

we obtain

$$I = \frac{1}{2}\int_{-\infty}^{\infty}[F(-i\omega)+F(i\omega)]\frac{d\omega}{2\pi} \tag{7}$$

$$= \frac{1}{2}\mathfrak{I}^{-1}\{F(i\omega)\}\big|_{t=0} + \frac{1}{2}\mathfrak{I}^{-1}\{F(-i\omega)\}\big|_{t=0}. \tag{8}$$

where $\mathfrak{I}^{-1}$ denotes the inverse Fourier transform.

$$\therefore\ I = \tfrac{1}{2}f(0)+\tfrac{1}{2}f(0) = f(0). \tag{9}$$

But

$$f(0) = \int_{-\infty}^{\infty} f(t)\delta(t)dt \tag{10}$$

$$\therefore\ \delta(t) = \frac{1}{2\pi}\int_{-\infty}^{\infty}\cos\omega t d\omega \tag{11}$$

as before. This example shows that no difficulty is encountered in the use of (3) as an identity.

Similar questions to those discussed here can be raised regarding the higher-order δ-functions (doublets, triplets, etc.). These questions are resolved in much the same manner as those discussed in this note.

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[3] B. Friedman, "Principles and Techniques of Applied Mathematics," John Wiley and Sons, Inc., New York, N. Y.; 1957.

Comments on the Delta Function and the Fourier Integral*

There is no question of the great usefulness of the delta function in its diverse applications to analysis in engineering and physics. On the other hand, there is the possibility of some latent difficulties in a too-literal acceptance of the usually stated properties of the delta function. In particular, in the first paragraph of my original letter, one of the properties given the delta function was essentially that given in Weiss's letter as (1a)–(1c),

$$\delta(x) = \infty \quad \text{at} \quad x = 0 \tag{1a}$$

$$\delta(x) = 0 \quad \text{for} \quad x \neq 0 \tag{1b}$$

$$\int_{-\infty}^{\infty}\delta(x)dx = 1. \tag{1c}$$

Eq. (1c) is incompatible with (1a) and (1b). On any interval not containing the origin, the integral of $\delta(x)$ is clearly zero; however, on any interval containing the origin, $\delta(x)$ is not Riemann integrable and its Lebesgue integral is zero.

Furthermore, if we accept (1a) and (1b), it is an inescapable conclusion that the function $\delta(ax)$ is in fact identical with $\delta(x)$. That is, according to (1a) and (1b) for $\delta(x)$, it would then follow that

$$\delta(ax) = 0, \quad \text{for} \quad ax \neq 0 \tag{2}$$

but

$$ax = 0 \tag{3}$$

is simply

$$x = 0, \quad \text{for} \quad a \neq 0 \tag{4}$$

and

$$\delta(ax) = \infty \quad \text{at} \quad ax = 0 \quad (\text{or } x = 0). \tag{5}$$

Therefore, in summary, based on (1a) and (1b), both $\delta(x)$ and $\delta(ax)$ are zero everywhere except at $x=0$; and at $x=0$, both $\delta(x)$ and $\delta(ax)$ are $\delta(0)$. This is the inherent danger of a too-literal acceptance of (1a)–(1c), because $\delta(x)$ and $\delta(ax)$ thus appear to be identical, and is of course the source of the apparent "paradox" which arose from (7) and (9) of my original letter.

I am grateful to Weiss for his observation on the incorrectness of the integral

$$\int_{-\infty}^{\infty}\delta(0)dt. \tag{6}$$

Instead of (6), the limits on the integral should be changed according to the description afforded by (1a) and (1b). Since $\delta(x)$ is zero everywhere except at the origin, the limits of integration should be reduced accordingly to include only the interval on which $\delta(x)$ is nonzero, so that the integral

$$\int_{-\infty}^{\infty}\delta(t)dt \tag{7}$$

immediately reduces literally to

$$\int_{0}^{0}\delta(0)dt = \delta(0)\int_{0}^{0}dt = (\infty)\cdot(0) \tag{8}$$

which is quite generally regarded as meaningless even in the extended real-number system that includes the symbol ∞ (infinity). Thus (8) is an example of a too-literal acceptance of (1a) and (1b), and it shows that (1c) contradicts (1a) and (1b).

Brown, in his letter, refers to the delta function as the "delta fiction," which is the description reportedly originated by no less than John Von Neumann. Brown's use of the Stieltjes integral is exactly what I had in mind as a rigorous counterpart of the delta function when I wrote the original letter. As Brown shows, in terms of the unit step function, the two Stieltjes integrals

$$\frac{1}{2\pi}\int_{-\infty}^{\infty}e^{i\omega t}du(\omega) \tag{9}$$

and

$$\frac{1}{2\pi}\int_{-\infty}^{\infty}e^{i2\pi ft}du(2\pi f) \tag{10}$$

both yield the unambiguous answer $1/2\pi$, because $u(2\pi f)\equiv u(f)$. It should be emphasized that the Stieltjes-integral approach is not a rigorous justification of the delta function, but rather the delta function is a heuristic interpretation of the rigorous Stieltjes integral.

One way, perhaps, to appraise both $\delta(x)$, as described by (1a)–(1c), and the integral in question

$$\delta(t) = \int_{-\infty}^{\infty}\cos 2\pi ft df \tag{11}$$

is to say that both expressions have well-

* Received by the IRE, December 19, 1961.

understood properties in mathematical physics and mathematical engineering, but the same properties do not carry over intact into mathematical mathematics.

In connection with Brown's comments in his last paragraph, on the inadequate description of $\delta(x)$ and on the suggestion of regarding δ as a functional, it seems that the single most desirable property of $\delta(x)$ is expressed by the so-called sifting integral:

$$f(a) = \int_{-\infty}^{\infty} f(x)\delta(x-a)dx. \qquad (12)$$

This integral has to be accepted on the same heuristic basis as $\delta(x)$ itself. However, in the sense of a functional whose domain is a set of functions and whose range is a set of numbers, the desirable property of the sifting integral could be expressed as

$$F_a[f(x)] = f(a). \qquad (13)$$

The functional is F_a, $f(x)$ is one of the functions in the set which constitutes the domain of F_a, and $f(a)$ is the particular number in the range of F_a which corresponds to the particular function $f(x)$; in other words, the image of $f(x)$ under F_a is $f(a)$. Then the sifting integral of (12) can be considered to be a heuristic symbolic representation of (13).

The heuristic equation (12) can be given rigorous meaning most simply by again using the Stieltjes Integral. Instead of (12), the equation is written as

$$\int_{-\infty}^{\infty} f(x)du(x-a) = f(a), \qquad (14)$$

where $u(x)$ is the unit step function mentioned previously, and the desired result is obtained in a perfectly straightforward manner.

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Infrared Oscillations from $CaF_2:U^{+3}$ and $BaF_2:U^{+3}$ Masers*

The transient response of a three-level optical maser has recently been investigated by Kaplan and Zier.[1] They treat the active ions in terms of a density matrix and show that when 1) the cavity losses are small, 2) the density of atoms in the ground state is fixed, and 3) the terminal state is empty, the set can be reduced to the following coupled equations previously suggested by Statz and deMars:[2]

$$\dot{n} = bP - n/\tau - n\alpha N/\tau, \qquad (1)$$

$$\dot{N} = \alpha nN/\tau - \beta N, \qquad (2)$$

where n is the total number of pumped ions and N is the number of coherent photons in the cavity. The terms in (1) represent optical ion pumping, spontaneous emission and the rate of stimulated emission. In (2), $\alpha nN/\tau$ is the rate of increase of stimulated photons and βN accounts for their losses. In this notation, $n_0(=\beta\tau/\alpha)$ is the minimum number of upper-state ions for oscillation to occur. This density is obtained when the pump power $P=P_0(=\beta/\alpha b)$.

Eqs. (1) and (2) are nonlinear and can lead to large signal instabilities. A stable, steady-state solution exists when $n=n_0$ and $N=N_0\equiv(P/P_0-1)/\alpha$. Near these steady-state values, (1) and (2) can be linearized by setting $n=n_0+\Delta n$ and $N=N_0+\Delta N$—where $\Delta n/n_0,\ \Delta N/N_0 \ll 1$. Substitution of these values yields

$$(\dot{\Delta N}) + (P/P_0\tau)(\Delta N) + (\beta/\tau)(P/P_0-1)(\Delta N) = 0. \qquad (3)$$

The solution to (3) for $\beta\tau \gg 1$ and $P=$const. is

$$\Delta N \approx \exp(-t/T \pm i\omega t), \quad \text{where } T^{-1} = P/2P_0\tau \text{ and}$$
$$\omega = [(P/P_0-1)(\beta/\tau)]^{1/2}. \qquad (4)$$

This shows that ΔN approaches the steady state by means of damped oscillations of frequency ω, and damping time T.

Damped oscillations from a $CaF_2:U^{+3}$ maser operated near threshold at 77°K are shown in Figs. 1 and 2. Similar oscillations have been observed from a $BaF_2:U^{+3}$ maser operated under the same conditions. In Fig. 1, the flash lamp output (as viewed by a photomultiplier) and the maser output at 2.5 μ, measured by an InSb photomagnetoelectric cell, are simultaneously displayed on a dual-beam Tektronix oscilloscope. Maser oscillations are first observed ~90 μsec after the lamp has reached peak intensity. This is due to the finite fluorescent lifetime of 60 μsec. From (1) the threshold density n_0 is obtained when $P[1-\exp-(t/\tau)]/P_0=1$. A ratio t/τ of 90/60 leads to a value $P/P_0=1.3$, which is consistent with experiment.

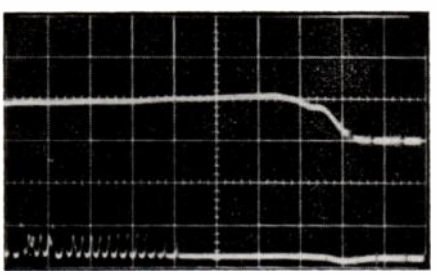


Fig. 1—Uniform oscillations obtained from a $CaF_2:U^{+3}$ optical maser operated at 77°K and pumped slightly above threshold. The time axis runs from right to left. The upper trace is the optical light pump viewed by a photomultiplier. The lower trace shows the maser oscillations. The time interval between uniform pulses is 7.2 μsec.

The uniform oscillations of Fig. 1 have a frequency of 1.4×10^5 sec^{-1}. The frequency computed from (4) is 1.2×10^5 sec^{-1} by using a ratio $P/P_0=1.2$—computed from the light pulse curve and a value $\beta=2\times10^8$ sec^{-1}—computed from $\beta=V(1-R)/L$ where $R(=0.96)$ is the reflection coefficient, $L(=4.3$ cm) is the length of the maser cavity, and $V(=2.1\times10^{10}$ cm/sec) is the velocity of light in the cavity. This agreement is considered good, for (4) gives the frequency dependence near the steady state which is somewhat greater than that of the initial oscillations. Computer solutions of (1) and (2) give close agreement with experiment.

The Q of the cavity can be determined from $Q=2\pi f/\beta$ where f is the frequency of the maser output. This relation yields $Q=2.7\times10^6$. The general expression for Q is $Q^{-1}=Q_r^{-1}+Q_d^{-1}+Q_c^{-1}$ where the subscripts r, d, and c denote losses due to reflection, diffraction, and cavity losses (associated with end faces, strain, inhomogeneities, etc.). Fox and Li[3] have recently shown that diffraction losses from an optical maser cavity should be negligible. The fact that β is determined largely by reflection losses—in the preceding analysis—shows that cavity losses for these masers[4] are also negligible.

In Fig. 2 the trace is increased to 10 μsec/div in order to examine the output pulses in more detail. The average damping time of the uniform pulse train is ~150 μsec. This time is consistent with $T=2P_0\tau/P\sim144$ μsec. Each picture in Fig. 2 is the output from the maser operated at 1 minute intervals. It is interesting to note: 1) the initial uniformity of the pulse train, 2) that one less "clean" pulse is obtained in going from (a) to (c), and 3) that interaction of modes interrupts the uniform oscillations. These effects appear to be due to lattice heating. For example, the thermal diffusivity of CaF_2 (doped with 0.1 per cent U^{+3}) is estimated to be $\sim2\times10^{-2}$cm^2/sec. Hence, thermal energy travels a mean distance of 1.0 cm in 1.0 minute. The fewer "clean" pulses in Fig. 2(b) and (c) can be caused by a higher

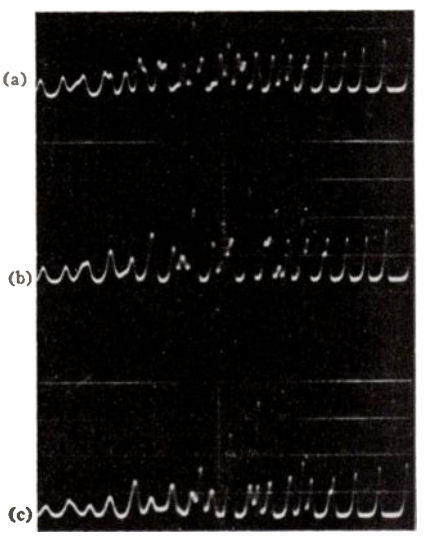


Fig. 2—Oscillations from a $CaF_2:U^{+3}$ optical maser operated at 77°K with a time interval of 1 minute between (a), (b), and (c). The trace is delayed 180 μsec by triggering the oscilloscope by the first maser pulse. The maser had been pumped several times prior to the operation shown in (a). The sweep time is 10 μsec/cm and runs from right to left.

* Received by the IRE, December 18, 1961. This material was presented by one of the authors (H. A. Bostick) in an invited paper at the WESCON Convention, San Francisco, Calif., August 25, 1961.

[1] J. I. Kaplan and R. E. Zier, "A Model for Transient Oscillations in a Three-Level Optical Maser," to be published.

[2] H. Statz and G. A. deMars, "Quantum Electronics," C. H. Townes, Ed., Columbia University Press, New York, N. Y.; 1960.

[3] A. G. Fox and T. Li, "Resonant modes in a maser interferometer," *Bell Sys. Tech. J.*, vol. 40, pp. 453-488; March, 1961.

[4] These materials were obtained from Optovac, Inc., and fabricated into maser cavities by Frank Cooke, Inc. Both of these companies are located in North Brookfield, Mass.

initial cavity temperature. To obtain CW operation, it is obvious that the cavity temperature will have to be kept constant.

Analysis of Figs. 1 and 2 leads us to believe: 1) the initial pulse train is either one mode or a small number of coupled modes, and 2) that a second mode becomes resonant as the cavity temperature is raised. It is of interest to note that the fluorescent line width of $CaF_2:U^{+3}$ at 77°K is a broad 64 cm^{-1} and increases to about 200 cm^{-1} at room temperature. This indicates that materials with large fluorescent line widths can be operated as optical masers with uniform output.

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[5] Operated with support from the U. S. Army, Navy and Air Force.

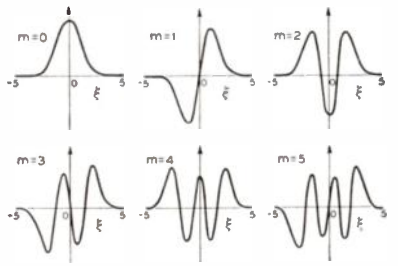


Fig. 1—Relative amplitude variations along one transverse coordinate (ξ) of the six lowest-order modes of a concave-reflector cavity.

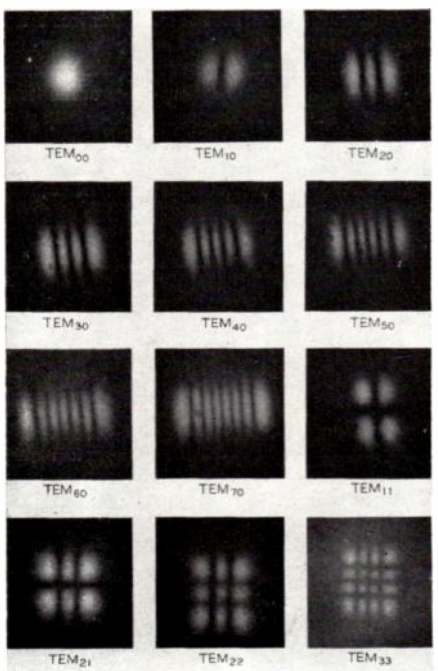


Fig. 2—Beam patterns of some pure modes of a concave-reflector cavity. All beams are polarized horizontally. (Superimposed ring patterns are due to interference between surfaces of the mirror, infrared filter, and window of the image convertor.)

Visual Display of Isolated Optical-Resonator Modes*

The resonant fields of cavities with concave spherical reflectors have been shown[1] to consist of TEM_{mnq} modes, with transverse (x, y) amplitude variations given approximately by

$$E = f_m(x)g_n(y) \tag{1}$$

with

$$f_m(x) \sim H_m(x\sqrt{2}/w_s') \exp\left[-(x/w_s')^2\right] \tag{2}$$

and

$$g_n(y) \sim H_n(y\sqrt{2}/w_s') \exp\left[-(y/w_s')^2\right]. \tag{3}$$

Here H_m is the mth Hermite polynomial and w_s' the "spot size" of the fundamental (0, 0) mode, defined by

$$w_s' = \left(\frac{d\lambda}{\pi}\right)^{1/2}\left[2\frac{d}{b'} - \left(\frac{d}{b'}\right)^2\right]^{-1/4} \tag{4}$$

where λ is the wavelength, d the reflector spacing and b' their common radius of curvature. The transverse field distributions f_m, g_n are recognizable as the Hermite orthogonal functions, which appear in quantum mechanics as the wave functions of the harmonic oscillator[2] (Fig. 1).

When concave reflectors are used as a maser cavity at optical wavelengths, the output beam pattern generally consists of many such modes, superimposed on one another. Photographs of a typical multimode pattern have been published recently.[3] They were obtained by means of an image convertor in the output beam of a helium-neon maser, operating near 11,500 Å. In this case, two concave mirrors outside of the gas-discharge tube formed a nearly confocal cavity ($b' \approx d$), a condition for which the spot size w_s' is a minimum, and is small compared with the reflector aperture.

A technique has been found, however, for isolating a number of pure modes of a concave spherical cavity. It consists of adjusting the external cavity to a nearly spherical condition ($d/b' \approx 2$), for which the spot size w_s' is an appreciable fraction of the aperture, so that only a limited number of modes can be excited. Then, by adjusting the maser gain and manipulating the mirrors and windows to utilize irregularities in their optical properties, it is possible to raise the losses for all but one mode above the threshold for oscillation.

The accompanying photographs of a dozen pure modes (Fig. 2) illustrate the results. As the photo-current in the image convertor is a measure of beam intensity rather than field amplitude, the patterns correspond in form to the square of the wave functions of the harmonic oscillator.

The photographs were obtained by simulating a near-spherical cavity with a concave mirror of about 130-cm radius, at a distance slightly less than 130 cm from a flat mirror (at the plane of symmetry of a concave-spherical cavity). For these conditions, w_s' was about 2 to 3 mm, and the aperture radius about 7 mm.

Measurements of the spacing between zeros of the patterns agree closely with the calculated values derived from (1)–(4). Purity of the individual modes was further assured by observing that the beat-frequencies between modes could not be detected when the single-mode patterns were obtained.

The mode patterns do not exhibit azimuthal symmetry, due to astigmatism introduced by the Brewster-angle windows. This will be analyzed in greater detail in a later publication.

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Coupling of Laser Rods*

A cylindrical-optical maser oscillation is bounded primarily by the reflecting ends of its rod enclosure, which is typically the order of 10^5 wavelengths in length. A qualitative model of the oscillation is a mode band having a "width" selected from a nearly continuous spectrum of possible modes. In terms of the signal emerging from one slightly transparent aperture, a characterization of the oscillation is primarily the distribution of certain properties of the disturbance evaluated in the plane of the aperture. These are the distributions of intensity, phase, and phase uncertainty (or, variation in the mutual coherence function), which may be investigated by near and far field studies of the output.

New information may be obtained by comparing the radiation emitted from both ends of the rod in an interference experiment. However, a typical solid laser length is of the same magnitude as the coherence length, L, of its radiation, where $L \sim c/\Delta\nu$ is the interference path difference for which fringe contrast becomes arbitrarily small.[1] Therefore, for interference, the electromagnetic field along any line parallel to the cylinder axis must exhibit a degree of standing wave behavior. A crude picture of such a standing wave pattern may be obtained by considering the reflection of a plane wave λ_0 at an angle of incidence θ, between opposing parallel perfect reflectors a distance L apart. For certain $\theta = \theta_n(\lambda_0)$ stationary waves having the so-called "cut-off" wavelength $\lambda_0/\cos\theta$, result along any line normal to the reflectors. The curved wavefront and frequency width of the laser may be simulated by a suitable superposition of such θ_n modes.

* Received by the IRE, December 11, 1961.
[1] G. D. Boyd and J. P. Gordon, "Confocal multimode resonator for millimeter through optical wavelength masers," *Bell Sys. Tech. J.*, vol. 40, pp. 489–508; March, 1961.
[2] D. I. Schiff, "Quantum Mechanics," McGraw-Hill Book Co., Inc., New York, N. Y., p. 66; 1949.
[3] W. W. Rigod, H. Kogelnik, D. J. Brangaccio, and D. R. Herriott, "Gaseous optical maser with external concave mirrors," *J. Appl. Phys.*, to be published.

* Received by the IRE, December 8, 1961.
[1] A line width of ~0.1Å is assumed.

For small θ, spacing of the modes is given by $\Delta\theta_n = \sqrt{2\lambda_0/L}$. Thus for $L \gg \lambda_0$, as in the laser, a nearly continuous mode spectrum exists even for monochromatic radiation. If a time-integrating detector such as photographic film is used, the necessary phase correlation must exist throughout the measurement for observation of interference. Heating of a pulsed laser changes its length by several wavelengths, yet the resulting mode shift should produce no noticeable effect on standing wave symmetry. This conclusion results from the assumption that the rise time of the oscillation is negligible so that mode shifting would not appear discontinuous. Factors such as time-dependent temperature gradients, spatially nonuniform magnetic fields, etc., which may disturb the stimulated emission process, would be more likely to influence results.

The experimental geometry employed for studying this type of interference consists of a two-element optical antenna using a total internal reflection 90° prism for coupling, as shown in Fig. 1. Parallel alignment of the ruby rods is insured by adjusting a special mount for perpendicularity with respect to the optically flat prism face. This is determined by the observation of Newton's rings from the back of the prism. The optical mismatch at the ruby-air-glass boundary is reduced by replacing the air layer with a refractory liquid of intermediate index. The degree of coupling is controlled by varying the transparency of the prism end of the rods. The exposed ends have dielectric high-reflection coatings. The rod axes are spaced about 1½ diameters apart ($10^4\lambda$) and the unit fits into a conventional General Electric helical xenon flashlamp assembly.

Interference was photographically observed as shown in Fig. 2 by placing a narrow slit (0.005×0.125 inch) at the center of each rod aperture, and adjusting the degree of coupling. When the transmission at the prism ends was approximately ½, time integrated stationary fringes of correct spacing were detected by far field microphotographic measurements.[2] Except for a factor of 3 reduction in fringe spacing, the over-all pattern is similar to that reported previously for narrow slits separated by 3000 λ at the end of a single ruby laser rod.[3] We note again the time-integrating property of the detector implying correlation throughout the laser output pulse. It is expected that evaluation of internal fields under various conditions will result from the current study of such interference patterns.

In the two-element experiment described above, the coupling geometry has additional properties worthy of comment. At first glance the system appears, except for a short optical path in the coupling prism, equivalent to a single rod of double length. In so far as the oscillating component of the radiation is concerned this is apparently correct. However, this is one of a class of configurations in which the uncorrelated component of the internal radiation flux in each rod is reduced and decoupled from the system. In this case, the mechanism for the

[2] In the absence of coupling no interference pattern is observed between the laser beams. This is as expected since any interference should not persist for times greater than the coherence time $1/\Delta\nu \sim 10^{-10}$ sec.

[3] J. Masters and G. B. Parrent, Jr., "Propagation in laser crystals," PROC. IRE (Correspondence), vol. 50, pp. 230–231; February, 1962.

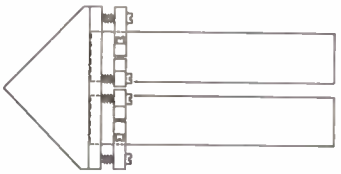


Fig. 1—Simple coupling geometry.



Fig. 2—Interference fringes from coupled laser rods, fringe spacing $\sim 10^{-4}$ radians.

effect is the total reflection property of the prism, which couples radiation whose direction is within a small axial elliptical conoid, rejecting the off-axis radiation by transmission out of the prism or by loss due to beam divergence within the prism path. This directional selectivity increases as the prism index approaches $\sqrt{2}$.

Multiple coupling of n separate rods of diameter d may be compared phenomenologically with an identical multiple aperture screen in front of a single large rod of diameter nd on the basis of uncorrelated flux, assuming both systems uniformly absorb the same pump energy per unit volume. If all bounding surfaces have only slight transparency for the fluorescence radiation, one would expect such uncorrelated radiation-energy density to be nearly independent of volume. However, whereas the transparency of bounding surfaces is actually appreciable, a large transparent surface to volume ratio is likely to be a geometric factor in uncorrelated radiation noise reduction.[4] In this regard, the greater transparent surface to volume ratio of the coupled system for a given aperture configuration (greater by the factor n) may be of significance.

From the results reported, it is clear that the coupling of independently pumped lasers is feasible, enhancing the possibility of shaped diffraction patterns and increased power densities, *i.e.*, optical antenna arrays.

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[4] N. G. Basov, O. N. Krokhin, and Y. M. Popov, "Generation, amplification, and detection of infra-red and optical radiation by quantum-mechanical systems," *Soviet Phys.*, vol. 3, pp. 718–719; March, 1961.

Cathode-Ray Frequency Meter and Signal Analyzer*

A cathode-ray deflection structure for measuring, practically instantaneously, the frequency of a signal has recently been constructed and tested. This cathode-ray frequency meter, proposed by R. C. Cumming, utilizes the transit-time effects of an electric deflection structure as the basis of frequency measurement. A straight-line display whose slope is a function of frequency and whose length is proportional to the square root of power is produced by a sinusoidal signal. More complicated signals create distinctive, useful patterns. The frequency range of the experimental tube is about 200 to 500 Mc. Its frequency resolution is 5 Mc or better. Design for another frequency ranges, centered either higher or lower, is equally feasible.

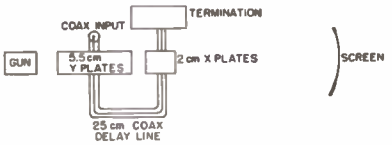


Fig. 1—Schematic drawing of the cathode-ray frequency meter.



Fig. 2—Multiple-exposure photograph showing the straight-line displays produced by sinusoidal signals of different frequencies applied one at a time.

It is well known that the magnitude of the electron deflection in a cathode-ray tube is a function of frequency when the electron transit time through the deflection plates is appreciable compared with the period of the signal. This dependence of deflection on frequency has been exploited for frequency measurement, as follows: A cathode-ray tube in which the y deflection plates are approximately 2.5 times as long as the x deflection plates was constructed, and the signal was applied in the manner shown schematically in Fig. 1. The 25-cm coaxial delay line shown in Fig. 1 delays the signal by an amount which is numerically equal to the time required for an electron to travel from the center of the y deflection plates to the center of the x deflection plates. This method of applying the signal to the plates causes the x and y deflections to be in phase, so that a straight-line display is obtained on the screen. Furthermore, the different lengths used for the x and y deflection plates make the x and y components of deflection vary differently with frequency, causing the slope of the straight-line display to be a function of frequency, which is the desired result.

Fig. 2 shows a multiple-exposure photograph of the displays produced by sinusoidal signals of different frequencies applied one at a time. The display for a given single frequency is a straight line. The slope of the straight line is a function of frequency, permitting calibration of frequency vs slope. The strength of the signal does not affect the frequency calibration; a stronger signal produces a longer line having the same slope. The longest line in Fig. 2 is a "zero-frequency" calibration line produced by a sinusoid whose frequency is so low that transit-time effects are negligible. The other lines, reading counter-clockwise from the LF line, are given by 200-, 250-, 300-, 350-, 400-, 450-, and 500-Mc sinusoids, respectively. An input power of the order of 1 w is required to produce a display about 1 in in length.

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A noteworthy feature of the cathode-ray frequency meter is its capability to indicate the frequencies and amplitudes of more than one signal applied simultaneously. The simultaneous application of N sinusoids, irrationally related in frequency, produces a polygonal display having N pairs of parallel straight sides. The slopes of the N pairs of parallel sides indicate the N-respective frequencies of the sinusoids. The slope-vs-frequency calibration is independent of the number and relative amplitudes of the sinusoids simultaneously present. Similarly, the lengths of the various sides are proportional to the amplitudes of the respective sinusoids.

Nonsinusoidal periodic signals create complex Lissajous displays; the shape of each particular display is characteristic of a certain signal waveform. These Lissajous displays may be used to identify signals or to check amplifiers for harmonic distortion, but not in general to measure frequency.

The electron transit time through the deflection structure is of the order of one RF cycle. An interesting consequence of this fact is that the frequency meter reads the carrier frequency rather than the Fourier spectrum of piecewise sinusoidal waves. Hence, the frequency resolution is not degraded when short rectangular RF pulses or phase-reversal signals are being observed. For example, 0.1-μsec RF pulses have a spectrum about 10-Mc wide. However, there is no fundamental reason why the cathode-ray frequency meter cannot resolve the carrier frequency of such pulses to an accuracy much better than 10 Mc.

Methods of improving the sensitivity of the tube are under consideration. One possibility is to utilize some form of traveling-wave deflection. Another is to provide a radial dc electric field in the region between the deflection plates and the screen. This field would enhance the initial radial deflection produced by the deflection plates.

This relatively simple electronic meter offers to be a useful alternative or adjunct to conventional spectrum analyzers. Its display provides a rather unique pictorial measure, or "fingerprint" of a signal.

The technical contributions of D. D. Luby and R. A. Moran are gratefully acknowledged.

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Causality*

The Editorial Board of the IRE has performed a public service in publishing the reflections of Golay.[1] Nothing could be of more "long-range importance and general interest"[2] to IRE members.

In discussing the creation of life, Golay states that "we evolved as a matter of course almost. . . ." The use of the word *almost* is scientifically inadmissible. It is a violation of the principle of causality, the principle that effect uniquely follows cause. In accordance with causality, a supercomputer that was constructed one billion years ago, and that was given the state of every particle and field in the universe at a certain time $t=0$, would have predicted that Golay's thought-provoking paper would be published at $t=$one billion years. There is no evidence whatsoever that living matter disobeys causality. Biophysical and biochemical studies show that a living cell is a complicated electrochemical engine;[3] an atom of carbon in a brain cell is no different than any other atom of carbon.

The principle of causality leads to the conclusion that we evolved as a matter of course inevitably, that *free will* and human creativity are illusions, that life must exist outside the solar system.

We have not been able to detect nonrandom signals from outer space. The likelihood of receiving these signals is extremely small because of the huge amounts of transmitter power that are required,[4] but it is quite possible, instead, that the signals are never launched because the evolutionary end of all "intelligent" organisms is self-destruction.

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* Received by the IRE, October 17, 1961.

[1] M. J. E. Golay, "Reflections of a communication engineer," Proc. IRE, vol. 49, pp. 1378–1382; September, 1961.

[2] "Information for IRE authors," Proc. IRE, vol. 48, pp. 1536–1539; September, 1960.

[3] Issue on "The Living Cell," *Sci. Am.*, vol. 205; September, 1961.

[4] S. S. Huang, "Life outside the solar system," *Sci. Am.*, vol. 202, pp. 55–63; April, 1960.

On Self-Organizing Systems*

The criticisms which appeared in these Proceedings by Harmon[1] and Coleman[2] on Hawkins' article[3] and, to a lesser extent, on Brain's article[4] illuminate a problem which is rapidly growing in importance.

The problem is this: Many of us in electronics today are engaged in the investigation of advanced communication concepts for dealing with the interpretation of multidimensional signals. We cannot avoid the fact that the inspiration for our concepts derives mainly from theories and suggestions as to how biological nerve-networks perform this task. The nub of the problem is that we find ourselves working deeper and deeper into an area of immediate concern to people in the biological sciences and, in most cases, we do not know their field well enough to discuss our findings in their terms. Nevertheless, the desire to use biological terms is strong and sometimes valid. The terms serve to dramatize the concepts and point out the potential which is, of course, always inherent in the mathematics but not always obvious.

I took the precaution of having a recent book review[5] which I was preparing sent to Marcel Verzeano of the UCLA Medical Center for comments. He took strong exception to the use of anthropomorphic terms and, in fact, was kind enough to aid in evaluating the biological aspects of the text under review. I recall Mark Kac, then of the Mathematics Department of Cornell University, now of the Rockefeller Institute, in reviewing some mathematical work of his associates, insisting that biological analogs be left out of the discussion and the concepts be developed in purely mathematical terms. Finally, R. Bolt of the National Science Foundation, in recent conversations on the need for closer contact between the communication specialists and the biologically oriented scientists, stressed the need for maintaining the relationship on the highest professional levels with no "scientific talking down" involved on either side.

People in the biological sciences have indicated their dissatisfaction before these recent critical notes. At the 1960 WESCON convention[6] a medical doctor apparently found himself assigned to comment upon certain aspects of medical electronics. He devoted a good part of his article to objecting to the topic which he was assigned and to correcting what he felt were misconceptions among engineers on the subject of medical instrumentation. Although not concerned with advanced communication theory, this last example reveals the common theme. All these examples indicate that there is a growing awareness of the problem of developing a mutual understanding between the two disciplines of electronics and biology.

Brain,[7] in his response to the criticisms, makes a rather picturesque case for vigor as well as rigor in science. I doubt that a journal of the stature of the Proceedings would like to take that argument as its main line of defense. It is obvious that the IRE is involved here as well as the authors, for these articles were presumably submitted to the usual review procedure. We don't want to lose the right to use anthropomorphic terms although they should be used with caution. Loose and general references to biological work, however, reduces the scientists in that area to a vague and general anonymity. The fact that our people wish to retain their identity with their own work makes it imperative that we be particularly careful of the rights of others at a time when we are obviously drifting into a close working relationship with them.

It is evident, however, that the lack, in IRE papers, of specific references to work in

* Received by the IRE, October 16, 1961.

[1] L. D. Harmon, "Neural analogs," Proc. IRE, vol. 49, pp. 1316–1317; August, 1961.

[2] P. D. Coleman, "On self-organizing systems," Proc. IRE, vol. 49, pp. 1317–1318; August, 1961.

[3] J. K. Hawkins, "Self-organizing systems," Proc. IRE, vol. 49, pp. 31–48; January, 1961.

[4] A. E. Brain, "The simulation of neural elements by electrical networks based on multi-aperture magnetic cores," Proc. IRE, vol. 49, pp. 49–52; January, 1961.

[5] P. M. Kelly, "Self-organizing systems, M. C. Yovits and S. Cameron, Eds.," Proc. IRE (Book Review), vol. 49, pp. 533; February, 1961.

[6] J. B. Dillon, "An anesthetized man in a normal environment," 1960 IRE WESCON Convention Record, pt. 5, pp. 73–74.

[7] A. E. Brain, "Author's comment to neural analogs," Proc. IRE (Correspondence), vol. 49, p. 1317; August, 1961.

the biological field springs, not from any desire to withhold credit but, simply, because of the lack of specific knowledge. Further, it is unrealistic to expect the communication scientist to keep up with the rapid developments in biology when he faces such a tremendous task in keeping up with just the literature in his own field. It might be helpful if the IRE were to solicit an article with a title somewhat as follows, "A Biologist Looks at Electronic Analogs of Neural Networks." Such an article, if it brought us up-to-date on the significant developments in the biological field, would be extremely helpful.

I realize that I have not commented in detail on either the original articles, the criticisms, or the rebuttals. It seems to be less important to dissect a particular situation than it is to recognize a growing problem of a general nature. It is in order to call for caution in the use of the biological analogs in presenting new concepts. It would probably be worthwhile, on the part of the IRE, if biologically-oriented scientists were involved in the review of papers in which such analogs were used to any extent.

It is important that we take positive steps to correct the situation described above for, without a doubt, these incidents will arise again, and it is necessary to show that we are well-intentioned and that the shortcomings revealed are not a part of our normal behavior pattern.

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Comment on Coherence*

Some of the confusion still surrounding the meaning of coherence appears to be due to the different viewpoints of the optician and of the communication engineer, whom the advent of the laser has obliged to communicate more with each other.

Taking the optical viewpoint, I have defined coherence as the essentially qualitative characteristic of a CW signal when compared to an optical emission line or to a narrowly filtered band of white noise,[1] for an optician will say he is dealing with a coherent light source when he observes that his light source forms a sharp interference pattern with indefinitely delayed portions of itself, even if this pattern is not stationary. This is because the optician has, in the past, termed coherent two interfering optical beams derived from the same light source and finitely delayed with respect to each other. On the other hand the communication engineer, long familiar with CW signals, takes for granted the existence of sharp interferences between two portions of the the same signal which are indefinitely delayed with respect to each other, and may prefer to reserve the term coherence to designate a measure of some desirable property of the signal source, such as the degree of stationarity of the interference pattern, or of the phasor formed by said two portions.

At this point the communication engineer runs into some interesting problems. For instance, if his source is an ideal CW generator installed on a planetoid having its own orbit around the sun, the communication engineer will attempt to track the weak signal he receives with a phase-locked loop, the behavior of which may be described by the differential equation of a damped system such as

$$\left(1 + T\frac{d}{dt}\right)^n \phi_s = \phi_r,$$

in which ϕ_s and ϕ_r are the phase of the received signal and of the slaved signal, and in which n and T have been selected so as to optimize the SNR.

Since a large T means a very narrow bandwidth of reception, T may be taken as a measure of the coherence of the received signal, even though there is a rapid rotation of the interference phasor formed by two portions of the signal delayed with respect to each other. Furthermore, the order n of of the characteristic differential equation should be taken into account, which might lead to the view that coherence is a multidimensional quantitative concept.

The concept of coherence as I have defined it is not without its own questions. For instance, should we term partially coherent signals characterized by a phasor-end distribution having the aspect of a molehill depressed at the top, and is coherence, even as an optician would define it, a quantitative concept after all?

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* Received by the IRE, October 25, 1961.

[1] M. J. E. Golay, "Note on coherence vs. narrow-bandedness in regenerative oscillators, masers, lasers, etc.," PROC. IRE (Correspondence), vol. 49, pp. 958–959; May, 1961.

Roughness and Radar Return*

In view of the recent discussion in the IRE correspondence columns (Hayre [1], Senior and Siegel [2], [3]) the author makes an attempt to quantize the word "rough" in terms of statistical properties of a rough surface. Surfaces are often referred to as quasi-smooth, smooth, almost smooth, rough and very rough, etc., but rarely is the roughness quantitatively specified. The order of roughness is closely related to the ratio of specular and scatter return of electromagnetic waves from a surface, where specular return is defined as the return from a mirror-like, perfectly smooth surface, and the scatter as that from a nonsmooth surface. In practice, the return from a rough surface may be said to be made up of both specular and scatter components. It is from this that some inference (Moore [4], Senior [2]) can be had about the nature of roughness. A motion of the radar with respect to a rough target area causes fading of the return signal. The fading spectrum has sometimes been used to detect the order of roughness, as fading statistics of signal return from a particular terrain are determined by the statistical properties of the surface, the relative motion, the carrier frequency and other radar parameters. The radar scattering cross section (σ_0) is also employed in the prediction of surface roughness of a target terrain.

Various models have so far been used to predict the roughness of the target terrain from the radar return data. The Davies-Moore-Hayre model (Hayre and Moore [5]) uses the experimental autocoveriance of the elevations and their probability density function, in applying the Kirchhoff-Huygens' principle. Certain standard assumptions about perfect conductivity, isotropic scatterers, near-vertical incidence and no part of the terrain being shadowed by any other part, etc. were employed. Contour maps were used to calculate the autocovariance functions. This resulted in the following expression for the radar scattering cross section:

$$\sigma_0 = (4\sqrt{2}\pi B^2/\lambda^2)(\theta \cos^2\theta/\sin\theta) \cdot \exp(-4k^2\sigma^2\cos^2\theta) \cdot \sum_{n=1}^{\infty} (4k^2\sigma^2\cos^2\theta)^n \cdot [(n-1)!(2B^2k^2\sin^2\theta + n^2)^{3/2}]^{-1} \quad (1)$$

where

σ = standard deviation of elevation of points on the surface above and below the mean surface
λ = wavelength
$k = 2\pi/\lambda$, wave number
v = angle of incidence measured from the outward surface normal
B = roughness characteristic constant (in auto-covariance function).

This expression has been shown to check well with experimental data on terrain and lunar return (Hayre and Moore [5], Hayre [1]). In the following paragraph, (1) will be used to infer something about the transitional value of σ/λ in going from a quasi-smooth to a rough surface.

Roughness may be specified quantitatively in terms of standard deviation (σ) and autocovariance constant (B), maximum and minimum size of the surface perturbations, or a roughness coefficient contained in the probability density functions for the areas at a particular elevation (Brown [6]).

A search of the literature on the subject shows that an order of roughness separating an "almost smooth" and a rough terrain is not clearly specified. For instance, Evans [7] and Kerr, *et al.* [8] assume that a surface is rough if the size of its perturbations is greater than one eighth of a wavelength. The distribution of such perturbations above and below this size is not discussed. Moore [4] uses an interesting technique to separate the specular and scatter radar return components. The reduction of the specular component with increasing roughness may be used to indicate a transitional order of roughness between an almost smooth

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surface and a rough surface. Moore lists a separation factor a which shows reduction in specular component due to energy going into the scatter component, equal to the exponent $(-2k^2\sigma^2)$. When the exponent $2k^2\sigma^2$ becomes unity, the factor a reduces to 0.368. The value of σ/λ corresponding to this reduction in specular component is 0.113, and it is comparable with the assumed value given by Evans and Kerr, *et al.*

Radar scattering cross section, vs angle of incidence curve, is very steep near zero angle of incidence for a nearly smooth surface. As the surface roughness is increased, this curve becomes flatter, but never reaches a horizontal line. For given dielectric properties, some of these curves of σ_0 vs θ for varying degrees of roughness, cross each other in the range of angles between zero and about ten degrees (Dye [9], Hayre and Moore [5].) The constants σ and B are indicative of the order of roughness of a surface: a small σ and a large B represent a surface that appears smooth to the eye and "nearly smooth" to the radar, while a comparatively large σ and a small B would represent a "relatively rough" surface. Small values of B and σ indicate a very rough terrain with small scale (vertical) roughness; whereas a large B and a large σ represent to the eye a relatively smooth surface with large scale perturbations, but to the radar, a rough surface. A study of σ_0 vs θ curves obtained using Davies-Moore-Hayre model seems to support the conclusion that $\sigma/\lambda \approx 0.08-0.125$ is a transition range from quasi-smooth to a rough surface.

The above discussion suggests that the specification of standard deviation and the autocovariance constant, B, in terms of the wavelength used, is desirable for describing a target terrain roughness, rather than the use of qualitative adjectives.

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Bibliography

[1] H. S. Hayre, "Radar scattering cross section applied to moon return," Proc. IRE (Correspondence), vol. 49, pp. 1433; September, 1961.
[2] T. B. S. Senior and K. M. Siegel, "A theory of radar scattering by the moon," *J. Res. NBS*, vol. 64D, pp. 317–229; May–June, 1960.
[3] T. B. A. Senior and K. M. Siegel, "A radar cross section of the moon," Proc. IRE (Correspondence), vol. 49, p. 1938; December, 1961.
[4] R. K. Moore, "Resolution of Vertical Incidence Radar Return into Random and Specular Components," Univ. of New Mexico Engr. Expt. Sta., Albuquerque, Tech. Rept. No. EE-6; July, 1957.
[5] H. S. Hayre and R. K. Moore, "Theoretical Scattering Coefficients for Near-Vertical Incidence from Contour Maps," *J. Res. NBS*, vol. 65D, pp. 427–432; September–October, 1961.
[6] W. E. Brown, Jr., "A lunar and planetary echo theory," *J. Geophys. Res.*, vol. 65, pp. 3087–3095; October, 1960.
[7] J. V. Evans, "Raido Echo Studies of the Moon," Lincoln Lab., Massachusetts Institute of Technology, Lexington, Rept. No. 3G-0001; August 30, 1960.
[8] D. E. Kerr, W. T. Fishback and H. Goldstein, "Propagation of Short Radio Waves," McGraw-Hill Book Co., Inc., New York, N. Y., p. 411; 1951.
[9] J. E. Dye, "Ground and sea return signal characteristics of microwave pulse altimeters," *Trans. Symp. of Radar Return*, pt. 1 (Unclassified), Univ. of New Mexico, Albuquerque, pp. 111–114; May, 1959.
[10] H. Davies, "The reflection of electromagnetic waves from a rough surface," *J. IEE*, pt. 4, vol. 101, pp. 209–215; August, 1954.

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.[1] The corrections reported here were arrived at by means of improved measurement methods based on NBS LF and VLF transmissions.

The WWV and WWVH time signals are also kept in agreement with each other. Also they are locked to the 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.

Subsequent changes were as follows:

Frequency Offset
January 1, 1960, −150 parts in 10^{10}
January 1, 1962, −130 parts in 10^{10}

Time Adjustments
December 16, 1959, retardation, 20 milliseconds
January 1, 1961, retardation, 5 milliseconds
August 1, 1961, advancement, 50 milliseconds

Adjustments were made at 0000 UT on the foregoing dates.

WWV Frequency
With Respect to U. S. Frequency Standard

1961 November	Parts in 10^{10}†
1	−150.2
2	−150.0
3	−150.0
4	−149.8
5	−149.7
6	−149.7
7	−149.4
8	−149.3
9	−149.3
10	−149.0
11	−148.9
12	−149.0
13	−148.9
14	−149.0
15	−148.9
16‡	−148.8
17	−149.1
18	−149.0
19	−148.9
20	−148.9
21	−149.0
22	−149.1
23	−149.2
24	−149.3
25	−149.3
26	−149.3
27	−149.1
28‖	−149.0
29	−149.5
30§	−149.7

† A minus sign indicates that the broadcast frequency was low. The uncertainty associated with these values is $\pm 5\times10^{-11}$.
‡ Frequency adjusted -0.3×10^{-10} on November 16.
‖ Frequency adjusted -0.8×10^{-10} on November 28.
§ Frequency adjusted -0.4×10^{-10} on November 30.

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* Received by the IRE, January 2, 1962.
[1] Refer to "National Standards of Time and Frequency in the United States," Proc. IRE, vol. 48, pp. 105–106; January, 1960.

Synthesis of Nonuniform Transmission Lines and Riccati's Differential Equation*

Recent work by Sugai[1] and Stickler[2] suggests a revival of interest in the exact solution of nonuniform transmission line problems.

It will be shown that if one specifies that the input reflection coefficient should vary in some predetermined manner along a line terminated in an arbitrary impedance, it is in general possible to synthesize a nonuniform line meeting this requirement.

Consider a transformation

$$y = \frac{Cf + Bg}{h(Cf' + Ag')}, \tag{1}$$

where y and x are the dependent and independent variables, A, B and C are arbitrary constants, f, g and h are arbitrary functions of x, while primes denote differentiation with respect to x.

Eliminating the constant C we find that

$$y' = \frac{1}{h}\frac{B(gf' - g'f)}{Bgf' - Ag'f} + y\left[-\frac{h'}{h} + \frac{(B-A)g'f'}{Bgf' - Ag'f} - \frac{Bgf'' - Ag''f}{Bgf' - Ag'f}\right] - y^2hA\frac{g''f' - g'f''}{Bgf' - Ag'f} \tag{2}$$

subject to the following conditions:

1) f and g as well as their first and second derivatives exist throughout the range of x under consideration;
2) $Bgf' - Ag'f \neq 0$;
3) $g''f' - g'f'' \neq 0$.

The third condition, while not essential for (2) to hold, ensures the nondisappearance of the square term in y.

The input reflection coefficient of a nonuniform transmission line satisfies a differential equation

$$\frac{d\Gamma}{dx} = -\frac{1}{2}\frac{d\ln K}{dx} + 2\gamma\Gamma + \frac{1}{2}\frac{d\ln K}{dx}\Gamma^2, \tag{3}$$

where Γ, K and γ are the reflection coefficient, nominal characteristic impedance and the propagation constant, respectively, while x is the distance along the line which may be normalized with respect to the overall length, *i.e.*, $0 \leq x \leq 1$.

Comparing (2) and (3), we deduce that

$$\frac{1}{h}\frac{B(gf' - g'f)}{Bgf' - Ag'f} \equiv -\frac{1}{2}\frac{d\ln K}{dx} \tag{4}$$

$$\frac{-h'}{h} + \frac{(B-A)g'f'}{Bgf' - Ag'f} - \frac{Bgf'' - Ag''f}{Bgf' - Ag'f} \equiv 2\gamma \tag{5}$$

$$-hA\frac{g''f' - g'f''}{Bgf' - Ag'f} \equiv \frac{1}{2}\frac{d\ln K}{dx}. \tag{6}$$

* Received by the IRE, October 23, 1961; revised manuscript received, November 14, 1961.
[1] I. Sugai, "A new exact method of nonuniform transmission lines," Proc. IRE, vol. 49, pp. 627–628; March, 1961.
[2] D. C. Stickler, "A note on Sugai's class of solutions to Riccati's equation," Proc. IRE (Correspondence), vol. 49, p. 1320; August, 1961.

From (4) and (6)

$$\frac{1}{2}\frac{d\ln K}{dx} = \pm\frac{\sqrt{AB(gf'-g'f)(g''f'-g'f'')}}{Bgf'-Ag'f} \quad (7)$$

$$k = \mp\sqrt{\frac{B(gf'-g'f)}{A(g''f'-g'f'')}} \quad (8)$$

while γ follows from (5) and (8).

The arbitrary constant C in (1) is determined by the reflection coefficient of the termination.

It can be noted that a transformation

$$y = \frac{Cf'+Ag'}{h(Cf+Bg)} \quad (9)$$

could have been chosen, yielding similar results.

Let

$$f = \exp(kx) \quad g = \exp(-mx) \quad (10)$$

$A = B = 1$, k, m—arbitrary constants.

Hence

$$k = \frac{-1}{\sqrt{km}} \quad (11)$$

$$\Gamma = \sqrt{km}\,\frac{C\exp(kx)+\exp(-mx)}{m\exp(-mx)-Ck\exp(kx)} \quad (12)$$

$$K = \sqrt{\frac{Z_0}{Y_0}}\exp(2\sqrt{km}x)$$

$$\gamma = \tfrac{1}{2}(m-k). \quad (13)$$

Let

$$2\sqrt{km} = \alpha \text{ and } \tfrac{1}{2}(m-k) = \sqrt{Z_0Y_0}. \quad (14)$$

Hence

$$k = -\sqrt{Z_0Y_0} \pm \sqrt{Z_0Y_0+\tfrac{1}{4}\alpha^2} \quad (15)$$

$$-m = -\sqrt{Z_0Y_0} \mp \sqrt{Z_0Y_0+\tfrac{1}{4}\alpha^2} \quad (16)$$

$$Z = Z_0\exp(\alpha x) \quad Y = Y_0\exp(-\alpha x), \quad (17)$$

Z_0 and Y_0 represent the series and shunt impedance per unit length of line, respectively, at the reference point (here $x=0$).

The nonuniform line will be recognized as being exponential.

A relatively simple pair of functions f and g was chosen with the object of illustrating the procedure and deducing known results. The example was a particular case of a substitution $f=g^{-n}$ which leads to ($A=B=1$, n—arbitrary constant)

$$K = \sqrt{\frac{Z_0}{Y_0}}f^{2\sqrt{n}} \quad (18)$$

$$\gamma = \tfrac{1}{2}(n-1)\frac{f'}{f}. \quad (19)$$

In a more general case one can, for example, assume that $f=\sum a_nx^n$ and $g=\sum b_nx^n$.

The input reflection coefficient must be dimensionless, its absolute magnitude cannot exceed unity at all frequencies and K must be real for all x; an investigation of requirements imposed on f and g is straightforward but outside the scope of this communication.

The input impedance of a nonuniform line satisfies another Riccati's equation while voltage and current lead to second-order linear homogeneous differential equations which can always be reduced to a Riccati's equation; hence these variables are amenable to the foregoing analysis as well.

In conclusion, the author wishes to express his gratitude to his thesis supervisor Prof. D. L. Waidelich, Visiting Professor of Electrical Engineering at the University of New South Wales, for valuable discussions and criticism.

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A Note on the Realization of Boolean Functions Using a Single Threshold Element*

A threshold element is a device with a finite number of inputs, to each of which there is applied a binary valued variable x_i. If x_i is applied to m_i of the inputs, it is said to have a multiplicity m_i. The single output F_n is given as

$$F_n = 1 \quad \text{if} \left[\sum_{i=1}^{n} m_ix_i\right] - t \geq 0 \quad (1)$$

$$F_n = 0 \quad \text{otherwise.}$$

F_n represents a Boolean function of the n input variables. An objective of interest which has been the goal of research in this area[1,2,3,4] is the establishment of a set of necessary and sufficient conditions for $F_n \in R$, where R is the class of Boolean functions which are capable of being realized as the output of a single threshold element. It is shown below that a theorem contained in the mathematics of convex sets supplies these conditions.

F_n may be expressed as the disjunctive logical sum of those input n tuples $v_i(x_1, x_2, \cdots, x_n)$ which satisfy inequality (1). Since there are 2^n such n tuples, F_n is given by

$$F_n = \sum_{i=0}^{2^n-1}\beta_iv_i \qquad \beta_i = (0, 1). \quad (2)$$

Each n tuple may be identified with the vertex of the unit n cube, U_n, by substituting a 1 in the position of an unnegated variable of the n tuple and a 0 in the position of a negated variable F_n, may thus, as is well known, be represented on a subset S_n of the 2^n vertices of U_n. S_n comprises those vertices in (2) for which $\beta_i=1$. The remaining vertices belong to the subset $\bar{S}_n$ upon which the complementary function $\bar{F}_n$ is represented. The 2^{2^n} possible Boolean functions of n variables are thus placed in a 1:1 correspondence with the 2^{2^n} possible partitions of U_n into the two disjoint subsets S_n and $\bar{S}_n$.

Inequality (1) may be geometrically interpreted as a closed half space and the n tuples satisfying the inequality are the subset of vertices comprising S_n.

The equation

$$\left[\sum_{i=1}^{n} m_ix_i\right] - t = 0$$

is an $(n-1)$ hyperplane, partitioning U_n into S_n and $\bar{S}_n$. Conversely, all subsets S_n (and also $\bar{S}_n$) formed by such a partition represent realizable Boolean functions. The necessary and sufficient conditions for the existence of this separating hyperplane are obtained from convex set theory.

A convex linear combination of n tuples in Euclidean n space, E^n, is an n tuple $v \in E^n$ given by

$$v = \sum_{i=1}^{k} a_iv_i \quad (3)$$

where

$$\sum_{i=1}^{k} a_i = 1.$$

A set S of n tuples in E^n is convex if for every pair $(v_1, v_2) \in S$ all linear convex combinations v of $(v_1, v_2) \in S$; that is

$$v = [av_1 + (1-a)v_2] \in S \quad 0 \leq a \leq 1. \quad (4)$$

Geometrically, (4) states that all points on the finite line segment joining $v_1, v_2 \in S$.

The convex hull, $C(S)$, of an arbitrary set of n tuples, is defined as the smallest convex set which contains S. An extreme n tuple of $C(S)$ is an n tuple which does not lie on a line segment joining two separate n tuples of $C(S)$.

There is a well-known theorem[5] which states that if two convex hulls in E^n are disjoint, that they may be strictly separated by a hyperplane. A separating hyperplane in U_n will exist if then

$$C(S_n) \cap C(\bar{S}_n) = \phi \quad (\phi = \text{null set}). \quad (5)$$

For (5) to hold and therefore for S_n and $\bar{S}_n$ to represent realizable Boolean functions F_n and $\bar{F}_n$, there must not exist any linear convex combination of n tuples in $C(S_n)$ equal to a linear convex combination of n tuples in $C(\bar{S}_n)$, or

$$\sum_{i=1}^{k} a_iv_i' \neq \sum_{i=1}^{r} b_iv_i'' \quad (v' \in S_n;\ v'' \in \bar{S}_n) \quad (6)$$

where

$$\sum_{i=1}^{k} a_i = \sum_{i=1}^{r} b_i \qquad (a_i, b_i \geq 0).$$

While inequality (6) establishes the necessary and sufficient conditions for $F_n \in R$, its application is too difficult to be practical. Inequality (6) is a very strong condition which applies to all pairs of convex sets. The nature of the convex hulls formed by S_n and $\bar{S}_n$ is, however, very special being 1) those whose extreme n tuples are the

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[1] M. C. Paull and E. J. McCluskey, Jr., "Boolean functions realizable with single threshold devices," Proc. IRE (Correspondence), vol. 48, pp. 1335–1337; July, 1960.

[2] Robert O. Winder, "Single stage threshold logic," *Proc. Natl. Electronics Conf.*, Chicago, Ill., October 10–12, 1960; vol. 16, p. 333, 1960.

[3] Calvin C. Elgot, "Truth functions realizable by single threshold organs," *Proc. Natl. Electronics Conf.*, Chicago, Ill., October 10–12, 1960, vol. 16, pp. 331–332, 1960.

[4] S. Muroga, *et al.*, "Theory of Majority Decision Elements," IBM Res. Ctr., RC 296; 1960.

[5] H. G. Eggleston, "Convexity," Cambridge University Press, Cambridge, England; 1958.

vertices of a unit n cube and, 2) those wherein the totality of extreme n tuples are the 2^n vertices of the unit n cube. It is possible that this special structure may give rise to a set of weaker conditions which would indicate nonintersection of the convex hulls and thus establish useable realizability criteria for $F_n \in R$.

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Consciousness and Research in Machine Intelligence*

The conventional notion of placing the "conscious act" in a place of pre-eminence in the search for the underlying principle behind the "ultimate" machine is altered by the psychoanalytic proposition that every mental act begins as an unconscious one. Experimental evidence, in particular the tachistoscopic picture exposure research (Mt. Sinai Hospital, N. Y.), confirms the notion that cognition involves an initial phase below "awareness" and a *potential* final phase in which a part of the system's reaction is what we think of as conscious perception. The failure to recognize the fully cognitive systems beneath the conscious level is secondary to the more inclusive deficiency of ignoring psychoanalytic theory and research which might elucidate the nature of sensory capacity, imaginative or "intuitive" thought, and symbolic processes of a metamathematical kind. The role of "neurotic" disturbances in learning is unclear and the suggestion from psychoanalytic considerations that learning should be an effortless process has an important effect on learning models and neuron analogs which incorporate the stimulus repetition phenomenon as arising from intrinsic properties of neural circuitry.

Further, the superfluousness of consciousness to the processes of intelligence casts serious suspicion on the current notions of Man's intellectual capacities—notions widely held among researchers in artificial intelligence.

The incessant activity of the brain may be regarded in terms of the interaction of nonisolated subsystems which are not in equilibrium but maintain instead a stable stationary nonequilibrium condition or move between such states due to altered external constraints (stimuli). The stability of such "open" systems undergoing irreversible processes is determined by minimum entropy *production* and such states may not be left by "spontaneous" irreversible processes, but only by a change of constraints. Then there is no real significance to the usual notions of "spontaneous behavior" or "act of will" so that we should take Freud's feeling that *all* behavior is triggered subconsciously as a sound approximate truth.

Minimum entropy production corresponding to constraints argues for a variational principle in mental activity rather than, *e.g.*, Selfridge's pandemonium paradigm or other trial-and-error or statistical decision process.

The whole of mental activity seems to be "more than the sum of its parts" and it is most tempting here to fall back to vitalism or the allusion to consciousness as the "force" which makes the Brain (or machine) a Mind. But the stationary-state coupling of irreversible processes in which use is made of the entropy increase of one process to promote the entropy decrease of some other process gives the illusion of "more than the sum of its parts" behavior. In this way seemingly incommensurate activities "cooperate" to maintain a system's autonomy. Such coupling is not by way of "phenomenological" or "transport" coefficients and appears on this account to thwart intuitive insights into the cause-effect relations.

The entropy coupling of two processes not connected phenomenologically may explain the data of the already mentioned flash-exposure experiments (tachistoscope) in which there is subconscious perception but no conscious experience of the event. By hypothesis, consciousness itself is "inert" in the chain of behavioral events, but these events produce changes somewhere in, *e.g.*, the reticular system of the brain (believed by many to be the seat of conscious sensation). These changes in excitation level might derive from the stationary state coupling of the processes of information transfer (a "transport" phenomenon) and the rate of acquisition of information in the transformation of a stimulus to its internal representation. Thus the "inert" system (consciousness) is coupled to a diffusion and rate process so that, barring "neurotic" disturbances, there are normal modes of "subception" which may or may not reach consciousness, depending on information diffusion and its transformation rate.

These results may be obtained from system equations which are well tried and indeed elementary and need not distract us from the more important comment on the logical difficulty in discussing the Mind by analogy with irreversible thermodynamics. On the other hand the same principles obtain in the "automated" cellular factory and while there is no desire to equate chemical regulation with "higher activities" there seems to be an important similarity between chemical subsystem communication and coupling problems, and these same questions which would have to be met in any self-sustaining information system more familiar to the research engineer.

To conclude, it is important to remark that from the viewpoint of psychoanalytic theory the current preoccupation with external superficial behavioral manifestations has fixed attention on processes which *follow* the generic events we seek—some of these externals being typical only of faulty system integration (neurotic processes) and some the evidence that the conscious level has no privileged status. Additionally, it is basic to psychoanalytics to regard *all* symbolic processes as being conjugated inextricably with what we have come to call "emotions" and "feelings."

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Bandwidth of Hybrid-Coupled Tunnel-Diode Amplifier*

The hybrid-coupled tunnel-diode amplifier previously described in the literature[1,2] usually employs a hybrid consisting of two coupled strip transmission lines.[3,4] Letting $[S]$ be the scattering matrix of the amplifier consisting of the hybrid and two tunnel diodes (see Fig. 1):

$$[S] = \begin{bmatrix} 0 & S_{12} & 0 & S_{14} \\ S_{12} & 0 & S_{14} & 0 \\ 0 & S_{14} & 0 & S_{12} \\ S_{14} & 0 & S_{12} & 0 \end{bmatrix}.$$

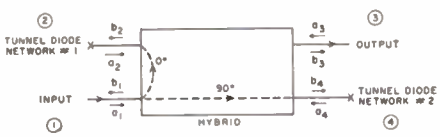


Fig. 1.

This matrix entails the following equations:

$$\begin{cases} b_1 = S_{12}a_2 + S_{14}a_4 \\ b_2 = S_{12}a_1 + S_{14}a_3 \\ b_3 = S_{14}a_2 + S_{12}a_4 \\ b_4 = S_{14}a_1 + S_{12}a_3 \end{cases}$$

where b_i is the reflected wave and a_i is the incident wave at port i. Now $\Gamma = a_4/b_4 = a_2/b_2$ for identical tunnel-diode networks at ports 2 and 4. (Γ is voltage reflection coefficient.) Now $a_3 = 0$ if the amplifier output at port 3 is terminated in a matched load at all frequencies. This reduces the equations to the following:

$$\begin{cases} b_1 = (S_{12}^2 + S_{14}^2)\Gamma a_1 \\ b_3 = 2S_{12}S_{14}a_1 \end{cases}$$

* Received by the IRE, November 24, 1961.

[1] J. Sie, "Absolutely stable hybrid coupled tunnel-diode amplifier," PROC. IRE, vol. 48, p. 1321; July, 1960.
[2] J. C. Greene, "Noise considerations for hybrid coupled negative-resistance amplifiers," PROC. IRE, vol. 49, pp. 626–627; March, 1961.
[3] J. K. Shimizu, "Stripline 3 db directional coupler," 1957 IRE WESCON CONVENTION RECORD, pt. 1, pp. 4–15.
[4] J. K. Shimizu and E. M. T. Jones, "Coupled transmission line directional couplers," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-6, pp. 403–410; October, 1958.

Letting G = the power gain of the amplifier,

$$G = \left|\left(\frac{b_2}{a_1}\right)\right|^2 = |2S_{12}S_{14}\Gamma|.$$

Putting G into db,

$$G = 10\log_{10}|2S_{12}S_{14}\Gamma| = 20\log_{10}|2S_{12}S_{14}| + 10\log_{10}|\Gamma|^2.$$

It is apparent that the power gain is determined by two factors: the loss factor $20\log|2S_{12}S_{14}|$ which accounts for the frequency sensitivity of the hybrid, and the gain factor $10\log|\Gamma|^2$ which accounts for the frequency sensitivity of the tunnel-diode network.

Previous published work has accounted for $10\log|\Gamma|^2$ but has neglected $20\log|2S_{12}S_{14}|$. If the tunnel-diode network is independent of frequency (in practice that is never true), the frequency sensitivity of the hybrid imposes an upper limit on the bandwidth of the amplifier. The purpose of this paper is to pin down this upper limit.

It can be shown that for the hybrid previously mentioned

$$S_{12} = \frac{V_2}{V_1} = \frac{jK\sin\theta}{\sqrt{1-K^2}\cos\theta + j\sin\theta}$$

(in phase component),

$$S_{14} = \frac{V_4}{V_1} = \frac{\sqrt{1-K^2}}{\sqrt{1-K^2}\cos\theta + j\sin\theta}$$

(quadrature component),

where K is a measure of hybrid coupling, θ = electrical length of coupled strip transmission lines. It can be shown that

$$20\log|2S_{12}S_{14}| = 20\log\left[\frac{2K\sqrt{1-K^2}\sin\theta}{(1-K^2\cos^2\theta)}\right].$$

This must be equated to $\frac{1}{2}$ to determine the value of θ at the half-power points.

Letting $K=0.707=\sqrt{2}/2$ for critical coupling

$$\sin^2\theta - 2\sqrt{2}\sin\theta + 1 = 0.$$

Solving for θ:

$$\theta_1 = 24.5^\circ \quad \text{and} \quad \theta_2 = 155.5^\circ$$

$$\frac{\theta_2}{\theta_1} = \frac{155.5^\circ}{24.5^\circ} = 6.34.$$

This means that the 3-db bandwidth of the hybrid-coupled tunnel-diode amplifier using a coupled $\lambda/4$ strip transmission line hybrid with $K=0.707$ will never exceed a 6.34 to 1 bandwidth. It can also be shown that this hybrid degrades the amplifier by only $\frac{1}{2}$ db over a 3 to 1 frequency range.

It is conceded that the bandwidth figures being mentioned must be modified for practical tunnel-diode networks and amplifiers improperly terminated.

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On Cyclotron-Wave Noise Reduction*

Fast-wave beam noise must be low in Adler tubes[1] to facilitate tuning. Brillouin beams have a high space-charge density and a fairly uniform axial electron velocity. Hence a large amplification[2,3,4] and a low re-introduction of noise can be obtained.[5] Unfortunately, the jump in the magnetic field (Fig. 1) increases the noise temperature.[3] However, our experiment shows that this noise can be reduced without impairing the gun performance seriously. The fast-wave noise power $P_{nv}=kT_{eq}df$ extracted from the beam by the input coupler was

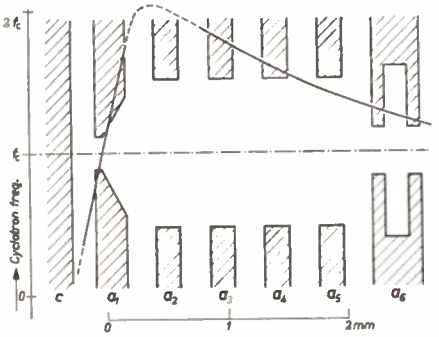


Fig. 1—Brillouin gun and shape of the magnetic field.

TABLE I

Case	a_1 (Volts)	a_2 (Volts)	a_3	a_4	a_5	a_6	i_{a6}	$i_{collector}$ (μa)
1	7.8	260	93	21	7.4	6.0	9.5	32.5
2	7.5	260	120	22.5	6.8	6.0	12	34
3	8.2	54	70	19.3	38	6.0	2	32
4	8.8	135	0	8	14	6.0	8.8	36.2
5	32	−48	290	1	20	6.0	80	30

measured for different potential profiles (Fig. 1 and Table I) as a function of frequency f (Fig. 2). k is Boltzmann's constant, df a small frequency range.

The profile in row 1 (Table I), represents the case of minimum beam-expansion and lens effects while the axial electron velocity u in the region of inconstant magnetic field is large. We see T_{eq} is large. The voltages in row 2 permit more expansion. In rows 3, 4 and 5, u is progressively decreased. We observe a pronounced reduction of T_{eq}. How can we explain this? The change of the noise temperature

$$T_n = \frac{f_c}{f}T_{eq}$$

of a beam is given by:[6]

$$T_{n2} = G\cdot T_{n1} + (1-G)T_e. \qquad (1)$$

Here G is the power gain of the length of beam involved, T_{n1} and T_{n2} are the input and output noise temperatures and T_e is the transverse electron temperature. G as a function of frequency has approximately the shape of a resonance curve. $G\leq 1$, $G\approx 1$ at $f=f_c$. The larger the spread in u, the larger the decrease of G with $|f-f_c|$.[5,6]

Let us divide the gun roughly into three sections, with cyclotron frequency f_{c1}, f_{c2}, f_c,

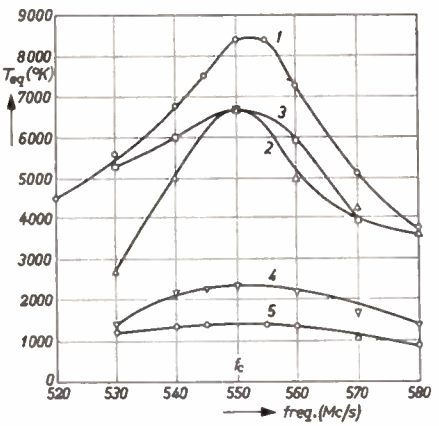


Fig. 2—Equivalent noise temperature. (T_{eq} is not corrected for coupler loss ≈0.3 db at $f=f_c$ and frequency response of the coupler, bandwidth 40 Mc/s.)

power gain G_1, G_2, G_3, electron temperature T_{e1}, T_{e2} and T_{e3}, respectively. Assuming $T_{n1}=T_{e1}$ = cathode temperature T_c, $G_3\approx 1$ for the frequencies of interest, and the jump in the magnetic field adiabatic, *i.e.*, P_{nv} is constant, we obtain with (1)

$$T_{n3} = G_2\frac{f_c}{f_{c1}}T_c + (1-G_2)\frac{f_c}{f_{c2}}T_{e2}. \qquad (2)$$

The output noise

$$P_{nv} = k\frac{f}{f_c}T_{n3}\,df$$

can be made small if G_2 is made small, *i.e.*, if f_{c2} and f_c are far apart and the spread in u in section 2 is large. Since f_{c1} should not be too small, a compromise between Brillouin focussing and low noise may be reached. The second term of (2) can be reduced if T_{e2} is made small, *i.e.*, if the beam is much expanded, if f_{c2} is chosen large, *i.e.*, if section 2 coincides with the peak in the magnetic field (Fig. 1).

* Received by the IRE, October 25, 1961.

[1] R. Adler, *et al.*, "The quadrupole amplifier, a low-noise parametric device," PROC. IRE, vol. 47, pp. 1713–1723; October, 1959.
[2] T. E. Everhart, "Space charge effects in a quadrupole pump," *Proc. Internatl. Mikrowellen Tagung*, Munich, Germany, June 7–11, 1960; Friedr Vieweg & Sohn, Braunschweig, Ger., pp. 350–352; August, 1961.
[3] E. I. Gordon, "Transverse electron beam waves in varying magnetic fields," *Bell Sys. Tech. J.*, vol. 39 pp. 1603–1616; November, 1960.
[4] P. A. H. Hart and C. Weber, "Parametric-amplifier electron gun," *Philips Res. Repts.*, vol. 16, pp. 376–388; December, 1961.
[5] C. P. Lea-Wilson, "Some possible causes of noise in Adler tubes," PROC. IRE (Correspondence), vol. 48, pp. 255–256; February, 1960.
[6] E. I. Gordon, "Noise in beam-type parametric amplifiers," PROC. IRE (Correspondence), vol. 49, p. 1208; July, 1961.

Theoretically the maximum value of $T_{n3} \approx 16{,}000°K$, ($G_2 = 1$, $f_{c1} \approx 50$ Mc/s, $T_c = 1420°K$). If we estimate $T_{e2} = 400°K$ and $f_{c2} \approx 1100$ Mc/s, we obtain from curve 1 in Fig. 2 $G_2 \approx 0.6x$. From other measurements we have the impression that this is reasonable.

The shape of the T_{eq} vs f curve is caused by the frequency response of the coupler. The shift of the peak in T_{eq} from $f = f_c$ (Fig. 2) is probably caused by the transition of f_{c2} to f_c being gradual, unlike the jump in our model.

Finally we can make the following conclusions. Since (2) does not contain T_{e3}, the beam can be refocused in section 3 so that little or no current is lost. There seems to be no reason why T_{n3} cannot be reduced far below cathode temperature, although our gun did not allow it. The slow cyclotron wave is reduced in the same degree as the fast wave. Therefore, the method may also be useful for other tubes involving cyclotron waves, provided f_{c1} and f_{c2} are chosen properly and sufficient expansion occurs.

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Broad-Band Nonuniformly Spaced Arrays*

The radiation pattern for a nonuniformly spaced array is given by

$$F(\theta, k) = \sum_{l=0}^{L} A_l e^{ikx_l \sin\theta}. \quad (1)$$

This radiation pattern is sensitive to frequency by the constant $k = 2\pi/\lambda$ where $\lambda = c/f$. By changing the frequency f the radiation pattern $F(\theta, k)$ will change. The radiation pattern is especially sensitive to changes of frequency if the antenna array consists of equispaced elements, half wavelength apart, as is usually the case.

It has been pointed out by King, *et al.*,[1] that antenna arrays with nonuniform unequally spaced elements are more broad band, and less sensitive to frequency changes.

The change of the radiation pattern $F(\theta, k)$ with respect to frequency can be given by the partial derivative $(\partial/\partial k)F(\theta, k)$. For broad-band purposes, one would like this change to be stationary, *i.e.*, maximum, minimum, or an inflection point. At those points, the changes of the radiation pattern with respect to frequency are the smallest. The condition for $(\partial/\partial k)F(\theta, k)$ to be stationary is that its derivative be zero.

$$\frac{\partial^2}{\partial k^2} F(\theta, k) = 0. \quad (2)$$

It has been shown previously[2,3] that by using the Jacobi expansion

$$e^{iz \sin\theta} = \sum_{n=-\infty}^{+\infty} J_n(z) e^{in\theta}, \quad (3)$$

one could rewrite (1) as

$$F(\theta, k) = \sum_{n=-\infty}^{+\infty} f_n(k) e^{in\theta} \quad (4)$$

$$f_n(k) = \sum_{l=0}^{L} A_l J_n(kx_l). \quad (5)$$

Since (5) is correct for every n, it could be rewritten in matrix form

$$f_n(k)] = [J_n(kx_l)]A_l]. \quad (6)$$

Substituting (1) in the condition for a broad-band array given by (2), one has

$$\frac{\partial^2}{\partial k^2} F(\theta, k) = (i \sin\theta)^2 \sum_{l=0}^{L} A_l x_l^2 e^{ikx_l \sin\theta} = 0. \quad (7)$$

The condition for a broad-band antenna array then becomes

$$\sum_{l=0}^{L} A_l x_l^2 e^{ikx_l \sin\theta} = 0 \quad (8)$$

or in the form

$$\sum_{l=0}^{L} A_l (kx_l)^2 e^{i(kx_l)\sin\theta} = 0. \quad (9)$$

Using the Jacobi expansion in (3), one gets

$$\sum_{l=0}^{L} A_l (kx_l)^2 J_n(kx_l) = 0. \quad (10)$$

Eq. (10) should be required for every n for a broad-band antenna, and therefore could be rewritten in a matrix form,

$$[J_n(kx_l)](kx_l)^2 A_l] = 0. \quad (11)$$

For these $L+1$ homogeneous linear equations to have a nontrivial solution, it requires that the determinant of the coefficients must be zero.

$$|[J_n(kx_l)]| = 0 \quad (12)$$

for $0 \leq n \leq L$ where $L+1$ is the number of elements.

From (12) one can see that once x_0, x_1, x_2, $\cdots$, x_{L-1} are given arbitrarily, x_L will be determined by solving the transcendental (12). Also, in that case A_1/A_0, A_2/A_0, A_3/A_0, $\cdots$, A_L/A_0 will be determined from (11), where A_0 will be arbitrary. Those determined numerical values will give the radiation pattern by substituting in (5) and (4).

It appears as though a trial and error method will be required to design an optimum broad-band antenna which will give a prescribed radiation pattern.

Another approach to the problem will be to design the array for nonuniform spacing and take the determinant in (12) as a criterion for broad-banding. The closer this Bessel function determinant is to zero, the more broad-band the antenna array is. Obviously, much more work should be done on the subject.

One additional important fact should be mentioned; a fact which can simplify our design problem above. If one requires the shortest possible antenna one has from (5),

$$J_n(kx_0) \ll J_n(kx_1) \ll J_n(kx_2) \ll J_n(kx_3) \ll \cdots J_n(kx_L), \quad (13)$$

where $J_n(kx_L)$ has a finite value and all the other array elements are in the decay part of the Bessel function. Using the above fact for other orders of Bessel functions, and substituting zero for the very small values in the matrix, (12) becomes

$$\begin{vmatrix} J_0(kx_0) & J_0(kx_1) \cdots J_0(kx_L) \\ (0) & J_1(kx_1) \cdots J_1(kx_L) \\ 0 & (0) \quad J_2(kx_L) \\ \vdots & \\ 0 & 0 \cdots (0)\ J_n(kx_L) \end{vmatrix} = 0 \quad (14)$$

and one gets a triangular determinant which can be easily solved. Usually the zeros in the brackets are not so small and a more involved calculation will be required.

In the case of a symmetric array, the odd-order Bessel functions in the determinant (12) should be omitted, as can be found from (10).

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A Mechanical Nondegenerate Parametric Oscillator*

It may not be generally realized that a very simple mechanical model of a nondegenerate parametric amplifier or oscillator may be built.

For the oscillator, the model (see Fig. 1) comprises two pendulums coupled by a soft iron electromagnet and pole-piece assembly. The electromagnet rotates with one pendulum, the pole piece with the other. The torque, which tends to align the two pendulums, may be considered as a spring force whose magnitude is controlled by the amount of current in the electromagnet. To see the analogy with the electrical diagram of Fig. 2, one can say that current in circuit No. 1 may either flow through circuit No. 2, or charge the capacitor C_V. The equivalent statement for the mechanical model is that a velocity existing on one pendulum may either couple directly to the other pendulum or cause a relative rotation between the electromagnet and the pole piece. Another viewpoint is that the magnet assembly is analogous to a mechanical differential driv-

* Received by the IRE, November 8, 1961.

[1] D. D. King, *et al.*, "Unequally spaced, broad-band antenna arrays," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 380–384; July, 1960.

[2] H. Unz, "Linear Arrays with Arbitrarily Distributed Elements," Electronics Res. Lab., University of California, Berkeley, Rept. Series No. 60, Issue No. 168; November 2, 1956.

[3] H. Unz, "Linear arrays with arbitrarily distributed elements," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 222–223; March, 1960.

* Received by the IRE, November 13, 1961.

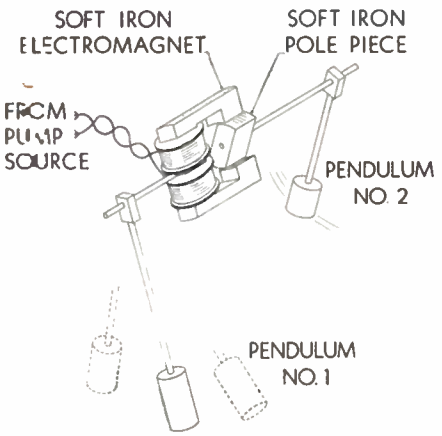


Fig. 1—Mechanical model of a nondegenerate parametric oscillator.

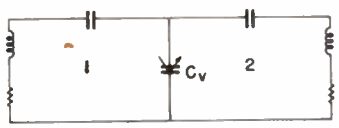


Fig. 2—Equivalent electrical circuit.

ing a modulated spring from the two inputs.

We have built such a model, and by "pumping" the electromagnet at the sum of the resonant frequencies of the two pendulums, sustained oscillations are obtained. In general, the two frequencies are not commensurate. Potentiometers (not shown) are mounted on each shaft, and used to generate signals to drive a multitrack pen recorder. The position of each pendulum may be recorded, and also the pump waveform. A fourth quantity which may be recorded easily is of considerable interest. This is the signal between the two potentiometer rotors. This indicates the relative offset of the electromagnet and pole piece, and this offset is the quantity analogous to the charge on the capacitor C_v.

The amplitudes of oscillation obtained on the two pendulums are in good agreement with the Manley-Rowe relations, that is, the energy delivered to each pendulum is proportional to its frequency. This has been observed over a range of values of viscous (magnetic) damping. The relative loading is determined by measuring the Q. Viscous damping is, of course, the correct equivalent to the resistance in the electrical circuit. Earlier experiments were carried out with frictional damping, and the Manley-Rowe relations were found not to hold for this case.

Danielson[1] in 1959, has proposed, without proof, an explanation of the phase relations in the nondegenerate case when using a rectangular pump waveform. We may adjust the amplitudes of the two pendulums to be equal, in order to duplicate the special case shown in Danielson's Fig. 2. It is interesting that we obtain on the pen recorder exact agreement with the waveforms and phase relations shown in Danielson's Fig. 2(c) and 2(d).

[1] W. E. Danielson, "Low noise in solid state parametric amplifiers at microwave frequencies," *J. Appl. Phys.*, vol. 30, pp. 8–15; January, 1959.

Establishment of the correct phase relationships in the model may be visualized as follows: the difference between the pump waveform and an assumed oscillation on either pendulum comprises a forcing function of specific phase for the other pendulum. Since this forcing function is at the resonant frequency of the other pendulum, the resulting oscillation is in phase with it.

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A Proposed Test for the Existence of a Lorentz-Invariant Aether*

Since the inception of the special theory of relativity more than half a century ago, many investigators have sought to overcome the obvious irreconcilability of its two postulates, and to obtain thereby some consistent explanation of the general phenomena. Einstein's first postulate together with his arbitrary definition of time [1], requires that all inertial frames be isotropic, particularly with respect to the propagation of light. This concept demands that light be propagated isotropically with respect to its source alone, a requirement upon which Dingle has based a new theory of special relativity [2], [3]. That this is also Einstein's viewpoint can be verified through his statement [4]. "What is essential is, that the electric and magnetic force of the light which is influenced by a moving body, be transformed into a system of coordinates at rest relatively to the body. By this means all problems in the optics of moving bodies will be reduced to a series of problems in the optics of stationary bodies."

Unfortunately, the phenomena are not so simple as given by this hypothesis, and in view of two recent electro-optical experiments [5], [6], this postulate must now be abandoned, along with its explanations for the symmetrical phenomena, the Michelson-Morley experiments, the Townes-Cedarholm maser experiments, Ruderfer's experiment [7], Fizeau's results with moving liquids, etc. Consequently, for experiments of this type, special relativity now has no explanation at all [8].

Einstein's remaining postulate, taken alone, requires a classical electromagnetic aether: In the isotropic reference frame of the inertial observer, the velocity of light, (taken one way), is the universal constant c. I have listed in my previous paper [8] several methods whereby this postulate can be unambiguously tested. The postulate is generally accepted by physicists, along with the Larmor-Lorentz hypotheses [9], and is used to explain the so-called asymmetrical phenomena of nature. The asymmetrists, perhaps unconsciously, employ a Lorentz-invariant aether that supposedly causes a measurable time dilation and a measurable Lorentz contraction in a noninertial system, neither of which effect is valid without the other. Because time dilation has supposedly been demonstrated [10] (I have shown elsewhere that this belief is false [11]), while all attempts to measure the Lorentz contraction have failed, the presently accepted viewpoint turns out to be paradoxical and is generally referred to as such.

Since Einstein's first and second postulates are incompatible, the first requiring a constant velocity for light with respect to its source alone, and the second a constant velocity with respect to the receiver alone, neither postulate seems capable of telling more than a half truth. A new mechanics is needed that, based upon a single postulate, will explain all of the general phenomena without involving paradoxes. Such a mechanics appears now to have been found [8], and preliminary experiments have verified its correctness. The purpose of this communication is to describe one such test and to suggest a further, crucial experiment that will test the basic consequence of the newly extended Newtonian relativity: that the velocity of light with respect to an inertial observer is not a constant as with Einstein's theory, but instead is $c(1+V^2/c^2)^{1/2}$, wherein V is the purely relative velocity of the source. The preliminary test is with Čerenkov light, and demonstrates directly that the periodic properties of light are propagated with the velocity C_{mn},[8], which equals $c(1-V^2/C_{mn}^2)^{-1/2}$. In Fig. 1, the observer, stationed arbitrarily in the y frame, notices that light, which originates because of the super-light velocity V of electrons in the x frame, reaches him at maximum intensity at the angle θ with respect to the line of action. By the Huygens' construction, reinforcement of wavelets occurs along the wavefront shown upon the assumption that the velocity of light in the medium of refractive index, n, is C^2_{xy}/n, so that

$$\text{Cos}\,\theta = C_{xy}/nV = c/n\alpha_{xy}V \qquad (1)$$

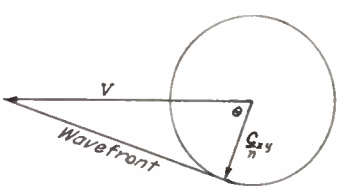


Fig. 1—Čerenkov light.

where $\alpha_{xy} \equiv (1-V^2/C_{xy}^2)^{1/2} \equiv (1+V^2/c^2)^{-1/2}$. Now it can be shown that the limit of the quantity $\alpha xy V$, as V is allowed to increase without limit is c. Hence,

$$\lim_{V\to\infty} \cos\theta = 1/n. \qquad (2)$$

This consequence of Newtonian relativity is in precise agreement with our experiments. As the velocity of the electrons used is increased, becoming infinite in the limit, $\cos\theta$ becomes $1/n$.

* Received by the IRE, October 20, 1961.

However, under Einstein's second postulate, the Huygens' construction yields

$$\cos\theta = c/nV, \tag{3}$$

wherein no limit is placed upon the velocity V either by the postulate or by the Huygens' construction. Hence, under relativity,

$$\lim_{V\to\infty} \cos\theta = 0 \tag{4}$$

so that the maximum possible angle for Čerenkov emission becomes $\pi/2$, which contradicts experience. Thus it is seen that Einstein's second postulate leads to asymmetries that are not inherent to the phenomenon. This failure, together with the failure of the asymmetrists to offer any explanation at all for the symmetrical phenomena, formerly explained under Einstein's now invalidated first postulate, indicates, as claimed by Palacios [11], that the second postulate, too, is invalid.

A crucial test of the group velocity of light remains to be conducted. Suppose, by way of example, that electrons of 0.5 bev energy are available and that they are caused to emit a pulse of Čerenkov light. At the source, a portion of this can be immediately intercepted by a photomultiplier tube, which in turn triggers a strobe flash with a known time delay, for example, 100 nsec. The Čerenkov pulse, together with the strobe pulse, is allowed to travel a convenient distance, such as 100 feet, to a point of detection. The Čerenkov pulse, traveling at C_{xy} velocity, which at 0.5 bev amounts to $1000c$, will traverse this path in one-tenth nsec, or almost instantaneously. However, the strobe pulse, traveling at c velocity, will require 100 nsec, and since it was originally emitted 100 nsec late, it will arrive about 200 nsec after the Čerenkov pulse.

At the point of detection, a photomultiplier detector can be used, feeding into a 100-megacycle universal counter-timer.[1] The latter consists of two basic parts: an electronic gate, which is opened by the first pulse to arrive and which is closed by the second pulse; and a 100-Mc timer, which measures accurately the duration of the open position of the gate. An interval of time as short as 10 nsec can be measured by this means, which is more than sufficient sensitivity for the purpose. In fact, the measurement of C_{xy} to two significant figures appears possible in the range of effective electron energies between 0.51 and 5.1 mev. Within this range, if the predictions of the extended Newtonian relativity are verified, Čerenkov light will travel from 1 to 10 times as fast as ordinary light.

If, in fact, forms of light exist that travel with superlight velocities, this knowledge will have a profound influence upon communications of the future. The time for an exchange of messages between us and the nearest possible stellar systems supporting life could be cut from 20 years to a matter of months or even days. There would be no theoretical limit to the velocities that space vehicles of the future could attain, and communication with such fast moving objects would be facilitated by their motion, which would cause their radio waves to be propagated with superlight velocities.

[1] Computer Measurements Co., Sylmar, Calif.

Regardless of such possible applications in the distant future, it is of scientific interest to measure the velocity of light to the second order, one way, with the same high degree of accuracy that the average velocity of the two-way signal has been measured to the first order. It should be quite obvious to all that outside of the Čerenkov measurements I have just cited, the velocity of light from a relatively moving source has never been tested to the second order.

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BIBLIOGRAPHY

[1] A. Einstein, "On the electrodynamics of moving bodies," in "Principle of Relativity," Dover Publications, Inc., New York, N. Y., p. 40; 1923.
[2] H. Dingle, "Relativity and electromagnetism: an epistemological appraisal," *Phil. Sci.*, vol. 27, pp. 233–253; July, 1960.
[3] ——, "A possible experimental test of Einstein's 2nd. postulate," *Nature*, vol. 183, p. 1761; June, 1959.
[4] Einstein, *op. cit.*, p. 59.
[5] O. Heckman, "The aberration of extragalactic nebulae," *Ann. Astrophys.*, vol. 23, pp. 410–415; 1960.
[6] A. M. Bonch-Bruevich, "A direct experimental confirmation of the 2nd. postulate of special relativity," *Optics and Spectro.*, vol. 9, p. 73; July, 1960.
[7] M. Ruderfer, "First order ether drift experiment using Mossbauer radiation," *Phys. Rev. Lett.*, vol. 5, pp. 191–192; September, 1960.
[8] P. M. Rapier, "An extension of Newtonian relativity to include electromagnetic phenomena," PROC. IRE, vol. 49, pp. 1691–1692; November, 1961.
[9] H. E. Ives, "The Doppler effect considered in relation to the Michelson-Morley experiment," *J. Opt. Soc. Am.*, vol. 27, pp. 389–394; October, 1937.
[10] C. W. Sherwin, "Some recent experimental tests of the clock paradox," *Phys. Rev.*, vol. 120, pp. 17–21; October 1, 1960.
[11] J. Palacios, "A reappraisal of the principle of relativity as applied to moving interferometers," *Rev. Roy. Acad. Exact Sci. Madrid*, vol. 55, pp. 3–14; April, 1961. (In English.)

Propagation in Laser Crystals*

Schawlow and Townes[1] in their article on optical masers, show, in theory, how the basic geometry of a cylinder with flat parallel reflecting ends can support a sufficiently loss-free, multiple-reflection mode. More recently theoretical papers that study internal laser propagation have appeared[2,3] in which highly idealized models have been employed.

In a real system, the emitted beam can be expected to be at most partially coherent and to have a nonuniform amplitude and phase distribution across the emitting aperture. The resulting diffraction pattern can deviate appreciably from that predicted by assuming coherent uniform illumination. For example, we show the diffraction pattern[4]

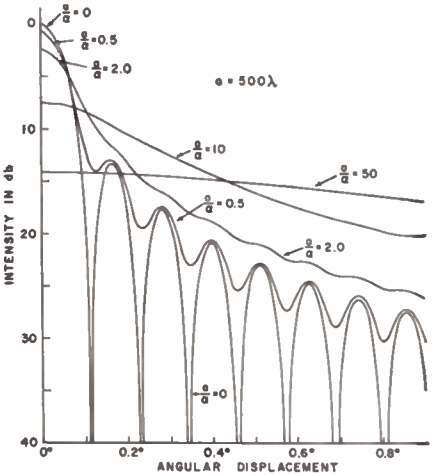


Fig. 1—Diffraction of partially coherent light (slit aperture).

produced by a slit aperature of width a illuminated by partially coherent radiation in Fig. 1. The coherence of illumination Γ_{12} was taken to be of the form $\Gamma_{12}=\exp(-|x_1-x_2|/\alpha)$. Here α, the correlation interval, is the separation of two points x_i such that the coherence between the disturbances at those points is equal to $1/e$. Thus $\Gamma_{12}=1$ implies perfect coherence. Although the illumination is taken as uniform, the nulls vanish almost immediately as the coherence is decreased, and the beamwidth spreads until in the limit the source is incoherent or Lambertian. Of course, any amplitude and phase distribution across the face of an emitting aperture will further complicate the far-field diffraction pattern.

Although theoretical results published to date are limited to idealized systems, our experimental data on various laser rods indicate a cooperative effect across the entire rod—in other words, an internal multiple reflection field characterized by a coherence function. This is shown by a far-field photograph of the laser signal from the entire aperture of a 0.25-in diameter ruby laser rod (Fig. 2). The dynamic range of photographic film (some 30 db) is not sufficient to show the outlying rings without saturation in the region of the central maxima (photographs using attenuating media complete the pattern). Any internal field asymmetry caused by imperfections adds an additional interference effect characterized in general by a modulation of the far-field pattern which is only slight in Fig. 2.

Prediction of far-field ring patterns is under study. In the meantime, it is noted that a simple physical optics argument as to their origin is not obvious. Although both the intensity and contrast of the rings decrease with ring radius in a manner expected of diffraction of partially coherent light by a circular aperture, ring radius varies approximately as $\sqrt{n}$ (characteristic of Haidinger interferometer fringes). Clarification requires a solution of the complete diffraction problem in a laser rod geometry.

Further evidence to support the argument that the coherence interval generally exceeds the laser rod diameter may be obtained by placing a two slit aperture at the

* Received by the IRE, November 14, 1961; revised manuscript received, November 20, 1961.
[1] A. L. Shawlow and C. H. Townes, "Infrared and optical masers," *Phys. Rev.*, vol. 29, pp. 1940–1949; December, 1958.
[2] A. G. Fox and Tingye Li, "Resonant modes in a maser interferometer," *Bell Sys. Tech. J.*, vol. 40, pp. 453–488; March, 1961.
[3] J. Kotik and M. C. Newstein, "Theory of laser oscillations in Fabry-Perot resonators," *J. Appl. Phys.*, vol. 32, pp. 178–186; February, 1961.
[4] G. B. Parrent, Jr., and T. J. Skinner, "Diffraction of partially coherent light by a plane aperture," *Optica Acta*, vol. 8, pp. 93–102; January, 1961.

eld of a rod. This is essentially the same experiment reported by Nelson and Collins[5] in which slits ~100 λ apart were used. We find that interference between two 0.005 in by 0.125-in slits spaced by 0.125 in (~3000 λ) is readily observed. This is illustrated in Fig. 3, where three features bear noting. First, the gross structure of which only the central lobe is shown here is the pattern produced independently by the non-uniformly illuminated rectangular slits. Secondly, internal to each lobe are fringes



Fig. 2—Far-field pattern (entire laser aperture).

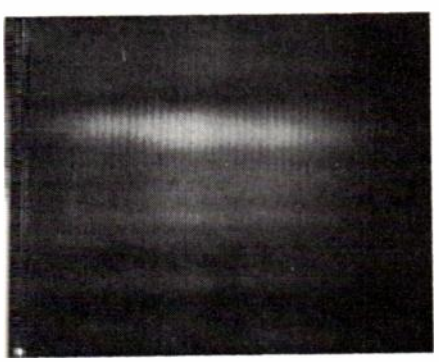


Fig. 3—Diffraction of partially coherent laser light (two-slit aperture).

of the correct spacing caused by the cosine modulation which is due to the slit separation. Thirdly, since the slits are finite, this pattern is reproduced above and below the central pattern.

It should be pointed out that while clean ring patterns, such as Fig. 2, indicate spatial coherence, across the face of the laser rod, their absence does not imply absence of coherence. In some samples the internal field asymmetry may be bad enough to destroy the ring structure, yet coherence may most certainly exist across such an apodized wave front. For example, the fringe pattern of Fig. 3 was produced by partially coherent light obtained from widely separated slits near the end of a rod which from its entire aperture produces a far-field pattern lacking ring structure. Furthermore, each of the many ruby samples tested in this laboratory yield high contrast interference fringes as in Fig. 3 whether or not they produce a ring pattern as in Fig. 2.

In this connection, by a series of experiments presently underway,[6] it has been demonstrated that the illumination of slits by quasi-monochromatic radiation does not generally produce time-integrated interference fringes. As expected theoretically, it also is necessary that the radiation have a degree of coherence consistent with slit spacing. These experiments employ parallel laser rods that are mounted so that their oscillations are coupled. Noticeable interference between the laser beams from the rods occurs only when the coupling is great enough to produce a sufficient amount of correlation between them.

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[5] D. F. Nelson and R. J. Collins, "Special coherence in the output of an optical maser," *J. Appl. Phys.*, vol. 32, pp. 739–740; April, 1961.

[6] J. I. Masters, G. B. Parrent, Jr., and T. J. Skinner, "Partial Coherence and Coupled Laser Systems," presented at NEREM, Boston, Mass., November 14, 1961.

A Classical Red-Shift*

If a sinusoidal oscillator S is in a state of uniform, translational motion $(0<v/c\ll 1)$ relative to an observer P, the waveform of this oscillation at P is represented by a frequency-modulated sinusoid whose amplitude is constant if the inverse-distance law is neglected. For a given time interval the amount of frequency deviation of this modulation is largest as the motion of S is "transverse" relative to P.[1]

For computational convenience let the sinusoid at S have a frequency $\nu_s=1/T_s=30$ *cps*. Let us assume a duration $\tau=20$ sec of the wave train at P. Outside this time interval the amplitude of the wave is zero. Let τ be centered upon the instant of closest approach between S and P. The terminated, frequency-modulated wave at P can be represented by

$$f(t)=\sin\left\{2\pi\Delta t\left(\frac{1}{T_s}+\frac{v^2}{2\pi\lambda_s}\cdot\frac{\left(\frac{\tau}{2}-\Delta t\right)}{\sqrt{R_m^2+v^2\left(\frac{\tau}{2}-\Delta t\right)^2}}\right)\right\},\quad(1)$$

where

$$f(t)=0 \text{ when } |t|>\frac{\tau}{2}.$$

$\lambda_s=c/\nu_s$, R_m is the minimum range between S and P. Let Δt increase from zero to τ in $(150)^{-1}$ second increments. The function $f(t)$ is fed into an IBM-704 computer and its Fourier-integral-transform is evaluated. Fig. 1 illustrates the results of this computation.

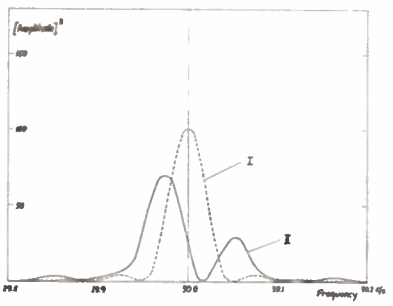


Fig. 1—Power spectrum for the "transverse" Doppler effect with wide acceptance angle at the observer.
I) $v=0$, II) $v=1000$ km, $R_m\ll\infty$.

Curve I represents the well-known spectral-power distribution of a terminated sinusoid $(v=0)$. Curve II is obtained for $v=1000$ km and $R_m=0.01$ km. This distribution reveals the presence of three effects due to the relative motion between S and P; a broadening, a splitting, and a red-shift. These effects are a consequence of the classical Doppler principle. If the amplitude of the oscillation at P varies in time the spectral power distributions will be different from those shown in Fig. 1, but a difference between cases I and II will remain.

For radio emissions from artificial earth satellites, because of relatively large ranges, small speeds, and low frequencies, all three effects are of negligible magnitude. In optical investigations, however, they become important for high speeds and close ranges of atomic emitters, as will be demonstrated elsewhere.

The phenomenon of frequency modulation discussed here follows from the consideration of a finite angular velocity between S and P. Usually, experimenters restrict the incremental Doppler effect to zero angular-velocity between S and P either because of existing conditions in nature, or, more frequently, by an implicit assumption, although zero angular-velocity is strictly obtained only if either a) S and P lie on a linear trajectory of motion, or b) the range between S and P is infinite.

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* Received by the IRE, November 24, 1961.

[1] K. Toman, "Minimum range to artificial earth satellites," *Nature*, vol. 190, pp. 333–334; April, 1961.

Explicit Forms for the Conductivity and Permittivity of Bulk Semiconductors in Waveguides*

Recent interest in the analysis of waveguides containing bulk semiconductors has been motivated by two different possible applications: 1) microwave measurement of electron transport phenomena[1-3] and 2) development of bulk semiconductor microwave devices.[4,5] In the first application, one desires to find the conductivity and/or permittivity (or changes in these quantities caused by carrier injection, dc electric and magnetic fields, etc.) from measurement of the transmission or reflection coefficient of a waveguide containing the semiconductor. For the second application, the inverse problem is generally of interest. The discussion in this note will be confined to the first application, the measurement problem.

Consider TE waves propagating in a waveguide that is perturbed by a semiconductor of arbitrary but uniform cross section between $Z=0$ and $Z=d$. In general normal modes of the perturbed and unperturbed sections will differ, resulting in an infinite series of modes excited at both discontinuities. If, however, mode solutions in the two regions are so similar that higher-order terms are negligible, the transmission coefficient of the principal mode is given by[2]

$$\frac{4\gamma_1\gamma_2}{(\gamma_2+\gamma_1)^2e^{\gamma_2 d}-(\gamma_2-\gamma_1)^2e^{-\gamma_2 d}} = \frac{E_{\text{Trans.}}}{E_{\text{inc.}}} = e^{-(A+j\phi)}, \quad (1)$$

where $\gamma_2=\alpha_2+j\beta_2$ and $\gamma_1=j\beta_1$ are the propagation constants of the principal mode in the perturbed and unperturbed regions, respectively.

We have programmed a digital computer to invert (1) by iterative interpolation. When the computer is presented measured values of A, ϕ, and $\beta_1 d$, it calculates $\alpha_2 d$ and $\beta_2 d$ to within 0.1 per cent in about five seconds.

Under the same conditions that (1) applies, σ and ϵ_r can be found explicitly in terms of α_2, β_2, and β_1 by perturbation methods. This calculation assumes nothing about the relative magnitudes of σ and $\omega\epsilon_r$ and thus applies to both high- and low-loss cases. We will now consider several examples.

Formal Solution of Completely Filled Waveguide

With the semiconductor completely filling the waveguide, the perturbed and unperturbed normal modes are identical and (1) is exact. The desired explicit forms can be derived from the formal equations for the propagation constants

$$\gamma_1^2 = \{2\pi/\lambda_c\}^2 - \omega^2\mu_0\epsilon_0$$
$$\gamma_2^2 = \{2\pi/\lambda_c\}^2 - \omega^2\mu_0\epsilon_0\hat{\epsilon}_r, \quad (2)$$

where $\hat{\epsilon}_r=(\epsilon_r-j\sigma/\omega\epsilon_0)$ is the complex relative permittivity of the semiconductor.

Subtracting,

$$(\gamma_1^2-\gamma_2^2) = \omega^2\mu_0\epsilon_0(\hat{\epsilon}_r-1), \quad (3)$$

and separating into real and imaginary parts yields

$$\sigma = \frac{2\alpha_2\beta_2}{\omega\mu_0},$$
$$(\epsilon_r-1) = \frac{\beta_2^2-\alpha_2^2-\beta_1^2}{\omega^2\mu_0\epsilon_0}. \quad (4)$$

These equations are exact and show that knowledge of *both* α_2 and β_2 is in general required to find *either* σ or ϵ_r.

The Perturbation Formula

In the field distributions of a perturbed mode $E_2(x, y)$, $H_2(x, y)$, and the corresponding unperturbed mode $E_1(x, y)$, $H_1(x, y)$ are known, the difference between their propagation constants can be calculated with the well-known formula[6]

$$\frac{\gamma_2-\gamma_1}{j\omega\epsilon_0} = \frac{\int(\hat{\epsilon}_r-1)E_2\cdot E_1^*dS}{\int(E_1^*\times H_2)\cdot\hat{k}dS+\int(E_2\times H_1^*)\cdot\hat{k}dS} \quad (5)$$

by integrating over the cross section of the waveguide. Eq. (5) is exact; however, E_2 and H_2 are generally not known exactly, requiring approximations to evaluate the integrals.

The Completely Filled Waveguide

Because fields of the perturbed and unperturbed modes have the same spatial variation for the completely filled waveguide, (5) can be evaluated exactly for this case. The fields of any perturbed mode are written

$$E_2 = E_1$$
$$H_2 = (\gamma_2/\gamma_1)H_1. \quad (6)$$

Thus, by noting that

$$\frac{\int E_1\cdot E_1^*dS}{\int(E_1\times H_1^*)\cdot\hat{k}dS} = j\frac{\omega\mu_0}{\gamma_1}, \quad (7)$$

one obtains again the results given in (4).

TE_{10} Mode, Thin Sample in Center of Waveguide

As an example of an approximate solution, consider a planar sample of thickness t extending from top to bottom in the center of a standard TE_{10} mode waveguide of width a, parallel to the electric field. The perturbed and unperturbed normal modes differ for this case; however, if the sample is sufficiently thin, (6) is a first order approximation to the perturbed TE_{10} mode fields. The conditions under which (6) applies are[7]

$$t \ll (2a/\pi), \quad (8)$$
$$t\,|\,\hat{\epsilon}_r-1\,| \ll (\pi/a)(2/\omega^2\mu_0\epsilon_0). \quad (9)$$

The integrand in the numerator of (5) only has value over the sample. Substituting (6) into (5) and noting that

$$\frac{\int(\hat{\epsilon}_r-1)E_1\cdot E_1^*dS}{\int(E_1\times H_1^*)\cdot\hat{k}dS} = j\left(\frac{2t}{a}\right)\left(\frac{\omega\mu_0}{\gamma_1}\right)(\hat{\epsilon}_r-1), \quad (10)$$

yields

$$\sigma = \left(\frac{a}{2t}\right)\left(\frac{2\alpha_2\beta_2}{\omega\mu_0}\right)$$
$$(\epsilon_r-1) = \left(\frac{a}{2t}\right)\left(\frac{\beta_2^2-\alpha_2^2-\beta_1^2}{\omega^2\mu_0\epsilon_0}\right). \quad (11)$$

The equations in (11) contain no assumptions concerning the relative magnitudes of the conduction and displacement currents and are valid as long as (8) and (9) are satisfied. Eq. (9) is usually the more stringent condition and requires samples much less than $\frac{1}{2}$ mm thick for germanium at X band. This condition may actually be more severe than is usually necessary. We have performed experiments with both sides of (9) of the same order of magnitude which yielded microwave measurements of σ and ϵ_r differing by only a few per cent from values obtained by other means.

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[1] A. P. Ramsa, H. Jacobs and F. A. Brand, "Microwave techniques in measurement of lifetime in germanium," *J. Appl. Phys.*, vol. 30, pp. 1054–1060; July, 1959.

[2] H. Jacobs, *et al.*, "Electrodeless measurement of semiconductor resistivity at microwave frequencies," PROC. IRE, vol. 49, pp. 928–932; May, 1961.

[3] H. A. Atwater, "Microwave determination of semiconductor-carrier lifetimes," PROC. IRE, vol. 49, pp. 1440–1441; September, 1961.

[4] H. Jacobs, *et al.*, "A new semiconductor microwave modulator," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 553–559; September, 1960.

[5] H. Jacobs, *et al.*, "Multiple reflections of microwaves propagating through a semiconductor medium," PROC. IRE, vol. 49, pp. 1683–1684; November, 1961.

[6] H. Suhl and L. R. Walker, "Topics in guided wave propagation through gyromagnetic media, Part III—Perturbation theory and miscellaneous results," *Bell Sys. Tech. J.*, vol. 33, pp. 1133–1194; September, 1954.

[7] D. B. Armstrong, "Waveguide perturbation technique for the measurement of microwave conductivity and permittivity of thin semiconductor samples," M.S. thesis, University of Minnesota, Minneapolis; 1961.

Contributors



S. Caputo

Salvatore Caputo was born in Bayonne, N. J., on February 9, 1932. He graduated from RCA Institutes, New York, N. Y., in 1959.

He served in the United States Air Force from 1951 to 1955, where he was an Instructor on microwave radio relay systems and radio teletype communications. He has been an employee of Bell Telephone Laboratories, Murray Hill, N. J., since March, 1959, where he has been working on the development of data communications techniques and equipment.

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R. I. B. Cooper

R. I. B. Cooper graduated in physics from Trinity College, Cambridge, England, in 1942, and did national service during the war at the Radar Establishment in Malvern. He returned to Cambridge, in 1946 and took his Doctor's degree, in 1948 for work on submarine gravity surveying.

Between 1946 and 1953 he took part in three submarine expeditions. Since 1954 he has worked in the electrical field, first in research on contacts and later in semiconductor technology. He is now Head of Science at the Cambridgeshire College of Arts and Technology, Cambridge, England. Previously he was Head of Controls Section in the Transistor Division of S.T.C. Ltd., England.

Mr. Cooper is a Fellow of the Royal Astronomical Society, and the Institute of Physics, and an associate member of the IEE.

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W. O. Eckhardt

Wilfried O. Eckhardt was born in Frankfurt/Main, Germany, on March 30, 1928. He received the Diploma in physics from the University of Frankfurt, Germany, in 1952, and the Ph.D. degree in physics from the same school, in 1958.

From 1953 to 1957 he was an Assistant at the Physical Institute of the University of Frankfurt; he did research in the fields of microwave and far infrared spectroscopy during that time. From 1958 to 1960 he worked on problems of parametric amplification, ultra-high-speed logic, and microwave light modulation at the RCA Research Laboratories, Princeton, N. J. In 1960 he joined the Hughes Research Laboratories, Malibu, Calif., where he is now engaged in research on plasma propulsion. He has published a number of papers on the subjects mentioned above.

Dr. Eckhardt is a member of the German Physical Society.

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H. D. Helms
S'55–M'60

Howard D. Helms was born in Chicago, Ill., on July 30, 1934. He received the B.S. degree in electrical engineering-physics and the Ph.D. degree in electrical engineering from Princeton University, N. J., in 1956 and 1961, respectively.

In 1956 and 1958 he worked as a summer employee at the RCA Laboratories, Princeton, N. J. From 1959 through 1960 he worked on information theory and sampling expansions as a member of the Acoustical and Visual Research Department of Bell Telephone Laboratories, Murray Hill, N. J. Since 1961 he has been with the Military Analysis Department of Bell Telephone Laboratories, Whippany, N. J., where he has been concerned with problems of acoustical transducers, nonjammable communications systems, and errors arising in certain sampled data systems.

Mr. Helms is a member of Sigma Xi and Phi Beta Kappa.



W. H. Highleyman
S'55–M'58

Wilbur H. Highleyman was born in Kansas City, Mo., on August 20, 1933. He received the B.E.E. degree, in 1955, from Rensselaer Polytechnic Institute, Troy, N. Y., the M.S. degree, in 1957, from Massachusetts Institute of Technology, Cambridge, and the D.E.E. degree, in 1961, from the Polytechnic Institute of Brooklyn, N. Y.

From 1955 to 1956 he was a Research Assistant at M.I.T. Lincoln Laboratory, Lexington, developing transistor circuits for digital computers. The following year, he continued work in transistor circuits and thin ferromagnetic films at M.I.T. on a fellowship from Melpar, Inc. In 1958 he joined the Bell Telephone Laboratories, Murray Hill, N. J., where he was engaged in the problem of character recognition until 1959. Since then, he has been concerned with the development of data communication equipment and the study of new devices and techniques for data communication problems. He is presently a Lecturer at the Polytechnic Institute of Brooklyn, N. Y.

Dr. Highleyman is a member of Tau Beta Pi and Eta Kappa Nu, and an associate member of Sigma Xi.

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C. Hilsum

Cyril Hilsum was born in London, England, on May 17, 1925. He received the B.Sc. degree in physics from University College, London, in 1945, and the Ph.D. degree from London University, in 1959.

In 1945 he joined the infrared group at the Admiralty Research Laboratory, Teddington, England, where he was engaged in research on infrared photocells and atmospheric transmission. Since 1950 he has been with the Services Electronics Research Laboratory, Baldock, studying compound semiconductors and is currently leader of the semiconductor group.

Dr. Hilsum is a Fellow of the Institute of Physics.



B. M. OLIVER
S'40–A'40–M'46–SM'53–F'54

B. M. Oliver was born in Santa Cruz, Calif., on May 27, 1916. He received the B.A. degree in electrical engineering from Stanford University, Calif., in 1935, and the M.S. degree from the California Institute of Technology, Pasadena, in 1936. Following a year of study in Germany under an exchange scholarship, he returned to the California Institute of Technology where he received the Ph.D. degree, in 1940.

From 1939 to 1952 he was employed at the Bell Telephone Laboratories, Murray Hill, N. J., in television research and radar development. He is now Vice President in charge of Research and Development at the Hewlett-Packard Company, Palo Alto, Calif.

Dr. Oliver is a member of the American Astronautical Society.

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J. E. ROWE
S'51–M'56–SM'61

Joseph E. Rowe was born in Highland Park, Mich., on June 4, 1927. He received the B.S. degree in electrical engineering and mathematics, in 1951, and the M.S. and Ph.D. degrees in electrical engineering, in 1952 and 1955, respectively, all from The University of Michigan, Ann Arbor.

Since 1951 he has been associated with The University of Michigan Office of Research Administration, engaging in fundamental research on microwave systems, microwave devices, noise theory, electromagnetic field theory, and plasmas. Formerly a Lecturer, Assistant Professor, and Associate Professor of electrical engineering, he is now Professor of electrical engineering and Director of the Electron Physics Laboratory at The University of Michigan.

Dr. Rowe is a member of the AIEE, the American Mathematical Society, the Society for Industrial and Applied Mathematics, and the American Physical Society. He is also a member of Sigma Xi, Phi Kappa Phi, Tau Beta Pi, and Eta Kappa Nu.

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F. STERZER
M'56

Fred Sterzer was born in Vienna, Austria, on November 18, 1929. He received the B.S. degree in physics from the College of the City of New York, N. Y., in 1951, and the M.S. and Ph.D. degrees from New York University, N. Y., in 1952 and 1955, respectively.

From 1952 to 1953 he was employed by the Allied Control Corporation, New York, N. Y. During 1953 and 1954 he was an Instructor in physics at the Newark College of Engineering, N. J., and a Research Assistant at New York University. He joined the RCA Tube Division, Harrison, N. J., in October, 1954, and transferred to the Princeton, N. J., branch in 1956, where he is now Group Leader in microwave physics. His work has been in the field of microwave spectroscopy, traveling-wave tubes, backward-wave oscillators, solid-state microwave amplifiers and oscillators, and microwave computing circuits.

Dr. Sterzer is a member of Phi Beta Kappa, Sigma Xi and the American Physical Society.



J. B. THOMAS
S'52–M'56–SM'60

John B. Thomas was born in New Kensington, Pa., on July 14, 1925. He received the A.B. degree from Gettysburg College, Gettysburg, Pa., in 1944, the B.S. degree from The Johns Hopkins University, Baltimore, Md., in 1952, and the M.S. and Ph.D. degrees from Stanford University, Calif., in 1952 and 1955, respectively.

During World War II he served in the U. S. Army. From 1946 to 1952 he was employed by Koppers Company, Inc., Baltimore, Md., in the Industrial Gas Cleaning Department, first as an Electrical Engineer and then as Assistant Chief Engineer of the department. Since 1955 he has been a member of the Department of Electrical Engineering at Princeton University, N. J., and is currently an Associate Professor of that department.

Dr. Thomas is a member of the AIEE, Tau Beta Pi, and Sigma Xi.

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C. Y. WRIGLEY

Charles Y. Wrigley was born in Bronxville, N. Y., on September 27, 1935. He received the B.A. degree in physics, in 1957 from Williams College, Williamstown, Mass., where he remained as a Graduate Assistant until 1958.

He then joined the Transistor Laboratory in the Research Department of the Sprague Electric Company, North Adams, Mass. He is now studying part-time at Williams College, and is engaged in the investigation of properties of transistors and other solid-state devices.

Books

Colour Television: N.T.S.C. System, Principles and Practice, by P. S. Carnt and G. B. Townsend

Published (1961) by Iliffe Books Ltd., Dorset House, Stamford St., London S.E. 1, England. 433 pages+7 index pages+2 bibliography pages+45 appendix pages. Illus. 5¾×8¾. 85s.

This excellent book describes the adaptation of the American N.T.S.C. color television system to the British 405-line standards. Written primarily for British readers, it is recommended to all those concerned with color television practice.

The authors, members of the technical staff of the British General Electric Company, have written clear accounts of the principles and practice of the N.T.S.C. system, and the properties of human vision on which its techniques are based. Throughout the book, the relationships between colorimetric quantities and their electrical counterparts are briefly but clearly reduced to explicit numerical form. Familiarity with monochrome television practice is assumed, but from this base the reader may safely embark on his own.

American and Continental European readers will find the numbers relating to frequencies unfamiliar, since these have been referred to the British standards (the color subcarrier frequency is 2.6578125 mc and the sight-sound carrier interval is 3.54375 mc). The authors have wisely refrained from confusing everyone by attempting to deal with the versions of these frequencies applicable to the 525-line and 625-line standards.

The book is comprehensive in scope. It treats colorimetry from first principles, and proceeds to the detailed effects of gamma correction on hue and saturation loci on the color triangle. Terminal equipment, including cameras and three types of picture tube (shadow-mask, index-sensing, and switching-grid), is described in some detail. One chapter is devoted to the intricacies of transmitter coding (forming the color subcarrier and burst) and another to specific coding circuits. The major portion of the book, some eight chapters, is devoted to receiver design, installation, test and trouble shooting.

Of eight appendices giving mathematical details, no fewer than six are concerned with particular aspects of the color-sync burst and its use in phase detection of chrominance values. This perhaps disproportionate attention to one phase of receiver design, reflects the concern of British and European engineers over the accuracy required for phase reference in the N.T.S.C. system. The authors admit that they were once numbered among the doubters that a 5 or 10 degree phase lock (equivalent to a timing error of less than 10 nanoseconds, British style) could be achieved in a widespread public service. Experience has prompted them to state that in practice, "generation of the *c-w* reference from the burst signal is practicable even under extremely adverse signal-to-noise ratio conditions . . . even when the the signal-to-noise ratio is below useable level for the monochrome signal."

Aside from its excellent organization and clear exposition, this book will be received with special interest by the 315 American engineers who participated in the deliberations of the N.T.S.C., because the book constitutes a rather complete endorsement of the principles of compatible color transmission developed by them. No set of electronic standards has ever undergone a more thorough examination for defects than did the N.T.S.C. system, at the hands of analysts on England and on the Continent. The last chapter, "Shortcomings of N.T.S.C. Systems," summarizes this study. The principal findings relate to the presently inescapable effects of gamma correction on the accuracy of the reproduced color values. But these are minor matters, and N.T.S.C. pictures are stated by the authors to have a quality "rather better than that obtained with good color photographs." They point out that the worst inherent defect of any television picture, monochrome or color, is the fact that the picture is quantized by its line structure, and that only the transition defects at boundaries between certain colors are more noticeable than the raster lines.

Members of the N.T.S.C. will note that no British translation of the American word "parc," describing color noise due to the overlap of the luminance and chrominance channels is attempted, the forthright term "gnud" having apparently been overlooked.

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Transistor Logic Circuits, by Richard B. Hurley

Published (1961) by John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. 355 pages+5 index pages+xvi pages+1 bibliography page. Illus. 6×9¼. $10.00.

This book is intended mainly as a text for senior students following a computer option in electrical engineering. About half the book is concerned with switching theory and Boolean analysis. The other half presents basic diode and transistor switching circuits. The presentation of these two subjects—the analysis of both logical and circuit characteristics of transistor logic circuits—in one text should be valuable and interesting to students. The book is well written, easy to read, and printed in clear type, with plenty of illustrations.

My main critical comments have to do with certain omissions. These points are generally omitted from most texts. Perhaps that is the academic custom. But any practicing engineer will testify that most of his work is concerned with *tolerances* and *reliability*. The fact that these considerations are left out of most texts (and this is no exception) cheats the student out of adequate preparation for his career. If there were a complete discussion of tolerances and worst-case design for every circuit in the book, it would be too long and unwieldy to teach. However, the worst-case design philosophy should be described and examples of how tolerances are handled in some situations should be analyzed. The reader should be clearly warned in all other areas that he is getting an idealized analysis, and circuit tolerances must be applied in useful designs. The innocent senior should also be warned that most manufacturers tell very little on data sheets about what happens to component values after the components leave his factory. However, the engineering departments of many of these manufacturers do know what happens to their components after long periods of time in various environments. In some cases, the engineer himself must perform tests to determine the proper allowances for aging.

In the area of logic, the author covers binary arithmetic, Boolean algebra, and single and multiple function minimization. Techniques for analyzing sequential logic, including state reduction and state assignment are also described. Most of the widely-used transistor switching circuits are studied. Single and multi-stage diode AND/OR networks, resistor NOR, diode-transistor NOR, DCTL, current switching transistor circuits, etc., flip-flops, single-shot and free-running multivibrators and blocking oscillator circuit analyses are included. Block diagrams of counter registers and several simple accumulators are illustrated and described.

The instructor using this book as a text should be pleased with the list of problems at the end of each chapter and the excellent references. There are a number of tables of equations of transistor voltage and current gain, etc., in various configurations. These will be of tutorial benefit as well as reference for the practicing engineer. In many cases, the engineering approximations are included along with the exact formulas. However, the lack of a list of symbols severely limits the ease of using the book for rapid reference.

Two curious items of terminology will immediately strike the reader. Transistors are called "triodes" throughout the text, but not in the title. The term "curtages" is used to indicate a system of equations, but this reviewer has not been able to find a definition of the word in several modern dictionaries of science and mathematics.

Striking a balance between the advantages and deficiencies discussed above, the book is a very worthwhile contribution, and should be useful in preparing men for careers in fields where switching circuits are designed and used.

Theodore H. Bonn
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Introductory System Analysis: Signals and Systems in Electrical Engineering, by William A. Lynch and John G. Truxal

Published (1961) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 436 pages+5 index pages+xx pages+9 appendix pages. Illus. 6¼×9¼. $7.50.

Written primarily as an introductory electrical engineering text for non-electrical majors in engineering and science, this book features a refreshingly new approach to "service-course" needs. Rather than dwell on dc and ac circuits and machinery, it focuses on linear system analysis techniques that are basic not only to electrical engineering, but also to automotive, aeronautical, marine, mechanical, and civil engineering, as well as applied mathematics and physics. Major emphasis is placed on 1) the model concept, 2) the transfer-function concept, and 3) analog simulation and computation.

The *model* concept is stressed in the first four chapters, wherein analytical models for physical *signals* and physical *systems* are discussed. Chapter 1 introduces the broad concepts of cause (input signal) and effect (output signal) associated with any situation (system). Chapter 2 is devoted to signal models, among which generalized exponentials, sine waves, and the step function are emphasized. In Chapters 3 and 4, models are developed for mechanical and electrical systems, in a manner that accentuates their circuit diagram and circuit equation analogies.

The *transfer-function* concept, a basic approach to the solution of dynamic problems, is developed in Chapters 5 and 6. Chapter 5 concentrates on the response of simple electric circuits, and, among other things, demonstrates that either signals or systems (or both) may be represented in terms of a transfer function. Chapter 6 shows how the transfer function may often be used to interpret system properties without the nuisance of writing differential equations and obtaining complete solutions. The "poles and zeros" concept, sinusoidal frequency response, complex frequency plane, resonance, and other related subjects are also treated.

Analog simulation and computation, an absolute necessity in the solution of very complex systems, is discussed in Chapter 7. Not only are the types, symbolisms and basic schemes of analog simulation examined, but also the technique of signal-flow graphs is developed.

The eighth and last chapter of the book pulls together many of the techniques previously discussed, by way of a single comprehensive example. This is only one of many excellent well-chosen examples that appear throughout the text. Also included are a generous supply of problems, as well as four appendices concerning node equations, the cathode-ray tube, simultaneous algebraic equations, and partial fraction expansions.

In the reviewer's opinion, this book has a somewhat higher level of sophistication than is optimum for undergraduate non-electricals. However, the authors have presented the material with sufficient clarity to allow a good instructor to accomplish well "service-course" introductory needs by appropriate selection and emphasis. Use of this text (for about the first third of the time) plus a more-conventional "service-course" text (for the remainder of the time) should yield a well-balanced program. The level of the book is probably near-optimum for an introductory course for undergraduate electricals or graduate non-electricals, as well as for an introductory reference for practicing engineers.

The reviewer thoroughly enjoyed reading this surprisingly error-free and stimulating introduction to system analysis.

H. M. Straube
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Physics of the Aurora and Airglow, by Joseph W. Chamberlain

Published (1961) by Academic Press, Inc., 111 Fifth Ave., New York 3, N. Y. 566 pages+34 index pages+xviii pages+78 bibliography pages+26 appendix pages. Illus. 6¼×9¼. $16.50.

Dr. Chamberlain published his doctorate dissertation in 1953 on the astronomical subject, "Collisional Excitation of Hydrogen in a Gaseous Nebula." A "tour of duty" at the Air Force Cambridge Research Center in Massachusetts brought him to a professional interest in auroral problems and then to a position on the astronomical staff of the Yerkes Observatory. Here he came under the influence of A. Meinel, at a time when the latter was energetically making outstanding contributions to our knowledge of auroral and airglow phenomena.

Now, only eight years after the publication of Dr. Chamberlain's dissertation, we are reviewing a scholarly book which fills a real need both for researchers and for teachers. As the author states in his introduction, the physical problems of the aurora and airglow cover a diversified field in an age of specialization, so that it is possible for a specialist working on a particular aspect of the problem to be out of touch with other pertinent researches. Chamberlain's book, with its excellent bibliography (more than 1600 references), author index and subject index, facilitates ready reference to a variety of topics.

The author is a theorist, but a rare one, with a discriminating understanding of observational and experimental results. Thus, his theoretical discussions are tempered and disciplined by "facts." This is especially noted by his attention to the optical spectrum of the aurora and airglow as a basis of their understanding. For example, the book contains a complete listing of the wavelengths and intensities of the 45 rotation-vibration bands of hydroxyl (OH), each band consisting of more than 25 individual lines.

In the past, most books on aurora and airglow have made only passing reference to the radio properties of aurora. Since the optical emissions of the aurora are due to the atomic and molecular constituents of the upper atmosphere, while the radio effects are due to the free electrons, the two frequency bands give complementary information. Dr. Chamberlain has performed a valuable service by including discussion of such radio effects as radar echoes from auroral ionization (a complete chapter is devoted to this topic), radio star scintillations and radio-wave absorption during aurora, and auroral sporadic E ionization and the emission of ionospheric and exospheric radio noise during aurora.

The book is divided into the following natural divisions: basic physics (3 chapters), the aurora (5 chapters), and the night, twilight and day airglow (5 chapters). An interesting feature of the book is the inclusion of problems at the end of each chapter, some of which are challenging research projects that will test the ability not only of students, but also of teachers. As a textbook it could be used for a graduate course extended over two semesters. As a reference book it will doubtless be of great value in the library of all active workers in upper atmosphere physics.

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Frequency Modulation Theory: Application to Microwave Links, by Jacques Fagot and Philippe Magne

Published (1961) by Pergamon Press, Inc., 122 E. 55 St., New York 22, N. Y. 480 pages+3 index pages +xi pages+6 appendix pages+references by chapter. Illus. 6¼×9¼. $15.00.

This is a reference book covering principally the design and testing of UHF and microwave line-of-sight FM radio links that are built to carry many simultaneous telephone conversations arranged in frequency division multiplex, or to transmit television video signals. Such a comprehensive coverage of the subject has not previously been available to the engineering fraternity. For this reason, the authors are to be congratulated for completing their ambitious task and excused for a number of mistakes, omissions, and misleading statements and analyses.

Experimentalists need not be alarmed by the term "theory" used in the title. Many of the results are expressed with mathematical symbols, but all of the mathematics could be understood by a college sophomore. Where lengthy or more sophisticated analysis is required, the authors are careful to state each problem, explain its significance, describe the general method of analysis, quote a result from the literature, and interpret that result. The book abounds with numerical calculations, graphs and circuit designs exemplifying the main results.

The key principle around which the book is organized is a familiar theme to telephone engineers: in practice, a link will employ many sections in tandem. In the FM links considered by the book, the noise and distortion introduced in the various sections combine to reduce the quality of the link. The quality of each link section must therefore be superb if the whole link is to be satisfactory. Following this reasoning, a major aim of the book is to discover, analyze, and, if possible, show means for eliminating the various sources of noise and distortion that emerge from a multi-section FM link.

The book is organized into five chapters. In Chapter I, on propagation, free-space loss between directive antennas, atmospheric absorption and refraction, and fading are covered. There is a good discussion of the concept of Fresnel zones; one section deals with knife-edge diffraction. A brief treatment of tropospheric propagation over the horizon is included. Chapter II introduces frequency modulation. Power series are used to describe the distortion caused by slightly nonlinear baseband amplifiers and by the curvature of the phase shift vs frequency characteristics of IF and RF systems. The baseband noise resulting from operation of a conventional FM receiver near, but above, threshold is calculated. Chapter III contains an unusual approximate analysis of the distortion caused by two-path transmission in the atmosphere and through a feeder that is mismatched at both ends.

Chapter IV applies the results of the first three chapters to television and multichannel telephone links. The statistics of a multi-channel FDM telephone signal are introduced. The standards recommended by the CCIR and CCIF for telephone trunks and the principal television standards are discussed, and the requirements placed by these standards on link subsystems are considered.

Slightly more than half of the book is occupied by the discussion of equipment in Chapter V. Topics covered include reflex-klystron and reactance-tube frequency modulators, phase modulators, limiters, discriminators, noise figure, IF filtering and delay equalization, video amplifiers, frequency stabilization, microwave mixers and filters, antennas, ferrite isolators, and methods of measurement. With the exception of the ferrite isolator and the semiconductor diode, solid-state devices are not mentioned. The ratio detector and "threshold-extension" demodulators are not discussed.

The French origin of the book and its publication in English (not American) provide a pleasant international flavor. Almost half of the book's numerous references are in English. The careful organization of chapters into sections makes the table of contents an easier point of entry for reference than the index.

J. GRANLUND
ITT Federal Laboratories
Nutley, N. J.

RECENT BOOKS

Allen, W. D., *Neutron Detection*. Philosophical Library, Inc. 15 E. 40 St., New York 16, N. Y. $10.00.

Birks, J. B. and Schulman, J. H., *Progress in Dielectrics, Vol. I*, John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. $11.00. Aims to provide a common meeting point for all interested in dielectrics, to coordinate current knowledge on dielectric phenomena, materials and technology, and to review recent progress.

Briggs, A., *The Birth of Broadcasting, Vol. I in The History of Broadcasting in the United Kingdom*. Oxford University Press, 417 Fifth Ave., New York 16, N. Y. $10.00. An untechnical account of the persons, committees, boards and shifting policies over a limited period in the early days of broadcasting, with a brief portion relating to broadcasting in countries other than England.

Carroll, J. M., Ed., *Design Manual for Transistor Circuits*. McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. $9.50. Composed solely of articles which have appeared in "Electronics Magazine."

Heumann, G. W., *Magnetic Control of Industrial Motors, Part 3: Direct Current Motor Controllers*. John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. $9.00. In what is primarily an application book, controllers for industrial type ac and dc motors are carefully analyzed and each type of motor is granted full treatment in conjunction with its associated controllers.

Klein, L., Ed., *Dispersion Relations and the Abstract Approach to Field Theory*. Gordon and Breach Publishers, Inc., 150 Fifth Ave., New York, N. Y. $4.95. An introduction to the axiomatic approach to field theory and its application to proving dispersion relations.

Kloeffler, Royce G., *Industrial Electronics and Control, 2nd Edition*. John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. $10.00.

Menzel, Donald H., Ed., *The Radio Noise Spectrum*. Harvard University Press, Cambridge, Mass. $12.50.

Newell, A., et al., Eds., *Information Processing Language-V Manual*. Prentice-Hall, Inc., Englewood Cliffs, N. J. $6.00. A manual for the computer language Information Processing Language-V. Section I gives an introduction and simplified account of the language. Section II sets out the complete rules for coding in IPL-V.

Osborn, J. A., Chairman, Publications Committee, *Proceedings of the Sixth Conference on Magnetism and Magnetic Materials*. McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. $10.00. Invited and contributed papers in the field of magnetism and magnetic materials presented at the annual Magnetism Conference in November, 1960, held under the joint sponsorship of the AIEE and the American Institute of Physics.

Television Field Trials of 405-Line and 625-Line Systems in the UHF and VHF Bands —1957–1958. The British Broadcasting Corp., 35 Marylebone High St., London, W. 1., England. 20s.

Von Koch, H. and Ljungberg, G., Eds., *Instruments and Measurements: Chemical Analysis, Electric Quantities, Nucleonics and Process Control*. Academic Press, Inc., 111 Fifth Ave., New York 3, N. Y. $16.00. Volume I of the *Proceedings* of the Fifth International Instruments and Measurements Conference, Sept. 13–16, 1960, in Stockholm, Sweden.

Wellard, Charles L., *Resistance and Resistors*. McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. $8.50. Groundwork in the basic aspects of resistance, and descriptions of each type of resistor, the materials used in its manufacture, and the characteristics associated with each type.

Scanning the Transactions

Why does stereo sound better than monophonic reproduction? It has been assumed that stereo provides greater enjoyment solely because of the spatial illusion it presents. It is certainly true that the directional effects of stereo give added reality, especially to those performances in which directional information is an important characteristic, such as in operas, chorals and plays. Orchestral music, however, is much less dependent on spatial effects. In fact, if the directional effects are too pronounced, they become distracting. And yet lovers of orchestral music prefer stereo over monophonic hi-fi, even when the dynamic range of stereo is inferior to hi-fi. Recent listening tests have shown that, entirely apart from the purely directional effects, stereo produces a fuller reproduction of the music itself as a consequence of binaural hearing. This is because a tone picked up exclusively by one ear does not cancel, or modulate, or otherwise interfere with another tone picked up exclusively by the other ear. In monophonic reproduction, identical notes played, for example, by violins and by woodwinds can, and frequently do, partially cancel each other because at times they reach the listener in opposite phase. With stereo, on the other hand, the violins and woodwinds do not cancel each other if the violins are heard by one ear and the woodwinds by the other. Studio reverberation is preserved more fully for the same reason. Moreover, a single note produces in some listeners a different subjective pitch in one ear than in the other when heard with different intensities, further helping to segregate identical notes coming from the left and right. Finally, two notes which differ in frequency can intermodulate to produce a discordant beatnote in the monophonic case, a condition which is largely avoided in the stereo case if one note is heard predominantly by one ear and the other note predominantly by the other. Thus, stereo permits the two ears to supply two nearly independent, non-interfering messages to the brain. The result is that the listener literally hears more music with stereo. (C. J. Hirsch, "The non-directional aspects of stereo," IRE TRANS. ON BROADCAST AND TELEVISION RECEIVERS, November, 1961.)

Theoretical versus practical. The Constitution of the IRE defines the aims of the society as including the advancement of the theory and practice of electronics and allied fields. The coupling of the words *theory* and *practice* in this fashion is a familiar and frequently encountered usage. The phrase implies that theory and practice complement one another; further, that they are in contradistinction to one another. This may explain why all too many persons, subconsciously at least, think of another familiar pair, *theoretical* and *practical*, as likewise being separate from one another, to the point of being in contrast to one another. From here it is but one short step to the notion that that which is theoretical is not practical. It is both interesting and illuminating to note what a famous applied scientist, Rankine, had to say about this matter nearly a century ago:

"In the history of mechanical art two modes of progress may be distinguished—the empirical and the scientific. Not the practical and theoretic, for that distinction is fallacious: all real progress in mechanical art, whether theoretical or not, must be practical. The true distinction is this: that the empirical mode of progress is purely and simply practical; the scientific mode of progress is at once practical and theoretic." (W. J. Miller, "What can the technical writer of the past teach the technical writer of today?", IRE TRANS. ON ENGINEERING WRITING AND SPEECH, December, 1961.)

Techniques for storing information have a history that is probably as old as man himself, going back beyond the first use of stone tablets and papyrus to a distant age when inscriptions were made on the walls of caves. The introduction of printing from movable type in the 15th century had a revolutionary impact, not only on methods of storing information, but also on the character and course of civilization. In recent times, great strides have been made in compressing storage space by means of photographic and magnetic recording techniques. It is probable that these systems in greatly improved ultra-high-density form will be as important to the world of tomorrow as the printing press is to the world of today. At present, storage densities of 10^5 to 10^6 bits per cubic centimeter can be obtained with digital magnetic tapes, stereo tapes and LP disk recording, while videotape, movie film and microfilm provide densities of 10^7 to 10^8 bits per cc. Magnetic core and drum memories have a considerably lower storage density, while special types of spectrographic film give densities over 10^{10} bits per cc. What are the ultimate limits of recording density? Theoretically, the ultimate limit would be set by the granular nature of matter. Most materials contain roughly 10^{23} atoms per cc. If the electron spin in each atom could be set in a prearranged pattern of plus or minus magnetic moments, one might say that the ultimate limit would therefore be 10^{23} bits per cc. Turning to biological systems, the genetic information which is contained in coded form in chromosomes yields an information storage density of about 10^{19} bits per cc. The human nervous system, including the brain, is considerably lower: about 10^{11} bits per cc. It is interesting to find that this figure may be surpassed by man-made systems of the future. Present projections indicate that densities of 10^{12} to 10^{13} bits per cc will probably be achieved with future photographic and magnetic recording systems. As a final observation, now that information density is regarded as such an important attribute, it may be well to remember that the next time someone calls you "dense," he may be paying you a compliment. (M. Camras, "Information storage density of magnetic recording and other systems," IRE TRANS. ON AUDIO, September–October, 1961.)

Errata and corrigenda are the bane of an author's or editor's existence. When errors occur in a technical work, it is of course important that a list of the errors with corrections be published. However, to those who perpetrate the errors—author, editor or printer—the publication of such lists is also a public admission that they goofed. It perhaps will be of some small comfort to these tortured souls to know that the problem of corrections has been with us for at least 2000 years. For evidence they need only refer to the cover of the December issue of IRE TRANSACTIONS ON ENGINEERING WRITING AND SPEECH, which carries a photograph of part of the Dead Sea Scrolls, estimated to have been prepared about 100 B.C. They will find that within two short columns, the "author" scribe had to make five corrections after the initial "printing" and that during the following 125 years various "proofreader" scribes found four more major errors, including the omission of entire lines of text. Other errors were never corrected. If occasional errors still occur today, at least the modern author does not have to be his own editor and printer.

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 79 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
Aerospace and Navigational Electronics	ANE-8, No. 4	$2.25	$3.25	$4.50
Audio	AU-9, No. 5	2.25	3.25	4.50
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Product Engineering and Production	PEP-5, No. 3	2.25	3.25	4.50

Aerospace and Navigational Electronics

VOL. ANE-8, NO. 4, DECEMBER, 1961

The Editor Reports—(p. 120)

New Chairman and Vice Chairman of PGANE—(p. 121)

Introduction to Automatic Altitude Reporting for Air Traffic Control—William T. Carnes (p. 122)

The Application of Altitude Telemetering to the ICAO Secondary Surveillance Radar System—Duncan G. Terrington (p. 125)

The operational requirements for telemetering altitude are listed, together with general aspects of the system which can affect implementation.

The special problems which arise in the application of altitude information to the ICAO SSR system are discussed, and it is shown that it is essential to incorporate certain design features in the communications channel and in the coding system to minimize garbling.

The experimental United Kingdom system of altitude telemetering is briefly described, together with proposals for expanding the encoding scale to cover any future requirement.

Altitude Reporting on the ATCRBS—An Impatient Appraisal—S. B. Poritzky (p. 131)

The airline industry, continuing its support for the development and implementation of the ATC radar beacon system, now strongly endorses the use of automatic altitude reporting. In discussing the pitfalls which have delayed implementation of the basic ATC radar beacon system, this paper calls attention to the need for automatic altitude reporting in the ATC system, and the need to provide it quickly. In describing the need for automatic altitude information for both en route and terminal ATC, the choice of an altitude-interrogation mode, the 100-foot-altitude reporting increment, and the flexible code structure being proposed which is intended to cover all significant altitude strata to provide both coarse-grain and fine-grain information as required, are discussed. The paper also describes both early and recent actions by the airline industry in the field of secondary radar.

ATC Operational Uses for Automatic Altitude Reporting—R. L. Biermann (p. 137)

Surveillance radar provides the air traffic controller with only a two-dimensional picture of traffic in a three-dimensional environment. Recently, the Federal Aviation Agency embarked upon the development of an automatic altitude-reporting capability utilizing radar beacon. Its addition to the system will enable presentation of both specific aircraft identification and altitude in alpha-numeric symbology directly on the controller's radarscope adjacent to related radar targets. Operational requirements for this new tool indicate a future capability of reporting altitude from the surface to 100,000 feet in 100-foot increments.

The three-dimensional capability achieved will minimize the need for pilots to make verbal position reports, and will virtually eliminate reports with regard to altitude alone. Giving the controller this capability will reduce the need for arriving and departing aircraft to level out during descent and climb in order to maintain altitude separation from others on conflicting courses. While the aircraft is en route, altitude change will be much less a problem to the controller, permitting the pilot much greater freedom in selection of optimum cruising altitudes.

Automatically reported altitude data is essential to the sophisticated computerized environments being developed for reducing ATC bottlenecks.

ATCRBS Experimental Altitude Reporting Subsystem and Developments—K. Wise (p. 139)

FAA's Aviation Research and Development Service (ARDS), has recently embarked on an active program of establishing an altitude reporting subsystem for the Air Traffic Control Radar Beacon System (ATCRBS), internationally known as Secondary Surveillance Radar (SSR). This action is in response to a long-standing requirement for the provision of altitude information to the Air Traffic Control System by a means other than flight plan and voice transmission of altitude information from the cockpit. The use of the ATCRBS for altitude reporting utilizes precisely the same techniques used for the identification function on Mode A. The major difference is that while codes are manually selected for identification, they would be automatically selected for altitude through an automatic encoding system from a source of barometric altitude. The essential characteristics of the experimental subsystem are detailed, and the Federal Aviation Agency's (FAA) progress to date is described.

Code Configuration for Automatic Altitude Reporting via ATCRBS—Allan Ashley (p. 144)

The development of an unambiguous unit-distance code configuration for automatic altitude reporting via the ATC Radar Beacon System is described. This code provides for compatible altitude reporting in increments of 100, 250, and 500 feet within one common system. The code is developed by using Karnaugh mapping techniques. Some assumptions and requirements are set forth to provide guidance for the determination of the most desirable code. A brief presentation of a developmental code in use in the United Kingdom is also given. It is concluded that the flexibility of the unit-distance code affords a satisfactory compromise which can satisfy the objectives of the various users.

Altitude Processing in the ATCRBS—J. Freibaum (p. 149)

Equipment to process and display beacon-derived altitude information is being developed to enhance the operational capability of the Air Traffic Control Radar Beacon System (ATCRBS). Beacon identity and altitude information assists the air traffic controller to establish positive identification quickly and reliably; radar handoff procedures are completed more efficiently.

The principle altitude functions of the beacon video processor under development are: 1) To provide a digital display, in decimal form, of altitude data received from aircraft equipped with modified transponders which reply to a ground-to-air altitude interrogation. 2) To provide simultaneous altitude zone filtering of the altitude reply codes and tri-coordinate (3-D) radar data. 3) To provide altitude information to associated beacon video tracking units for aircraft tracking.

The fully transistorized processor converts beacon video code trains from serial to parallel form. Leading edge decoding, pulse regeneration, pulse standardization, and destructive readout techniques are used to detect and eliminate spurious or garbled codes and to correctly process closely interleaved replies. Aircraft spaced as close as 344 feet in slant range are correctly read out.

The equipment is designed for operation in either a computerized or noncomputerized environment.

Altitude Sensing for Automatic Altitude Reporting—Jack Andresen (p. 153)

A short resume is given of the means used to measure altitude for reporting purposes, possible sources of error, followed by a description of three of the most commonly used analog-digital converters.

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Suggestions to Authors—(Inside back cover)

Audio

VOL. AU-9, NO. 5, SEPTEMBER–OCTOBER, 1961

The Editor's Corner—Benjamin (Franklin) Miessner (p. 133)

PGA News—(p. 134)

A New Model for Magnetic Recording—B. B. Bauer and C. D. Mee (p. 139)

Magnetic recording is a complicated process which has in the past been portrayed largely in terms of models based upon the magnetic characteristics of the recording medium. The previous models have been incomplete and sometimes incorrect, and they have left unexplained many of the experimental phenomena known to exist in magnetic recording. The proposed new model is not claimed to be precise in every detail. However, it has the virtue of presenting a unified and simplified picture of magnetic recording. The recording process is viewed as an interaction between the idealized properties of magnetic particles and the idealized geometry of the recording field. The latter appears as a series of "bubbles" and "shells" of critical field strengths which grow and collapse throughout the cycles of bias and record current.

Analysis of the Recording of Sine Waves—Irving Stein (p. 146)

Magnetic recording of the sine waves of all wavelengths is analyzed. The analysis shows that the recording process consists of two types of phenomena: 1) a virtual input scanning—the remanent tape magnetization (the recording point) appears at the front gap edge and moves over the gap towards the rear gap edge each half period. Thus, the tape is magnetically "scanned" by the input gap field in the direction opposite the tape motion. 2) A particular type of interference—as the half wavelength of tape passes over the gap, the input field will generally reverse polarity at least once. Thus, the different "scannings" on the same tape section will interfere with each other timewise. The net effect of these processes, even with the initial magnetization curve linear, is to produce a distorted tape magnetization output. The results clearly indicate that for longer wavelengths there is the greatest amount of distortion, but no amplitude attenuation or phase shift. Furthermore, in this wavelength region the recording is done over the entire gap. For smaller wavelengths, the reverse holds true; that is, there is a decreasing amount of distortion and an increasing attenuation and phase shifting. Also, the shorter the wavelength, the closer to the front gap edge is the recording point. It is also found that there is a theoretical lower limit to the wavelength that can be recorded and that at very short wavelengths the shape of the gap trailing field becomes most significant. The fact that increasing oversaturation at the shorter wavelengths decreases instead of increases the net tape magnetization is also explained.

The Mechanism of AC-Biased Magnetic Recording—Donald F. Eldridge (p. 155)

The mechanism of ac-biased magnetic recording is directly related to the physical and magnetic properties of the recording medium. A theory has been developed which describes this relationship, both qualitatively and quantitatively. The action of ac bias on a single-domain particle is examined, and the factors determining its remanent state are analyzed. When the analysis is extended to an aggregate of such particles, it is found that the interaction fields created by the particles themselves are responsible for the biased recording mechanism. The transfer characteristic obtained with ac biasing is derived directly from the distribution of interaction fields in the aggregate material.

Flutter in Magnetic Recording of Data—Charles B. Pear, Jr. (p. 159)

Among the sources of error in magnetic tape recording, one of the more important is the distortion of the reproduced time scale generally described by the term "flutter." Flutter is defined as the deviation in reproduced frequency from the original recorded frequency, which results from nonuniform speed of the recording medium during recording and reproduction.

The present paper provides a qualitative and quantitative discussion of flutter including:

1) Causes—eccentricities, resonances, frictional variations.
2) Effects—FM noise, AM noise, time displacement error.
3) Measurements—methods, units and relations between them.
4) Compensation—methods and limitations.

The object of the paper is to collect the essential characteristics of flutter in a way that will be of value to users of magnetic tape recording equipment.

Design of a Multichannel Magnetic Recording System for Frequency Multiplication—S. Himmelstein (p. 166)

This paper concerns a multichannel magnetic tape recorder/reproducer whose primary function is to provide stable, real-time multiplication (100 times) of all frequencies present at the input. This is accomplished by the use of rotating playback head assembly, located near and "downstream" of a conventional record station. This arrangement yields a short processing delay and, because of the iterative scans, makes possible time division multiplexing of other system components. System requirements and characteristics are presented along with a discussion of the major engineering problems that had to be solved.

Information Storage Density of Magnetic Recording and Other Systems—Marvin Camras (p. 174)

Information storage density in bits per cubic centimeter are compared for various media including magnetic tape, phonograph records, photographic film, the human nervous system, and biological genetics. Magnetic recording is on a par with other man-made systems; but all of these are poorer by many orders of magnitude than genetics. An analogy between contact printing of tapes and genetic replication shows the superiority of the latter. Factors which limit the ultimate density of magnetic recording are discussed: gap-length, straightness, shunt effect, azimuth; tape surface, stability, particle size; head-tape contact, drive stability; amplifier noise, and demagnetization.

Comments on "Enhanced Stereo"—H. K. MacKechnie (p. 180)

Contributors—(p. 181)

Broadcast and Television Receivers

Vol. BTR-7, No. 3, November, 1961

1962–1964 PGBTR Administrative Committee—(p. 2)

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PGBTR Technical Communication—John F. Bell, Zenith (p. 7)

Note on International Convention Meeting—(p. 8)

Performance Characteristic of Television Receivers and Antenna for the UHF-TV Project—E. W. Chapin, F.C.C. (p. 9)

An Airborne UHF Television Transmitter for Educational TV—Frank G. Mullins, Jr., Westinghouse (p. 13)

Complementary Symmetry Sliding Class *A* Audio System—Zbigniew Wiencek and James Chua, Warwick (p. 21)

Specification of Transistor Parameters for Broadcast Receiver Design—H. J. Laurent and P. St. C. Manson, Bendix (p. 29)

For high volume commercial radio production to custom specifications, the specification of meaningful parameters can result in better quality, low cost transistors.

Where it is necessary to put narrow limits on Beta and Gm, a system of subclassification according to these parameters is suggested.

The important parameters for audio circuitry and for RF circuitry are discussed and methods of measuring these parameters are suggested.

The Non-Directional Aspect of Stereo—Charles J. Hirsch, RCA (p. 36)

Design Considerations in the Development of High Quality Stereo Multiplex Receivers—C. D. Knight and O. E. Bechman, Zenith (p. 40)

General requirements of the FM stereo receiver are reviewed with emphasis on high quality stereo performance commensurate with the system capability. In the commercially manufactured receivers described, proper attention has been given to each of these requirements, and the stereo demodulation circuits are integrated within the radio chassis. Two approaches are covered; one, a complete self-contained stereo FM radio, and the other, a novel solution to the conversion problem for existing FM receivers. The latter consists of a "Multiplexer" receiver which is not an "adaptor" and which requires no connection to the monaural receiver. Circuit details and performance data are included.

A Stabilized Ultrasonic Electrostatic Transducer—William L. Hughes, GE (p. 45)

The Noise Figure of Transistor Converters—Roger R. Webster, Texas Instruments (p. 50)

The noise figure of a transistor converter may be expressed in terms of the usual transistor and circuit parameters (source resistance, base resistance, emitter-base diode impedance, current gain, cutoff frequency, and conversion conductance). Theoretical values are in good agreement with measured values.

The converter noise figure is always higher than the corresponding amplifier noise figure, but may approach the amplifier noise figure at high source impedance. The common-emitter converter (signal into base) is superior to the common-base converter (signal into emitter) except at very low source impedance, where the noise figures tend to converge.

The principal difference between amplifier and converter noise figure comes from thermal noise generated in the base resistance, which is converted from many frequencies in addition to the signal frequency.

Stereophonic FM-Receivers and Adaptors—Daniel R. von Recklinghauser, H. H. Scott, Inc. (p. 66)

A Two-Stage DC Coupled Transistor Video Amplifier—P. A. van Berkum and D. J. Herick, Zenith (p. 72)

This report describes a three-transistor dc coupled video amplifier.

The low frequency gain, thermal stability, input impedance, frequency response, distortion, rise time and signal handling capability of the amplifier are discussed and the results of the measurements are given.

Circuit Theory

Vol. CT-8, No. 3, September, 1961

PGCT—Its Growth and Some Problems—J. H. Mulligan, Jr. (p. 182)

Numbers vs Circuit Theory—Philip R. Geffe (p. 183)

Programming the Minimum-Inductance Transformation—K. Yamamoto, K. Fujimoto, and H. Watanabe (p. 184)

This paper presents a method of calculation for the frequency transformation of a low-pass filter into a band-pass filter with a minimum number of inductances. When the chain matrix of a low-pass filter with the order N is factorized into the product of $N/2$ partial matrices, the circuit elements of the required band-pass filter can be found one by one from these partial matrices by means of $N/2$ recurrence equations. The computation formulas and rules, which are described in detail in this paper, are extremely suitable for the use of an automatic digital computer. The flow charts of the programming for these calculation procedures are also illustrated in some detail with a numerical example, which is carried out by the transistorized digital computer Type NEAC-2203.

Design of Lossy Ladder Filters by Digital Computer—C. A. Desoer and S. K. Mitra (p. 192)

Nonavailability of ideal elements is a major drawback in the development of filters having some prescribed characteristic. In this paper, a method for designing lossy filters built with elements having unequal dissipation factors using a high-speed digital computer as the main tool is presented along with the results of a preliminary study of the method. The magnitude function, within the first order, is shown to be a multilinear function of the element values for very small change in the element values. The basis idea is to perturb the element values of the lossy filter with the aim of making the magnitude function of the lossy filter proportional to that of the ideal filter. An error function is defined as the sum of squared differences between the magnitude characteristic of the lossy filter and the same of the ideal filter (multiplied by a suitable constant factor) at some discrete frequencies in the pass-band and the stop band. The error function is then minimized by the steepest descent method of minimization. Results of using the suggested method in designing a lossy Tchebycheff filter of degree 9 are included. It is found that the minimization of the pass-band ripple is associated with a decrease in the stop-band attenuation. The paper also includes details on programming methods.

A Combinatorial-Digital Computation of a Network Parameter—Daniel C. Fielder (p. 202)

In order to avoid numerical iteration in the calculation of a typical return-loss coefficient of an LC ladder structure, combinatorial analysis methods are used to form equations which defer inclusion of the problem data until an advanced stage in the solution.

Three different combinatorial procedures are included. The network parameter calculation serves as a convenient example to demonstrate these combinatorial methods and the applicability of digital computers to combinatorial procedures.

Analysis and Design of Parametric Amplifiers with the Aid of a 709 Computer—B. J. Leon and C. A. Bean (p. 210)

The linear circuit model usually used to characterize the small signal performance of single-diode parametric amplifiers consists of a single sinusoidally varying capacitance imbedded in a fixed parameter network. A computer program for precise analysis of this circuit model is presented. The method of analysis used is one of the exact methods that gives precise solutions regardless of the filtering properties of the passive circuit.

The program is set up so as to be useful in all stages of amplifier design. The admittance of the fixed parameter circuit is included as a separate subroutine. Thus it can be put into the computer as a formula in the initial design phases or as a graph of laboratory measurements to aid in the final alignment of an actual circuit. Besides computing the response at the normal signal and idler frequencies, the program also computes the amplitudes of the voltages at the various other frequencies generated by the mixing action of the time-variant element.

Synthesis of All-Pass Delay Equalizers—Simon Hellerstein (p. 215)

A systematic-synthesis technique has been developed whereby a given delay characteristic is approximated within a given error with a *small* number of all-pass network sections. The method is a generalization of Darlington's using the Tchebycheff polynomial series. As an application, a three-section delay equalizer for color television was designed which had better performance than a four-section equalizer now in use.

A Digital Computer Program for the Design of Phase Correctors—J. V. Fall (p. 223)

A digital computer program for the design of an all-pass function, with n natural modes, to correct the phase of a prescribed transfer function is discussed. The program is based on the expansion of both the desired phase and that of the all-pass function as a Tchebycheff polynomial series, together with the selection of the all-pass natural modes, so that the first n coefficients in the two series agree. If the over-all delay is unimportant, it may be chosen either to increase the number of matched coefficients by one, or to seek a minimum peak error over the approximating bandwidth.

The design process is based on a continued fraction expansion, and a detailed technique for adjusting the over-all delay is discussed. Recurrence relationships are established for expanding the phase of any transfer function as a Tchebycheff polynomial series. The dependence of peak error on approximating bandwidth is discussed, and a routine for the automatic adjustment of the bandwidth to meet a specified peak error is described.

The design accuracy depends primarily on the root extraction technique, and on the loss of significant figures during continued fraction expansion. Adequate results have been obtained with 8 or 10 natural modes. The speed of the program depends on the number of basic iterations of the design process, which varies from one to ten. On a Ferranti Mercury Computer with $n=8$, the basic iteration time is 20 seconds.

The Design of Transistor IF Amplifiers Using an IBM 650 Digital Computer—G. H. Cohen and D. Platnick (p. 237)

This paper describes a method of designing IF amplifiers using a digital computer. G. S. Bahrs has developed a method for designing stable amplifiers using potentially unstable elements. The present paper extends this approach so that it is possible to include bandwidth as a design parameter. Input data consists of sets of two-port parameters measured over the frequency range of interest, a stability factor ρ and the desired bandwidth. Using these values, the computer calculates the values of all elements needed for a complete amplifier. The method is also applicable to multi-element configurations which can be viewed externally as a single unit, *e.g.*, cascode circuits and partially neutralized amplifiers. All calculations are performed by an IBM 650 computer. Several examples are given.

A New Theory of Cascade Synthesis—D. C. Youla (p. 244)

This paper presents a new result generalizing Richards' theorem. It is shown that this result leads to a complete, simple and unified theory of cascade synthesis which yields the types A, B, Brune, C and D sections in a direct and natural manner. The element values of the various sections are obtained in closed form in terms of three or six indexes. Thus the extraction cycle is performed once and for all for the whole class of positive-real functions. Several problems are worked out in detail and a chart is constructed to facilitate the computations. The formulas are easily programmed on a digital computer.

Synthesis of Active RC Networks—J. M. Sipress (p. 260)

Synthesis procedures are presented which establish that one grounded 3-terminal negative-impedance converter, embedded in an unbalanced grounded RC structure, is sufficient to realize

1) any driving-point function,
2) any two of the four short-circuit admittance parameters of a two-port network, and
3) certain sets of n short-circuit admittance parameters of an $(n+1)$-terminal network,

where each of the parameters is specified as the ratio of any two polynomials in the complex-frequency variable, with real coefficients.

Furthermore, the required RC networks can always be made to take the form of grounded ladder-type structures, some of which, in particular cases, reduce to two-terminal admittances.

Polynomial Decomposition in Active Network Synthesis—R. E. Thomas (p. 270)

This paper deals with the use of polynomial decomposition as a basis for the synthesis of active RC networks. The first part briefly reviews the formation and important properties of the Horowitz decomposition. Next, the relationships between the decompositions of two polynomials are presented in the form of three basic theorems. Significant in the development of these theorems is the use made of the theory of positive real functions. The active network function is treated as the even or the odd part of a positive real impedance, and hence the wealth of information available on positive real functions is brought to bear on the problem of active synthesis. The tie between active and passive synthesis is then illustrated by means of a method of active driving point synthesis, which closely parallels the synthesis of lossless terminated passive networks.

A New Design Approach for Feedback Amplifiers—M. S. Ghausi and D. O. Pederson (p. 274)

A new design procedure is presented which is based on the root-locus technique. The realization of the following is possible: a) desired closed-loop response bandwidth (or rise time) and gain level (or amount of desensitivity); b) desired response shapes such as flat magnitude or flat delay; and c) desensitivity of both low-frequency and bandedge responses. The key feature of the root-locus technique is the proper use and location of phantom zeros (transmission zeros of the feedback path). To illustrate the technique, a shunt-shunt feedback configuration is used together with the realization of flat-magnitude-type closed-loop responses. The basic amplifier is a three-stage c-e transistor cascade. Typical design examples with experimental results are included.

Two Extensions of the Darlington Synthesis Procedure—D. Hazony (p. 284)

Two methods are presented which make it possible to synthesize a driving-point impedance function by a Darlington-type synthesis without surplus factors. One method relies on the use of nonreciprocal devices (gyrators) in the synthesized network. The additional constraints due to presence of gyrators are discussed. It is shown that the residue condition for physical realizability now becomes $k_{11}k_{22}-k_{12}^2-\beta^2>0$ (β is the coefficient of the z_{12} term which is of the form $\beta\omega_0/(s^2+\omega_0^2)$). It is shown that two real-part conditions now exist. In addition to the usual Brune real-part condition $Re\ z_{11}\ Re\ z_{22}-(Re(z_{12}+z_{21}))^2/4>0$ for reciprocal elements, there exists another restriction $Re\ 1/y_{11}\ Re\ 1/z_{22}-(Re(z_{12}-z_{21})/2z_{22})^2\geq 0$ on the j-axis (the equal sign is dropped for $Res>0$).

The second method relaxes the cascade net-

work requirement, and from a study of $EvZ(s)$ allows the synthesis of the network by a Darlington-type synthesis. Examples are given of both new methods.

An Approximate Method for Analytically Evaluating the Response of Time-Variable Linear Systems—Elmer G. Gilbert (p. 289)

For all but the most elementary time-variable linear systems, solutions of the response problem must be particular solutions obtained by machine computation. Thus, the insight into system response characteristics, which is gained from analytic procedures, is lost. This paper proposes an approximate representation of the time-variable weighting function $h(t, \tau)$ (system response at time t to a unit impulse at time $t-\tau$) which permits a straight-forward analytic evaluation of response for both deterministic and stochastic inputs. Although the evaluation of the approximate representation may require machine computation, once the approximate representation is obtained, no further machine calculations are necessary.

The approximation has the form $h(t, \tau) = \sum_{i=1}^{N} a_i(t)\phi_i(\tau)$ where the $\phi_i(\tau)$ are arbitrary approximating functions. Methods are given for evaluating the $a_i(t)$. Several families of orthonormal approximating functions are discussed which simplify both the evaluation and utilization of the approximation. For example, when the $\phi_i(\tau)$ have rational Laplace transforms, the $\alpha_i(t)$ are computed directly from the solutions of certain time-variable linear differential equations. For stationary stochastic inputs, simple series expressions for the output correlation function are derived, and it is shown that propitious choice of the $\phi_i(\tau)$ greatly reduces the number of integrations which must be made in evaluating the series. Two example problems illustrate the feasibility of the proposed procedures.

Selectivity Concentration and Integration Efficiency—W. Gersch and K. S. Miller (p. 296)

The relationship between the maximum achievable peakedness of the selectivity characteristic and the signal detection performance of a spectrum analyzer whose input is a finite duration sinusoid immersed in white noise is considered. The spectrum analyzer is assumed to be the equivalent of a bank of filters whose outputs are measured at a single point in time. Such systems have application to Doppler radars and speech analysis. In the Doppler radar application, the peakedness of the selectivity characteristic is a measure of frequency estimation performance capabilities of the spectrum analyzer system against single targets. The signal detection performance of a system may be characterized by its integration efficiency, which is a relative measure of the system's signal-to-noise ratio using the matched filter as a standard. An expression for integration efficiency in terms of filter impulsive response is developed. The selectivity characteristic is defined as the spectrum analyzer normalized output signal power. An expression for the selectivity characteristic is also developed in terms of the filter impulsive response. The mathematical problem considered is the minimization of the even moments of signal concentration for fixed integration efficiencies. The result obtained by variational techniques is the specification of an impulsive response that minimizes the selectivity concentration for a fixed integration efficiency. The utility of this criterion is that it permits a trade off of accuracy of signal frequency estimation performance vs signal detectability performance. The second moment case is treated completely, and a special case of higher order moments is also examined. Graphical results of the selectivity characteristics for some special cases are also presented and compared with the results of the spectral shaping windows of Blackman and Tukey.

Pulse Networks with Parabolic Distribution of Poles—S. K. Mullick (p. 302)

In this paper an attempt is made to find the pole configuraation for a transfer function whose zeroes are all at infinity and which will yield a transient response showing improvement over both the Butterworth, Thomson or Transitional Butterworth-Thomson ones. Some discussions of the monotonicity conditions of the time response suggest that the poles may conveniently be located on parabolic contours in the left half of the p plane. On investigation of such networks, it is found that they show very little or no overshoot, and small rise time. Also the overshoot decreases with increase of the order of the network.

A Simple Method for the Analysis of Demodulator Compensating Networks—E. V. Bohn (p. 306)

A simple approximate method for the analysis of demodulator compensating networks is developed which is valid for a large carrier to signal frequency ratio. The method is based on determining suitable modifications to a dc-prototype network. An equivalent network is derived by noting that the poles of the voltage transfer function of both networks are identical, and that the capacitor voltages of the networks differ by a constant factor. The equivalent circuits of a variety of lead, lag and lead-lag networks are derived.

Conditions for the Realizability of a Conductance Matrix—G. Biorci and P. P. Civalleri (p. 312)

It is shown that it is possible to determine whether a given conductance matrix of order n is realizable by a network with $n+1$ nodes without going through the actual realization of the tree. This realizability criterion requires only a proper arrangement of the rows and columns of the sign matrix, corresponding to the assigned conductance matrix.

On the Realization of an nth-Order G Matrix—E. A. Guillemin (p. 318)

Synthesis of a single-element-kind network from a given short-circuit parameter matrix of order n, leading to a topological configuration involving $n+1$ nodes, is now known, having been given by the author and several others working on this problem almost simultaneously. This form of realization is, however, not completely general, and therefore it is possible for a given matrix to be physically realizable even though the $(n+1)$-node realization procedure fails. The present paper discusses this more general problem and leads to some useful results which do not, however, represent a complete solution.

Topological Considerations in the Realization of Resistive n-Port Networks—I. Cederbaum (p. 324)

The topological implications of irreducibility of the admittance or impedance matrix of an n-port network are studied. Special attention is given to the cases when in $(n-1)$ rows of such a matrix of order n the main diagonal elements are equal to the absolute values of some off-diagonal elements. It is shown that the conditions of realizability of a Y or Z matrix of this type reduce to the known conditions of realizability by means of a network with $(n+1)$ vertices or exactly n independent circuits. Some examples show realizable Z matrices which cannot be realized as Y matrices and vice versa. Other examples give nonrealizable, paramount Y and Z matrices showing that paramountcy is not a sufficient condition for realizability.

Some Optimum Properties of n Ports—Harry Gruenberg (p. 329)

It is known that it is impossible to construct certain types of n ports for which too many designs constraints are required. For instance, a lossless, reciprocal three port cannot be completely matched.

The paper is mainly concerned with n ports that provide an equal power split. Scattering matrices are derived which provide for maximum power transfer for the two limiting cases of zero absorption loss and zero reflection at the input ports. It is shown that the first of these cases corresponds to an absolute maximum power transfer and that no other division of the over-all loss into reflection and absorption loss will yield better results. For matched inputs, it is shown that the optimum transmission can always be improved by going from a reciprocal to a nonreciprocal n port. While no matched, reciprocal three ports (providing an equal power split from all ports) can be built with less than 50 per cent loss, an equivalent nonreciprocal device can have losses as low as $33\frac{1}{3}$ per cent. Similar results are given for n ports in general.

Stability of Logical Networks and its Application to Improvement of Reliability—K. K. Maitra (p. 335)

McCulloch showed that certain redundancy networks composed of neuron-like elements exhibit the property of "logical stability" even though the individual neurons may undergo transformation of states caused by shift in neural threshold. It was speculated that the stability of neural nets contributes to the improved behaviorial reliability of human beings, a problem that was originally raised by Von Neumann. This paper extends the concept of fallible neurons to conventional logic modules as found in technological systems. It is shown that in a single logic module, transformation of states may occur due to malfunction of the internal components and/or drift in the thresholds of the signal and control inputs. The existence of multiple states in a single module lends itself to the possibility of constructing networks with redundant modules which are functionally more stable than the individual modules. Finally, it is shown that suitably chosen, logically stable structures may be used to replace the individual modules in order to perform logical operations with greater reliability.

Approximation Theory for Filter-Networks—Hitoshi Watanabe (p. 341)

This paper presents a general design theory for filter-networks constructed on the basis of theory of the Abelian Integral. An ideal transmission function is defined to be an Abelian Integral $w(\lambda)$ with the following properties: 1) $Re\ w(\lambda) = u(\lambda) = A_k$ in given regions B_k, 2) $\{u(\lambda) - \log|\lambda + a_i|\}$ is regular at any given point a_i, and 3) otherwise $u(\lambda)$ is a harmonic function. The application of appropriate analytic transformation techniques to $w(\lambda)$ leads to a generalized characteristic function $\phi(\lambda)$. All kinds of realizable transfer functions can be derived by the use of linear transformations of $\{\phi(\lambda)\}^2$ with respect to λ^2.

Further, this treatise gives the design methods for filter-networks having one or two pass bands, in which they exhibit the Tchebycheff performance. Besides the method is also established to synthesize a reactance band-pass network with the order N using only $[(N-1)/2]$ coils. Three design examples, such that the band-pass filter comprising only 6 coils for $N=13$ and 4 coils for $N=10$, and the double-pass-band filters with $N=8$ and 4 coils, are described in quite detail. All of the examples have the Tchebycheff performance in their pass bands.

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Engineering Writing and Speech

VOL. EWS-4, NO. 3, DECEMBER, 1961

The Cover—(p. 67)

The Objectives of PGEWS—(p. 68)

What Can the Technical Writer of the Past Teach the Technical Writer of Today?—Walter J. Miller (p. 69)

When in despair over the problems of presenting technical information, today's engineer, scientist, and technical writer can gain fresh inspiration by reading the classics of engineering literature. Major writers, including ancient and medieval authors as well as outstanding moderns like Ammann and Raymond, can teach us a lot about variations in form and approach and about appropriate use of rhythm and metaphor. Further, our great technical writers make us reconsider our contemporary notion of "objectivity" and its values.

Creative Aspects of Engineering Writing—Herbert B. Michaelson (p. 77)

Writing an engineering paper is shown to be an intrinsic part of the engineering work, rather than a mere recording of results. The process of writing a paper for publication in an engineering journal has some rather interesting aspects of creativity.

This paper shows how any careful attempt to record and analyze the results of an engineering project will actually contribute to the engineer's understanding of his own work. Writing a paper not only affords a fresh insight to the author, but may even inspire a more thorough study and a new approach to laboratory work before the manuscript is finished.

In addition, this paper exposes three popular fallacies about the creative aspects of writing.

The Writing of Abstracts—Christian K. Arnold (p. 80)

The abstract is not only the most important section of your report or paper, but also the most difficult to write. Effective abstracts 1) must contain enough specific information to satisfy the needs of a research worker looking for sources of information and of an administrator looking for a progress or status definition; 2) must be a complete, self-sufficient unit; 3) must be made as short as possible without violating accuracy or eliminating essential information; 4) must be written in fluent, easily understood language; 5) must be consistent in tone and emphasis with the parent report or paper; and 6) should make the widest possible use of numerals and standard, generally recognized abbreviations.

The Use of Semantic Structure in Information Systems—J. D. Sable (p. 83)

An approach to the problem of obtaining a semantically structured vocabulary for use in information systems is presented. It is felt that this approach is valid for a system whose users are a relatively homogeneous group in a well-defined problem area. The semantic structure of the technical vocabulary of such a group is defined and is presented as a useful representation of a relatively stable relationship between the words appearing in queries about information in the system and the words in the information-carrying documents or records themselves. Procedures are given for the discovery of semantic structure and its use in an index to information and for a formal language for describing information.

Testing the Readability of Engineering Writing—Stewart Doctor (p. 91)

Preparing a Bibliography—Brian F. Eden—(p. 97)

Preparing concise, acceptable bibliographies is one of the unpleasant details of report writing facing the engineer, writer, and editor. Here, for use as a guide, is a comprehensive summary of the essentials of their preparation.

Correction to "Apply Quality Control of Engineering Writing"—Ronald S. Blicq (p. 100)

Abstracts of Recent Articles on Engineering Writing and Speech—(p. 101)

Book Reviews—(p. 102)

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Product Engineering and Production

Vol. PEP-5, No. 3, November, 1961

Message from the Editor—David Ehrenpreis (p. 1)

Grateful Acknowledgment—(p. 2)

Organizing and Planning for Production—Loren D. Butler (p. 3)

Value-Engineering—An Engineer's Heritage and Discipline—A. D. Zappacosta (p. 8)

Encapsulation Techniques for Functional Blocks in Molecular Electronics—J. M. Steiner (p. 14)

Metal Diffusion Techniques and Glass Sealing Problems in the Electron Tube Industry—R. Wachtell and M. W. Leighton (p. 20)

This paper deals with early techniques of chrome diffusion and the more recent methods of chromizing of metals. Historical applications of the process as applied to certain glass sealing metals, its effect on the coefficient of thermal expansion and the interface between glass and metal, as well as proper seal geometry, are discussed. Precleaning, oxide formation, glass application, seal types, and post-removal of the oxide are explained.

Machine Preparation of Printed-Wiring Masks—Elis A. Guditz (p. 22)

A major source of errors in making photographic masks for precision etched wiring is the distortion introduced while photographically reducing master drawings to working size. Such errors can be eliminated by direct exposure of high-contrast photographic film with a fine pencil-like beam of light.

A tape-controlled, semi-automatic machine based on this principle has produced 8×10 in wiring patterns with line-position accuracy of ±mil from the pattern edge. The exposures are limited to straight, parallel, continuous or discontinuous lines 2 mils, or greater, in width. Masks requiring perpendicular lines are contact-printed from a composite of parallel-line patterns aligned under a microscope.

The unit consists of a reconditioned milling machine, modified to provide a controlled light source above the motor-driven work table, and the electronic equipment necessary for control. The light is program-controlled from a punched paper tape whose drive mechanism is directly coupled to the work table. A manual punch is used to prepare the tapes from a dimensioned sketch of the wiring.

The machine, under continual improvement for the past four years, has produced many precision masks for experimental printed wiring.

A source of error in making photographic masks for precision etched wiring is distortion-introduced while photographically reducing a master drawing to working size. Such errors can be eliminated by using a pencil-like beam of light to expose directly a photographic film. The tape-controlled, semi-automatic machine to be described has produced wiring patterns 8 in ×10 in in size with line-position accuracy of ±0.001 in from a reference edge. It can expose straight, parallel, continuous or discontinuous lines as narrow as 0.002 in. Utilizing timed exposures, it can produce patterns of round dots, or lands. The wiring patterns are exposed on high-contrast photographic film under darkroom conditions. Note the class limitation, i.e., patterns consisting of only parallel and perpendicular lines. This is because the machine is power-driven in only one direction. Patterns which contain perpendicular lines must be contact printed from composites of patterns of patterns of parallel lines.

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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The 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.

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.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears each June and December at the end of the Abstracts and References section.

The Index to the Abstracts and References published in the PROC. IRE from February, 1960 through January, 1961 is published by the PROC. IRE, May, 1961, Part II. It is also published by *Electronic Technology* and appears in the March, 1961, issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.14-8:535.215:546.48'221 1

Ultrasonic Amplification in CdS—A. R. Hutson, J. H. McFee, and D. L. White. (*Phys. Rev. Lett.*, vol. 7, pp. 237–239; September 15, 1961.) Application of a dc electric field in the direction of wave propagation in photoconductive CdS gave amplification of ultrasonic waves. Net acoustic gain was demonstrated even in the absence of an input signal. Results show gains of 18 db at 15 Mc, and 38 db at 45 Mc in a 7-mm length of CdS.

534.232:621.318.134 2

Effect of a Longitudinal Static Pressure upon a Magnetostrictive Ferrite—G. L. Wilson. (*J. Acoust. Soc. Am.*, vol. 33, pp. 1127–1130; August, 1961.) Transducers for underwater sound propagation may be exposed to considerable hydrostatic pressure. For a pressure in the element of 4600 lb/in², the effective coupling coefficient dropped from 0.21 to 0.05 and the clamped inductance decreased by 40 per cent.

534.283-8:537.312.62 3

Ultrasonic Attenuation in Tin Single Crystals at Low Temperatures—K. Kamigaki. (*J. Phys. Soc. Japan*, vol. 16, pp. 1141–1144; June, 1961.) The ultrasonic attenuation coefficient and velocity in normal and superconducting crystals were measured in the range 1.5–4.2°K. Anisotropy of attenuation for long wavelengths occurs for different crystallographic directions and the ratio of longitudinal to shear-wave attenuation was larger than predicted from the ree-electron model.

534.283-8:538.221 4

Ultrasonic Attenuation in Magnetite at Low Temperatures—K. Kamigaki. (*J. Phys. Soc. Japan*, vol. 16, pp. 1170–1174; June, 1961.) Coefficients were measured over the temperature range 4.2–300°K and two peaks were observed below the transformation temperature of ionic ordering.

534.286-8:537.228.1 5

Absorption of Sound in Insulators—Woodruff and Ehrenreich. (See 179.)

534.78:681.841.3 6

The Phonetograph: Present and Future—J. Dreyfus-Graf. (*Tech. Mitt. PTT*, vol. 39, pp. 160–172; May, 1961. In French.) Details are given of the facilities provided by Mark III of the equipment described in 3021 of 1957. New features planned for the phonetograph Mark IV are also mentioned.

534.88:621.396.677.3 7

Optimum Envelope Resolution in an Array Correlator—M. J. Jacobson. (*J. Acoust. Soc. Am.*, vol. 33, pp. 1055–1060; August, 1961.) A system with two identical collinear arrays of uniformly spaced elements is considered. For a sinusoidal input an envelope mainlobe width reduction of 30 per cent is obtained over a uniform amplitude system by introducing complex amplitude factors after each element, but there is a resultant signal-to-noise degradation.

621.395.614:534.76.001.57 8

Operational Research on Microphone and Studio Techniques in Stereophony—D. E. L. Shorter. (*B.B.C. Engr. Div. Monographs*, pp. 5–23; September, 1961.) Conclusions drawn from operational experiments are discussed with reference to the problems of securing compatibility in transmission and to the design and operating requirements of the special studio equipment necessary.

621.395.616 9

Miniature Condenser Microphone—D. L. Subrahmanyam and K. D. Pavate. (*J. Inst. Telecommun. Engrs., India*, vol. 7, pp. 129–134; May, 1961.) The design of a condenser microphone is explained. It has a frequency response of 40 cps–10 kc within ±2.5 db and a sensitivity of −55 db with respect to 1 v/dyne cm⁻².

621.395.623.7.001.4 (083.74) 10

IRE Recommended Practices on Audio and Electroacoustics: Loudspeaker Measurements, 1961—(PROC. IRE, vol. 49, pp. 1553–1560; October, 1961.) Standard 61 IRE 30.RP1.

621.395.623.75 11

Model Investigations on the Velocity Transformation in Pressure Chambers for Horn-Type Loudspeakers—E. Meyer and E. Schunk. (*Akust. Beih.*, no. 1, pp. 248–253; 1961.) Model tests were carried out with a large flat chamber in which the motion of oil droplets suspended in air was observed.

ANTENNAS AND TRANSMISSION LINES

621.315.212 12

High-Frequency Power Cables—U. Queck. (*Elektrotech. Z.*, vol. 82, pp. 501–507; July 31, 1961.) The construction, properties and installation of coaxial feeder cables are described; the external conductor is a corrugated-aluminum sheath and the cables can be pressurized.

621.372.8:537.226 13

Shielded Dielectric Waveguide—K. W. H. Foulds and P. Sampaio. (*Electronic Tech.*, vol. 38, pp. 393–397; November, 1961.) The electric field distribution measured in air and in the dielectric of the waveguide is shown to decay exponentially outside the dielectric slab. Experimental values of decay and of guide wavelength are in close agreement with theoretical results.

621.372.823+621.372.829 14
The Observed 33 to 90 kMc/s Attenuation of Two-Inch Improved Waveguide—A. P. King and G. D. Mandeville. (*Bell Sys. Tech. J.*, vol. 40, pp. 1323–1330; September, 1961.) Theoretical and experimental results are given for a solid-wall copper waveguide and a helix waveguide. The loss measurements were made with a shuttle-pulse technique using a single backward-wave oscillator.

621.372.823:621.396.41 15
Interference Waves in Waveguide Transmission and Intermodulation Noise in F.M. System for Waveguide Communication—Miyauchi. (See 291.)

621.372.824 16
The Imperfectly Conducting Coaxial Line—T. T. Wu. (*Quart. Appl. Math.*, vol. 19, pp. 1–13, April, 1961.) A theoretical study of em wave propagation along and through a coaxial line with imperfectly conducting walls. The "mode" concept of transmission is shown to have a very limited scope of application.

621.372.83/.85:621.318.134 17
Ferrite Components for U.H.F. and Microwave Systems—J. C. Cacheris and N. G. Sakiotis. (*Electronics*, vol. 34, pp. 37–42; September, 1961.) Recent designs of ferrite isolators, circulators, switches and phase shifters are described.

621.372.852 18
Contacting Piston for a Small-Height Waveguide—F. A. Braun. (*Rev. Sci. Instr.*, vol. 32, p. 992; August, 1961.) A simple contacting piston is described which has operated satisfactorily and with very low loss in a waveguide only 0.01 in high.

621.372.852.32 19
Power Limiting in the 4-kMc/s to 7-kMc/s Frequency Range using Lithium Ferrite—F. C. Rossol. (PROC. IRE, vol. 49, p. 1574; October, 1961.) Evidence indicating the possibility of extending the useful range of a coincidence limiter to about 7 Gc is discussed.

621.372.852.323:538.221 20
A Ferrimagnetic Limiter-Isolator—J. Brown and G. R. Harrison. (PROC. IRE, vol. 49, p. 1575; October, 1961.) A limiter-isolator is described which is based on the subsidiary resonance effect observed at high field strengths in ferrites and garnets. The device can also act as an interference suppressor.

621.396.67:621.397.62 21
Receiving Antenna and Distribution Systems for Operation of Multiple Television Receivers—F. R. Lackey. (*Proc. IRE (Australia)*, vol. 22, pp. 429–441; July, 1961.)

621.396.674.37 22
Calibration of Loop Antennas at V.L.F.—A. G. Jean, H. E. Taggart, and J. R. Wait. (*J. Res. NBS*, vol. 65C, pp. 189–193; July-September, 1961.) Field-strength measurement technique and suitable equipment are described.

621.396.677:621.391.822 23
Antenna and Receiving-System Noise-Temperature Calculation—L. V. Blake. (PROC. IRE, vol. 49, pp. 1568–1569; October, 1961.) Antenna noise temperature in the frequency range 100–10,000 Mc is presented in graphical form, for fixed contributions of cosmic, solar and ground noise. The applications of the results are discussed.

621.396.677.4 24
On the Impedance of Long Wire Suspended over the Ground—J. R. Wait. (PROC. IRE, vol. 49, p. 1576; October, 1961.) An expression is derived for the impedance of an infinitely long wire located at a given height over a homogeneous flat ground.

621.396.677.4 25
Experimental Study of Inverted *L*-, *T*-, and Related Transmission-Line Antennas—S. Prasad and R. W. P. King. (*J. Res. NBS*, vol. 65D, pp. 449–454; September/October, 1961.) Experimental and theoretical data are given. The apparent measured impedance is corrected for end effects and approximate theoretical impedances are calculated for very small heights for which measured values cannot be obtained readily.

621.396.677.831 26
The Computation of Radiation Diagrams and Surface Utilization of Plane Reflectors—G. F. Koch. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 60–68; May, 1961.) Various approximation methods based on Kirchhoff's principle are compared. See also 509 of 1961.

621.396.677.85 27
Experimental Investigations of Dielectric Spherical Lenses and Reflectors—B. Chiron and F. Holvoet-Vermaut. (*Onde élect.*, vol. 41, pp. 481–489; May, 1961.) The construction of a complete lens system whose components can be individually focused is described and methods of measuring its electrical characteristics are given. A comparison is made between a Luneburg dielectric reflector and several metallic reflectors of equivalent size for *X*-band operation.

AUTOMATIC COMPUTERS

681.142 28
Theory of Majority-Decision Elements—S. Muroga, I. Toda, and S. Takasu. (*J. Franklin Inst.*, vol. 271, pp. 376–418; May, 1961.) Various problems concerning a single majority-decision element are considered.

681.142 29
Majority Decision Functions of Six Variables—I. Toda, M. Kondo, and S. Muroga. (*Rev. Elec. Commun. Lab., Japan*, vol. 9, pp. 253–278; May/June, 1961.) See also 28 above.

681.142 30
A Fourier Transform Generator: Parts 1 & 2—G. F. Newell and W. K. E. Geddes. (*Electronic Eng.*, vol. 33, pp. 556–561 and 646–649; September and October, 1961.)

681.142 31
The Address System and Increment Store for a Digital Differential Analyser—P. L. Owen, M. F. Partridge, and T. R. H. Sizer. (*Electronic Eng.*, vol. 33, pp. 656–663; October, 1961.)

681.142 32
Digitizer-to-Tape Punch-Coupling Unit—L. Molyneux, E. E. Schneider, and G. R. Sharp. (*J. Sci. Instr.*, vol. 38, pp. 390–395; October, 1961.) The output in twelve binary digits of a digitizer is briefly stored before operating a high-speed five-track tape punch.

681.142:538.632 33
Study of the Application of the Hall Effect in the Construction of Logic-Circuit Elements—J. Vermot-Gaud (*Ann. Télécommun.*, vol. 16, pp. 133–151; May/June, 1961.)

681.142:539.23 34
The Use of Thin Films for Computer Storage—C. I. Cowan. (*Electronic Eng.*, vol. 33, pp. 642–645; October, 1961.)

681.142:621.318.042.1 35
Diode Steering increases Speed of Magnetic Memories—A. Melmed, R. Shevlin, and W. Orvedahl. (*Electronics*, vol. 34, pp. 68–70; September 15, 1961.) By using the recovery time of junction diodes for half-selection of the same word during the second half of a "read" or "store" operation, a 3-μsec cycle time is achieved in a magnetic storage system with 4096-word capacity.

681.142:621.318.57 36
Design of Logic Circuits using Thin Films and Tunnel Diodes—T. A. Smay and A. V. Pohm. (*Electronics*, vol. 34, pp. 59–61; September 15, 1961.) Switching circuits which have operated successfully are described and designs suitable for a wide range of logic functions are suggested.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.22:621.391.822 37
Resistance as Dissipation into Many Reactive Circuits: Landau Damping and Nyquist's Noise Theorem—O. Buneman. (*J. Appl. Phys.*, vol. 32, p. 1783; September, 1961.) Nonradiative dissipation is not a loss of energy but its dispersal into a large number of degrees of freedom. Resistance can be simulated by purely reactive elements. This leads to an understanding of Landau damping and Nyquist's noise theorem.

621.3.049.7 38
Techniques of Microminiaturization—D. H. Roberts and D. S. Campbell. (*J. Brit. IRE*, vol. 22, pp. 281–292; October, 1961.) The techniques of evaporated film and semiconductor solid-state circuit design are described and their merits discussed.

621.318.57:538.221 39
Metal Transfluxors—R. Deller. (*Z. angew. Phys.*, vol. 13, pp. 180–182; April, 1961.) The choice of materials for, and the design of a three-aperture transfluxor [*e.g.*, 3509 of 1955 (Rajchmann and Lo)] are considered.

621.318.57:621.382.3 40
Transient Response of Grounded-Emitter Switching Circuits—S. C. Chao. (*Semiconductor Prod.*, vol. 4, pp. 27–33; May, 1961.) Analysis of the effects of collector capacitance and load resistance. A ramp input voltage is assumed, which is closer to the practical case than a step function.

621.372.4/.5:681.142 41
The Representation of Branched Circuits on an Analogue Computer—W. Schüssler. (*Arch. elekt. Übertragung*, vol. 15, pp. 215–226; May, 1961.) Extension of earlier work (2111 of 1961) to cover another class of branched network.

621.372.412:537.228.1 42
The Effect of Driving-Electrode Shape on the Electrical Properties on Piezoelectric Crystals—J. A. Lewis. (*Bell Sys. Tech. J.*, vol. 40, pp. 1259–1280; September, 1961.) Formulas are derived for the electrical admittance of a piezoelectric crystal, in terms of its resonant frequencies and static and dynamic capacitances. The results are applied to the investigation of the effect of electrode shape on the resonance spectrum and the capacitance ratio of the crystal.

621.372.414:621.372.543.2 43
The Expanded Reactance Series Resonator —'ERSER'—E. O. Willoughby. (PROC. IRE, vol. 49, p. 1583; October, 1961.) A description is given of a circuit element wherein the bandwidth between the infinities of a series resonant circuit can be exchanged for increased steepness of the reactance slope through resonance.

621.372.44 44
Time-Varying Networks: Part 1—L. A. Zadeh. (PROC. IRE, vol. 49, pp. 1488–1503; October, 1961.) An analysis is presented of

some significant developments in the last decade, in the theory of time-varying network. 138 references.

621.372.44 45
Nonsymmetrical Properties of Nonlinear Elements in Low- and High-Impedance Circuits—A. L. Helgesson. (PROC. IRE, vol. 49, pp. 1569–1570; October, 1961.)

621.372.5 46
On Gyrator Networks—K. K. Nair and M. N. S. Swamy. (*J. Inst. Telecommun. Engrs., India*, vol. 7, pp. 111–116; May, 1961.) The attenuation and image impedance characteristics of four-terminal networks using nonreciprocal elements are considered. In networks incorporating gyrators it is possible to obtain finite attenuation in one direction and infinite attenuation in the other.

621.372.5 47
Nonrealizability of the Complex Transformers—R. W. Newcomb. (PROC. IRE, vol. 49, pp. 1585–1586; October, 1961.) A proof is given to show that for a particular case the complex transformer [2225 of 1955 (Carlin)] cannot be realized.

621.372.51 48
High-Frequency Broad-Band Impedance Matching—H. R. B. Seetharam and S. Datta. (*J. Inst. Telecommun. Engrs., India*, vol. 7, pp. 135–144; May, 1961.) The design, initial adjustments and performance of a confluent band-pass filter are discussed and specific examples covering frequency ranges of 1:1.5 and 1:2.3 are given.

621.372.54:621.391 49
An Adaptive Communications Filter—C. S. Weaver. (PROC. IRE, vol. 49, pp. 1587–1588; October, 1961.) A filter is described which is optimum for a completely random communication signal.

621.373:621.391.822:681.142 50
Generating Random Noise with Radioactive Sources—J. B. Manelis. (*Electronics*, vol. 34, pp. 66–69; September 8, 1961.) The system comprises a radioactive isotope mixed with a light-emitting phosphor, and a Type-931A photomultiplier tube. It was designed for application with a repetitive analog computer and provides both a random telegraph wave with a mean count rate of 80 kc and Gaussian noise with a power spectrum flat from dc to 20 kc.

621.373.44 51
The Generation of Variable Square Pulses in the Nanosecond Range—F. Maisenhälder, H. D. Purps, and E. Pfender. (*Arch. elekt. Übertragung*, vol. 15, pp. 253–256; May, 1961.) A generator is described which is capable of producing pulses of duration variable between 20 nsec and 1.4 μsec with pulse rise and decay (between 10 per cent and 90 per cent of pulse height) within 10 nsec. Correction, *Arch. elekt. Übertragung*, vol. 15, p. 334; July, 1961.

621.374 52
Pulse Analysis—R. W. Harris. (*Proc. IRE (Australia)*, vol. 22, pp. 441–448; July, 1961.) "The mathematical aspects of the determination of the response of systems by pulse techniques is discussed with particular reference to many of the practical problems involved in applying the techniques."

621.374.33:621.396.96 53
Radar Boxcar Circuits using Nuvistors, Transistors and Zeners—P. E. Harris. (*Electronics*, vol. 34, pp. 66–67; September 15, 1961.) Two versions of a waveform sampling device with Zener-diode biasing are described which give a 60-db SNR at room temperature.

621.375.1:621.372.543.2 54
Amplifiers with Band-Filter Coupling—J. Harmans. (*Elektron. Rundschau*, vol. 15, pp. 203–206; May, 1961.) The design of multistage amplifiers with flat or Tchebycheff-type response is considered, design formulas and response curves are given.

621.375.4 55
Transistor Amplifier Stages with Prescribed Gain and Sensitivity—J. H. Mulligan, Jr., and S. S. Shamis. (*Elect. Engrg.*, vol. 80, pp. 443–448; June, 1961.) A design procedure is developed for circuits meeting prescribed gain and sensitivity criteria in terms of anticipated operational variations and available component data.

621.375.43 56
An Engineering Approach to the Design of Transistor Feedback Amplifiers—E. M. Cherry. (*Proc. IRE (Australia)*, vol. 22, pp. 303–320; May, 1961.) A technique is developed for amplifier design which produces very accurate expressions for the transmittance without requiring any quantitative information about the transistors.

621.375.5 57
The Dielectric Harmonic Amplifier (a Controllable Harmonic Generator)—G. Bullinger. (*Arch. Elektrotech.*, vol. 46, pp. 141–148; May 9, 1961.) Circuits incorporating ferroelectric capacitors for producing harmonics of power frequencies are described. Applications as amplifiers or controllable frequency changers are mentioned.

621.375.9:621.372.44 58
Lossy Coupling in Parametric Amplifier—K. Grabowski. (PROC. IRE, vol. 49, p. 1573; October, 1961.) A method of analyzing a two-resonant-circuit parametric amplifier with a lossy coupling branch.

621.375.9:621.372.44 59
Experiments with Nonreciprocal Parametric Devices—A. Korpel and P. Desmares. (PROC. IRE, vol. 49, p. 1582; October, 1961.) A report is given of the behavior of two nonreciprocal parametric devices. One device is a 200-Mc unilateral amplifier, and the other is a 100-kc circulator.

621.375.9:621.372.44:621.372.2 60
The Helix Parametric Amplifier: a Broad-Band Solid-State Microwave Amplifier—C. L. Cuccia and K. K. N. Chang. (*RCA Rev.*, vol. 22, pp. 219–244; June, 1961.) The amplifier used a helix as a slow-wave distributed interaction structure and has variable-capacitance solid-state diodes distributively coupled to the helix. Gains above 30 db have been obtained in the 2-Gc region. See also 2673 of 1960 (Conrad, *et al.*).

621.375.9:621.372.44:621.385.6 61
Generalized Coupled Mode Theory—Pease (See 331.)

621.375.927:537.312.62 62
Phase Inverters utilizing Controlled Superconductors—P. M. Chirlian and V. A. Marsocci. (PROC. IRE, vol. 49, pp. 1574–1575; October, 1961.) Two arrangements for phase inversion in a controlled-superconductor push-pull amplifier are illustrated.

621.376.3 63
Transient Response of F.M. Signals—S. J. Cotton. (*Electronic Tech.*, vol. 38, pp. 414–419; November, 1961.) Practical investigation of distortion introduced by networks and lines into a signal when its frequency suddenly changes. The effects of filters, coaxial terminations, and resistive and RC networks are considered.

621.376.332 64
Linear Frequency Discriminator—U. S. Ganguly. (*Electronic Tech.*, vol. 38, pp. 401–406; November, 1961.) A method of distortionless demodulation of LF-FM signals with large deviations is described. Circuit design and performance are discussed and analyzed to determine conditions for maximum linearity.

GENERAL PHYSICS

537.311.3 65
Electrical Conduction Phenomena in Anisotropic Media: Part 1—Strongly Convergent Solutions of Linear Operational Equations, particularly of Bloch's Integral Equation—D. Langbein. (*Z. Phys.*, vol. 162, pp. 542–556; April 27, 1961.) See also 1914 of 1959.

537.533 66
A Note on the Relativistic Boltzmann Equation and some Applications—S. Yadavalli. (*J. Franklin Inst.*, vol. 271, pp. 368–375; May, 1961.) The relativistic equation in the absence of collisions is derived and applied to the case of a velocity-modulated drifting electron beam.

537.533:621.3.032.26 67
Electron Trigonometry—a New Tool for Electron-Optical Design—K. Schlesinger. (PROC. IRE, vol. 49, pp. 1538–1549; October, 1961.) A method is developed for predicting the image formation in electron-optical systems as a substitute for trajectory tracing. It is based on approximating axial potential distributions by linear or parabolic segments. Linear sections are replaced by equivalent drift tubes of appropriately scaled length; parabolic sections observe certain triangle rules for the sum of three phase angles of flight. Applications of the method in the design of electrostatic, magnetic and combined lens systems are illustrated.

537.533.8 68
The Theory of Secondary-Electron Emission of Metals: Part 2—G. Bimschas and G. U. Schubert. (*Z. Phys.*, vol. 162, pp. 382–399; April 14, 1961.) Part 1: 2910 of September (Bimschas).

537.54 69
Average Kinetic Energy of Diffusing Particles in Discharge and Electron Tubes—M. A. Cayless. (*Brit. J. Appl. Phys.*, vol. 12, pp. 523–524; September, 1961.) For particles flowing under a concentration gradient or across a region of a small field in the body of a gas, the mean kinetic energy of particles with a Maxwellian velocity distribution is $5/2\ k\ T$.

537.56 70
The Origin of Plasma Waves in Strongly Ionized Gases—H. Gerstenkorn. (*Z. Phys.*, vol. 162, pp. 363–381; April 14, 1961.) The development of plane waves in a gas plasma without external fields is considered.

537.56 71
Plasma Engineering: Part 3—Applications of Plasma—M. E. Wolff. (*Electronics*, vol. 34, pp. 29–35; September 1, 1961.) Uses of plasma are considered in the fields of thermonuclear research, thermionic and magnetohydrodynamic power conversion, propulsion and electronics. The generation and heating of plasma, and measurement of its parameters are described in Parts 1 and 2, *Electronics*, vol. 34, pp. 47–53 and 33–39; July 14 and August 4, 1961.

537.56 72
Cavity Method for Measurement of High Electron Densities in Plasmas—J. M. Anderson. (*Rev. Sci. Instr.*, vol. 32, pp. 975–978; August, 1961.) Conventional microwave cavity techniques may be used to measure electron

densities considerably above that corresponding to plasma resonance at the particular microwave frequency employed.

537.56:534.13 73
Observations of Ionic Sound Waves in Plasmas—I. Alexeff and R. V. Neidigh. (*Phys. Rev. Lett.*, vol. 7, pp. 223–225; September 15, 1961.) Ionic sound waves were produced in a spherical plasma. Six different gases were used and a relation between the lowest fundamental frequency and atomic mass was found.

537.56:538.566 74
New Form of Plasma Dispersion Formula Usable for Analog Computing—H. Porsche and K. Rawer. (PROC. IRE, vol. 49, pp. 1582–1583; October, 1961.) A correction term is derived which takes into account the influence of a magnetic field.

537.56:538.566 75
Representation of Propagation Parameters for the Plasma in a Magnetic Field—R. E. Haskell and E. H. Holt. (PROC. IRE, vol. 49, p. 1584; October, 1961.) The effect of a dc magnetic field on the properties of a uniform plasma for the propagation of em waves is described using three-dimensional models. See also 1959 of 1960 (Bachynski, *et al.*).

538.56:538.566 76
Study of Electromagnetic Wave Polarization in Magneto-plasmas by a Matrix Method of Crystal Optics—R. E. Haskell and E. H. Holt. (PROC. IRE, vol. 49, pp. 1584–1585; October, 1961). A method is given of studying em wave propagation through magnetoplasmas which is appropriate to microwaves.

537.56:538.63 77
Alfvén Waves in Partially Ionized Gases—T. Watanabe. (*Cancd. J. Phys.*, vol. 39, pp. 1197–1221; August, 1961.) The conditions for formation of an Alfvén wave are determined and fundamental equations are derived to describe the electrodynamic behavior of the ionized gas.

538.566 78
Electromagnetic Surface Waves on a Plane Interface—B. A. Lengyel and K. M. Mitzner. (*J. Appl. Phys.*, vol. 32, pp. 1758–1763; September, 1961.) Surface waves are defined as waves exponentially decreasing on both sides of a plane interface. The relations between the propagation parameters and the material constants are derived in a general case which includes media with different magnetic permeabilities.

538.566 79
Fundamental Solution of the Vectorial Wave Equation in Spherical Coordinates—M. Bouix. (*Ann. Télécommun.*, vol. 16, pp. 125–132; May/June, 1961.) Two families of vectors are determined which, under certain conditions, together form an orthonormalized system tied to the geometry of a sphere.

538.566:535.42 80
Vertex-Generated Waves Outside Metallic Wedges—W. E. Williams. (*Proc. Camb. Phil. Soc.*, vol. 57, pt. 2, pp. 393–400; April, 1961.) The waves generated by a magnetic line source placed at the vertex of a wedge of high conductivity and arbitrary angle are studied. The effect of a dielectric coating on the wedge is examined and the propagation of surface waves considered briefly. See also 2161 of 1961.

538.566:535.43 81
The Low-Frequency Scattering by a Perfectly Conducting Strip—D. S. Jones and B. Noble. (*Proc. Camb. Phil. Soc.*, vol. 57, pt. 2, pp. 364–366; April, 1961.) Previous approximate solutions have not been satisfactory as no expansion parameter suitable for all terms has been found. Results of an analysis suitable for thin cylinders and arbitrary angles of incidence are given.

538.569:539.2 82
Induced and Spontaneous Emission in a Coherent Field: Part 4—I. R. Senitzky. (*Phys. Rev.*, vol. 123, pp. 1525–1537; September 1, 1961.) Previous theoretical work on the interaction of molecules and em fields in a resonant cavity is extended to take account of large perturbations. Part 3: 515 of 1961.

538.569.4:537.228.5 83
Splitting of Electron Spin Resonance Lines by an Applied Electric Field—G. W. Ludwig and H. H. Woodbury. (*Phys. Rev. Lett.*, vol. 7, pp. 240–241; September 15, 1961.)

538.569.4:537.228.5 84
Linear Effect of Applied Electric Field in Electron Spin Resonance—F. S. Ham. (*Phys. Rev. Lett.*, vol. 7, pp. 242–243; September 15, 1961.)

538.569.4:538.222 85
Second-Order Effect of Spin-Orbit Interaction on the Paramagnetic Resonance Spectra of Ions with a Single 3*d* Electron—D. K. Ray. (*Nuovo Cim.*, vol. 21, pp. 1–6; July 1, 1961. In English.)

538.569.4:621.375.9:535.61–2 86
Some Operating Characteristics of Flash-Pumped Ruby Lasers—J. C. Cook. (PROC. IRE, vol. 49, pp. 1570–1571; October, 1961.)

538.569.4:621.375.9:535.61–2 87
Repetitive Hair-Trigger Mode of Optical Maser Operation—M. L. Stitch, E. J. Woodbury, and J. H. Morse. (PROC. IRE, vol. 49, pp. 1571–1572; October, 1961.)

538.691 88
Charged Particle Trajectories in Static Electric and Magnetic Fields—K. J. Ausburn. (*Aust. J. Phys.*, vol. 14, pp. 310–313; June, 1961.) Relative expressions are derived for the curvature and torsion of a trajectory in terms of electric and magnetic field distributions.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164 89
An Examination of the Steady-State Model in the Light of Some Recent Observations of Radio Sources—M. Ryle and R. W. Clarke. (*Monthly Notices Roy. Astron. Soc.*, vol. 122, pp. 349–362; June, 1961.) The unsatisfactory nature of the theoretical model is demonstrated by a comparison of predictions with observational results.

523.164 90
An Investigation of the Clustering of Radio Stars—P. R. R. Leslie. (*Monthly Notices Roy. Astron. Soc.*, vol. 122, pp. 371–380; June, 1961.) An examination of two recent surveys produced no evidence of clustering and an upper limit of 10 per cent could be set to the percentage of sources which occur in clusters.

523.164 91
The Number/Flux-Density Relation for Radio Sources Away from the Galactic Plane—P. F. Scott and M. Ryle. (*Monthly Notices Roy. Astron. Soc.*, vol. 122, pp. 389–397; June, 1961.) Observations at 178 Mc have been used to derive the number/flux-density relation for sources situated at galactic latitudes $>20°$.

523.164:551.507.362.2 92
Radio Astronomy from Space Vehicles—R. C. Jennison. (*Brit. Commun. Electronics*, vol. 8, pp. 502–505; July, 1961.) The usefulness of a radio receiving system in space is considered. Satellite experiments at frequencies outside the range 15–15,000 Mc are suggested.

523.164.3 93
Counts of Extragalactic Radio Sources—R. Van der Borght. (*Aust. J. Phys.*, vol. 14, pp. 295–303; June, 1961.) A formula is derived for the cumulative totals of Class II radio sources and compared with observational measurements.

523.164.3 94
Some Attempts to Detect Linear Polarization of Galactic Radio Emission from the Spur at $l^{II}=30°$—I. K. Pauliny-Toth, J. E. Baldwin, and J. R. Shakeshaft. (*Monthly Notices Roy. Astron. Soc.*, vol. 122, pp. 279–283; June, 1961.) Evidence from measurements at 160, 178 and 408 Mc suggests that the theory of Tunmer (1176 of 1959) does not adequately explain the spur-shaped feature of radio emission.

523.164.32:523.75 95
Slow Drift Solar Radio Bursts: Harmonic Frequency Ratios, Solar Longitude Dependence, and Frequency Drift Rates—M. B. Wood. (*Aust. J. Phys.*, vol. 14, pp. 234–241; June, 1961.)

523.165 96
Latitude Dependence of a Forbush Type Decrease—L. Vir and P. S. Gill. (*Nuovo Cim.*, vol. 20, pp. 498–501; May 1, 1961. In English.)

523.165 97
Radiation Measurements in the Slot between the Van Allen Belts to an Altitude of 1415 Kilometres—C. A. Gurtler. (*J. Geophys. Res.*, vol. 66, pp. 3050–3054; September, 1961.)

523.165 98
Longitudinal Dependence of Radiation-Belt Scattering and Primary Auroral Particles—C. J. Loughnan. (*Planet-Space Sci.*, vol. 8, pp. 13–22; October, 1961.)

523.165 99
World-Wide Distribution of Diurnal Variation of the Neutron Intensity of Cosmic Rays—T. Kanno. (*Nature*, vol. 192, pp. 250–251; October 21, 1961.) Data from stations throughout the world for the period July, 1957 to December, 1958 have been analyzed to detect possible latitude and diurnal effects in periods averaged over six months.

523.165:550.383.4 100
Drift Rate in a Dipole Field—J. S. Lew. (*J. Geophys. Res.*, vol. 66, pp. 2681–2685; September, 1961.) Expressions are found for drift rate and period, and the results are applied to the drift of trapped electrons and protons in the geomagnetic field.

523.165:550.385 101
The Distribution of Trapped Particles in a Changing Magnetic Field—E. N. Parker. (*J. Geophys. Res.*, vol. 66, pp. 2641–2652; September, 1961.) The changes in distribution which result from various changes in the field parameters are calculated.

523.165:551.507.362.2 102
Dynamics and Structure of the Outer Radiation Belt—C. Y. Fan, P. Meyer, and J. A. Simpson. (*J. Geophys. Res.*, vol. 66, pp. 2607–2640; September, 1961.) An analysis of electron measurements made in Explorer VI is presented in terms of four time-independent parameters of the outer Van Allen belt. The belt has two peaks of intensity E_2 max and E_3 max. During a storm E_3 coincides with dipole lines of force. E_3 max is used as a tracer to follow changes in the magnetic field. Some changes are accounted for only by invoking an

irreversible energy gain or loss in the belt. The peaks are observed to move radially inward during a storm and peak electron intensities fall by about 15 per cent.

523.42:621.396.96 103
Radar Contact with Venus—V. A. Kotelnikov. (*J. Brit. IRE*, vol. 22, pp. 293–295, Discussion, pp. 300–303; October, 1961.) In the U.S.S.R., in April, 1961, a transmitter and receiver operating at about 700 Mc were used in conjunction with analyzing and integrating system which enabled both the range of the planet and Doppler spread of received signals to be determined.

523.42:621.396.96 104
Radar Measurements of the Planet Venus—L. R. Malling and S. W. Golomb. (*J. Brit. IRE*, vol. 22, pp. 297–300, Discussion, pp. 300–303; October, 1961.) By means of radar equipment operating at 2388 Mc a new value for the astronomical unit was determined, accurate to 500 km, and the rotation period of Venus was estimated as ~225 days.

523.5:621.391.8 105
Amplitude Distribution for Radio Signals Reflected by Meteor Trails: Part 1—A. D. Wheelon. (*J. Res. NBS*, vol. 64D, pp. 449–454; September/October, 1960.) VHF signals are propagated to distances of 1500 km by reflection from meteor trails. A probability distribution for the envelope of the received signal is derived for the case of reflections from numerous small meteors coincident with residual signals from infrequent large meteors.

550.375:551.594.5 106
Auroral Effects on the Earth's Electric Field—G. D. Freier. (*J. Geophys. Res.*, vol. 66, pp. 2695–2702; September, 1961.) Abnormal conduction currents at the earth's surface are interpreted in terms of currents generated in the atmosphere during three auroras.

550.38:629.19 107
Interaction of West Ford Needles with Earth's Magnetosphere—S. F. Singer. (*Nature*, vol. 192, pp. 303–306; October 28, 1961.) The influence of electrostatic drag forces on the West Ford needles is studied and is shown to be predominant at heights above 1500 km.

550.385.523.75 108
Sun-Earth Relations and Associated Periodicities—I. Paghis. (*Nature*, vol. 192, pp. 346–347; October 28, 1961.) A periodic analysis shows that ssc magnetic impulses tend to recur at 3-day intervals, and that the interval between a flare and a magnetic impulse is about $1\frac{3}{4}$ days.

550.385+523.75]:621.391.812 109
Some Relationships between Short-Wave Fade Outs, Magnetic Crochets and Solar Flares—L. W. Acton. (*J. Geophys. Res.*, vol. 66, pp. 3060–3063; September, 1961.) Statistical analysis indicates a strong association between sudded SW fade outs and magnetic crochets. No significant difference could be found in fade-out type between those fade outs associated with flares of importance greater than 2 and those with flares of importance less than 2.

550.385.3:523.78 110
On the Geomagnetic Effect of the Solar Eclipse of 1961, February 15th at Genoa—M. Bossolasco, A. Elena, and A. Caneva. (*Geofis. pura e appl.*, vol. 48, pp. 102–108; January–April, 1961. In English.)

550.385.4 111
An Idealized Problem of Plasma Dynamics that bears on Geomagnetic Storm Theory—S. Chapman and P. C. Kendall. (*J. Atmos. Terr. Phys.*, vol. 22, p. 142–156; October, 1961.) An idealized problem is solved: a cylindrical sheet of neutral ionized gas is projected into a magnetic field which is aligned along the cylinder axis and varies as R^{-n}, where R is the distance from the axis. For a magnetic field comparable in magnitude with that of the earth the sheet would be stopped at 6.8 earth radii for a 200-γ storm and at 14.6 earth radii for a 20-γ storm.

550.385.4:523.16 112
Interpenetration of Two Ionized Gas Clouds—J. N. Tandon. (*Indian J. Phys.*, vol. 35, pp. 193–198; April, 1961.) Counterstreaming resulting from the collision of two ionized gas clouds is unstable. The double stream forms small clouds of space charge which may be the cause of geomagnetic storms and the radio noise of colliding galaxies.

550.389(73):551.507.362.1 113
Rocket Measurements of the Magnetic Field above New Mexico—R. Hutchinson and B. Shuman. (*J. Geophys. Res.*, vol. 66, pp. 2687–2693; September, 1961.) Absolute total intensity was measured in a rocket at heights up to 236 km and compared with the intensity predicted by spherical harmonic coefficients.

551.507.362.2 114
Measuring the Doppler Frequency Shift on Satellite Transmissions—R. E. Henderson. (*Brit. Commun. Electronics*, vol. 8, pp. 506–512; July, 1961.) The use of a phase-locked oscillator for observation of Doppler shift in transmissions in the range 20–324 Mc is described and the effect of signal modulation on the accuracy of measurements is illustrated by examples.

551.507.362.2 115
Theory of an Accurate Intermediary Orbit for Satellite Astronomy—J. P. Vinti. (*J. Res. NBS*, vol. 65B, pp. 169–201; July–September, 1961.)

551.507.362.2 116
On the Long-Period Lunar and Solar Effects on the Motion of an Artificial Satellite: Part 2—P. Musen. (*J. Geophys. Res.*, vol. 66, pp. 2797–2805; September, 1961.) The theory of lunisolar effects on artificial satellite orbits is presented in a form suitable for numerical integration and for the development of the perturbations into Fourier series with numerical coefficients. Part 1: 3384 of 1961.

551.507.362.2 117
A Comparison of Theory and Observation of the Echo I Satellite—R. Bryant. (*J. Geophys. Res.*, vol. 66, pp. 3066–3069; September, 1961.) A comparison is made between theoretical and observed changes in the perigee height of Echo I due to radiation pressure. The change in atmospheric drag gives information on densities in the 900–1500-km height region.

551.507.362.2 118
Infrared Circuits in Tiros Satellites: Parts 1 and 2—F. Schwarz. (*Electronics*, vol. 34, pp. 43–45 and 136–137; September 22 and 29, 1961.) A 5-channel radiometer for measuring parameters relating to the earth's heat balance, and horizon sensors to determine spin rate and orientation are described.

551.507.362.2 119
Determination of the Earth's Gravitational Potential from Satellite Orbits—D. E. Smith. (*Planet-Space Sci.*, vol. 8, pp. 43–48; October, 1961.) "From the motions of the orbital planes of four satellites, values for the second, fourth and sixth harmonics in the earth's gravitational potential have been obtained. Allowance is made for atmospheric, lunar and solar perturbations and the results are compared with those of other authors."

551.507.362.2:523.75 120
Radioactivity Produced in Discoverer XVII by November 12, 1960, Solar Protons—J. T. Wasson. (*J. Geophys. Res.*, vol. 66, pp. 2659–2663; September, 1961.)

551.507.362.2:551.510.53 121
Latitude and Diurnal Variations of Air Densities from 190 to 280 km as Derived from the Orbits of Discoverer Satellites—G. V. Groves. (*Proc. Roy. Soc.* (*London*) *A*, vol. 263, pp. 212–216; September 5, 1961.) An analysis of orbital information obtained from nine Discoverer satellites shows that variations in air density between day and night in the region 190–280 km are <25 per cent. Curves of daytime density plotted against height indicate the possibility of a region of local heating at 220 km which disappears at night.

551.507.362.2:621.396.43 122
Courier Communication Satellite—Siglin and Senn. (See 293.)

551.510.53 123
Chemical Heating above the Polar Mesopause in Winter—W. W. Kellogg. (*J. Met.*, vol. 18, pp. 373–381; June, 1961.)

551.510.535 124
The Electron Content of the Ionosphere in Winter—J. V. Evans and G. N. Taylor. (*Proc. Roy. Soc.* (*London*), *A*, vol. 263, pp. 189–211; September 5, 1961.) Observations of the Faraday rotation of moon-reflected radio waves at two closely-spaced frequencies are described. Accurate values were obtained for the total electron content of the ionosphere for many hours on successive days. 50 references.

551.510.535 125
The Total Electron Content through the Entire Ionosphere by the Faraday Rotation Technique—O. Burkard. (*J. Geophys. Res.*, vol. 66, pp. 3058–3059; September, 1961.) A linear relation between total electron density and $\sqrt{\cos\chi}$ is demonstrated which presents a possible method of determining variations in intensity of solar ionizing radiation.

551.510.535 126
The Conductive 'Chapman' Layer—C. H. Cummack. (*J. Geophys. Res.*, vol. 66, pp. 2751–2755; September, 1961.) The shape of the electron production function, when ionizing radiation acts as an atmospheric heat source, is given for atmospheres in mixed or diffusive equilibrium. See also 3392 of 1961.

551.510.535 127
Studies of the New Ionospheric Model: Part 2—O. Burkard. (*Geofis. pura e appl.*, vol. 48, pp. 57–84; January–April, 1961. In English.) The original model is perfected on the basis of $N(h)$ profiles and results obtained from moon-echo and radar-scatter observations. Part 1: 148 of 1960.

551.510.535 128
Wind Systems in the Mesosphere and Lower Ionosphere—E. S. Batten. (*J. Met.*, vol. 18, pp. 283–291; June, 1961.)

551.510.535 129
Study of Horizontal Drifts in the *E* Region at Waltair—R. R. Rao and B. R. Rao. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 81–99; October, 1961.) Results obtained over a 2-yr period including the steady 24-hr and 12-hr components found by harmonic analysis are presented. Other drift results, and the *E*-region drifts deduced from *Sq* variations are quoted for comparison.

551.510.535 130
Height Gradient of Horizontal Drifts in the *E* Region over Waltair—R. R. Rao and B. R.

Rao. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 158–160; October, 1961.) A value of 0.74 m/sec/km is obtained between 100 and 120 km. See also 129 above.

551.510.535 131
Evidence of some Geomagnetic Control on the *F*1 Layer—C. H. Cummack. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 157–158; October, 1961.) The bifurcation of the noon *F* layer is found to be related to geomagnetic latitude.

551.510.535 132
Analysis of the *F*1–*F*2 Transition Region—G. A. M. King. (*J. Geophys. Res.*, vol. 66, pp. 2757–2762; September, 1961.) From a detailed study of $N(h)$ profiles, the atmospheric scale height and the relative values of production rates and loss coefficients are deduced.

551.510.535:523.164 133
Effect of Electron-Ion Collisions in the *F* Region of the Ionosphere on the Absorption of Cosmic Radio Noise at 25 Mc/s at Ahmedabad. Changes in Absorption Associated with Magnetic Storms—K. R. Ramanathan, R. V. Bhonsle, and S. S. Degaonkar. (*J. Geophys. Res.*, vol. 66, pp. 2753–2771; September, 1961.) Riometer measurements at a frequency of 25 Mc have been compared with calculated values of total nondeviative absorption (due to collisions with neutral particles and ions) which occurred below *F* maximum, and the differences ascribed to absorption occurring above the layer maximum. The results show a depletion of electrons above the *F* maximum on the day following a sudden commencement and are compared with satellite observations on particle fluxes in Van Allen belts during storms.

551.510.535:523.164:523.78 134
Riometer Observations during a Solar Eclipse—M. A. Gordon. (*Nature*, vol. 192, pp. 249–250; October 21, 1961.) An average decrease of 0.01 db/km in the absorption of cosmic noise transmitted via the *D* region was observed during the total eclipse of October 2, 1959.

551.510.535:523.75 135
Solar Flares and Association with the Lower Ionosphere—T. Goh. (*J. Radio Res. Labs., Japan*, vol. 8, pp. 175–192; May, 1961.) SW fade-outs and f_{min} are adopted as ionospheric parameters. The statistical relations between these and solar flares and solar RF emission are studied. The analysis includes flares from Class 3^+ down to Class 1. The number of fade-outs is correlated with flare activity and with radio noise increases.

551.510.535:523.75 136
A Model of Solar-Flare Induced Ionization in the *D* Region—R. C. Whitten and I. G. Poppoff. (*J. Geophys. Res.*, vol. 66, pp. 2779–2786; September, 1961.) The energy distribution of the *X* radiation in the range 0.4 Å–70 Å was estimated from rocket measurements during a Class 2^+ flare on August 31, 1959. Simultaneous measurements of cosmic noise at 18 Mc showed the changes in *D*-layer absorption during the flare. The absorption can be accounted for by assuming a recombination coefficient of about $2\times10^{-7}/cm^3$ sec and an electron-density profile which varies linearly from 10^4 cm^{-3} at 75 km to 8×10^4 cm^{-3} at 90 km.

551.510.535:523.75 137
Possible Solar Flare Effects in the *F* Region of the Ionosphere—G. H. Munro: R. W. Knecht and K. Davies. (*Nature*, vol. 192, pp. 347–348; October 28, 1961.) Sudden changes in f_0F2 are suggested as being due to the effects of traveling disturbances in the *F*2 layer and not to solar flares [2969 of 1961 (Knecht and Davies)]. In reply the authors point out that on November 12, 1960, a sudden change in f_0F2 occurred simultaneously at stations 10,000 km apart which invalidates Munro's hypothesis. Also, a change in f_0F2 occurs only when the height of the peak of the layer at the time of the flare is less than 315 km.

551.510.535:551.507.362.1 138
Rocket Measurement of a Daytime Electron-Density Profile up to 620 km—J. E. Jackson and S. J. Bauer. (*J. Geophys. Res.*, vol. 66, pp. 3055–3057; September, 1961.) Results are compatible with ground-based and satellite-derived measurements.

551.510.535:551.507.362.1 139
First Pulsed Radio Soundings of the Topside of the Ionosphere—R. W. Knecht, T. E. Van Zandt, and S. Russell. (*J. Geophys. Res.*, vol. 66, pp. 3078–3081; September, 1961.) Preliminary report on a rocket test of the fixed-frequency system to be used in 1962 in a U. S. satellite. While the rocket was above the *F* layer, pulse echoes were observed at 4 and 6 Mc from the ordinary ray *o* and extraordinary ray *x*, reflection levels below. Unexpected phenomena were a) spread and field-guided echoes near the *z*, *o* and *x* reflection levels, and b) local spread echoes at 900-1000 km.

551.510.535:551.507.362.2 140
An Annual and a Semiannual Variation of the Upper Air Density—H. K. Paetzold and H. Zschörner. (*Geofis. pura e appl.*, vol. 48, pp. 85–92; January–April, 1961. In English.) Fluctuations in the acceleration of several satellites are analyzed. See also 3776 of 1961.

551.510.535:551.507.362.2 141
Satellites and the Earth's Outer Atmosphere—D. G. King-Hele. (*Quart. J.R. Met. Soc.*, vol. 87, pp. 265–281; July, 1961.) The structure of the atmosphere as determined by recent experiments is discussed.

551.510.535:621.391.812.63 142
World Maps of f_0F2 during I.G.Y.—K. Sawada and H. Shibata. (*J. Radio Res. Labs., Japan*, vol. 8, pp. 219–286; May, 1961.) Maps showing contours of f_0F2 are reproduced for each month of the year and are based on measurements made during the I.G.Y. when solar activity was very high. Constant Local Time maps are given for 00 and 12; those for a given Universal Time refer to the hours 00, 03, 06, . . . 21.

551.510.536:551.594.6 143
Transresonant Electron Acceleration—E. N. Parker. (*J. Geophys. Res.*, vol. 66, pp. 2673–2676; September, 1961.) Describes a mechanism, additional to that of Helliwell and Bell (3508 of 1960), for the acceleration of Van Allen electrons by whistlers. Nonrelativistic electrons would be accelerated by whistlers of about 1 mv/m field strength. The mechanism depends on the whistler frequency, Doppler shifted by the electron velocity, passing through the electron gyrofrequency.

551.594.5 144
Diurnal Variation of High-Latitude Auroral Frequency on Magnetically Quiet and Disturbed Days—K. Lassen. (*Nature*, vol. 192, p. 345; October 28, 1961.) The frequency of occurrence as a function of time of day is plotted separately for three sectors: north and south of the zenith and centered on the zenith. Quiet and disturbed days are considered separately. The frequency distributions differ from each other and can be interpreted in terms of the movement of the auroral zone relative to the station at Nord, Greenland.

551.594.6 145
Energy Spectrum of Atmospherics and Attenuation of Different Low Frequencies with Distance—S. R. Khastgir and R. S. Srivastava. (*Proc. Nat. Inst. Sci., India*, vol. 26A, Supplement 2, pp. 58–66; December, 1960.) Energy-spectrum curves of atmospherics produced at known distances are derived from simultaneous measurements on four frequencies in the range 3–15 kc. Curves of attenuation for the range 150–1500 km are similar to the theoretical curves of Budden (1652 of 1951).

551.594.6 146
A Note on Whistlers Observed above the Ionosphere—J. C. Cain, I. R. Shapiro, J. D. Stolarik, and J. P. Heppner. (*J. Geophys. Res.*, vol. 66, pp. 2677–2680; September, 1961.) 100 whistlers were observed between September and December, 1959, in satellite Vanguard III between heights of 510 and 3750 km; 90 per cent of these occurred between 1800 and 0600 Local Time indicating low ionospheric absorption at night.

551.594.6 147
Terrestrial Extremely-Low-Frequency Noise Spectrum in the Presence of Exponential Ionospheric Conductivity Profiles—J. Galejs. (*J. Geophys. Res.*, vol. 66, pp. 2787–2792; September, 1961.) Expressions are derived which are applicable to a nonhomogeneous ionosphere. The exponential model gives slightly lower resonant frequencies and less damping of the resonant peaks than a sharply bounded model [2985 of 1961 (Raemer)]. Agreement with measurement is improved.

551.594.6:551.510.535 148
A Study of Ionospherics at Macquarie Island—B. A. McInnes. (*Aust. J. Phys.*, vol. 14, pp. 218–233; June, 1961.) A systematic study of VLF radio emissions of natural origin recorded at one station is presented and the generation of these "ionospherics" is considered on the basis of an extension of the Cherenkov radiation model.

551.594.6:551.510.535 149
Stanford-Pacific Naval Laboratory Conjugate Point Experiment—J. E. Lokken, J. A. Shand, C. S. Wright, L. H. Martin, N. M. Brice, and R. A. Helliwell. (*Nature*, vol. 192, pp. 319–321; October 28, 1961.) Good correlation was found in the times of occurrence of VLF em phenomena observed at high-latitude conjugate locations.

551.594.6:621.391.812.63 150
Computation of Whistler Ray Paths—I. Yabroff. (*J. Res. NBS*, vol. 65D, pp. 485–505; September/October, 1961.) Calculations of whistler-ray paths in the outer ionosphere are illustrated for a variety of electron-density profile models. The variation of paths with frequency, latitude and other variables is examined.

LOCATION AND AIDS TO NAVIGATION

621.396.93:621.391.812.4 151
The Detection of Seasonal Delay Fluctuations of Waves in the 100-kc/s Band by means of an Evaluation of Observations carried out over Many Years on the German Decca Chain—Feyer (See 261.)

621.396.96 152
The Echo Box, its Theory and Application in Radar Techniques—G. Pusch. (*Nachrtech. Z.*, vol. 14, pp. 169–174; April, 1961.)

621.396.96 153
Coincidence Techniques for Radar Receivers employing a Double-Threshold Method of Detection—K. Endresen and R. Hedemark. (Proc. IRE, vol. 49, pp. 1561–1567; October, 1961.) A statistical method is given for assessing whether an object detected on a PPI system is a target or pulsed interference. The

criterion is based upon the number of times the pulses exceed a fixed intensity threshold.

621.396.96:621.391.812 154
Theoretical Scattering Coefficient for Near-Vertical Incidence from Contour Maps—H. S. Hayre and R. K. Moore. (*J. Res. NBS*, vol. 65D, pp. 427–432; September/October, 1961.) Theoretical impressions are derived for the scattering cross section of a terrain. An exponential autocorrelation function is assumed, based on measurements from maps of typical terrain. The expressions are in agreement with experimental results of other authors.

621.396.96:621.391.826 155
Interference in Radar Equipment by Reflections from Sea Movements—J. Brauns and H. G. Möller. (*Hochfrequenz und Elektroak.*, vol. 70, pp. 39–46; May, 1961.) The dependence of the magnitude of sea echoes on various parameters is investigated. The formulas derived are verified experimentally and the design of sea-clutter suppression is discussed.

621.396.969:551.594.2 156
Radar Reflectivity Profiles in Thunderstorms—R. J. Donaldson, Jr. (*J. Met.*, vol. 18, pp. 292–305; June, 1961.)

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 157
Sorption and Replacement of Ionized Gases at a Tungsten Surface—R. B. Burtt, J. S. Colligon, and J. H. Leck. (*Brit. J. Appl. Phys.*, vol. 12, pp. 396–400; August, 1961.) An extension of earlier work relating to the practicability of ion pumping [3266 of 1955 (Brown and Leck)]. Sorption values are given for Ne, A and Kr and show that a maximum sorption exists at any particular bombarding ion energy. A value for the depth of penetration of the ions is derived.

535.215 158
Cross-Section Ratios of Sensitizing Centres in Photoconductors—R. H. Bube. (*J. Appl. Phys.*, vol. 32, pp. 1707–1709; September, 1961.) The charge on sensitizing centers is determined by measuring the ratio of hole-capture cross section to electron-capture cross section. Most centers are singly charged although doubly-charged centers are also found.

535.215 159
Photovoltaic Effects in Cs_3Sb Films—F. Wooten. (*J. Appl. Phys.*, vol. 32, pp. 1789–1790; September, 1961.) Photovoltaic effects and their implications with regard to band bending are considered.

535.215:537.533.8 160
Effect of Contact Material on the Cathodoconductivity of Cadmium Sulphide and Cadmium Selenide—A. V. Simashkevich, M. V. Kot, and L. M. Panasyuk. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1035–1037; April, 1961.) Investigation by means of an electron-beam "probe" of the effect of ohmic and nonohmic contacts on the distribution of electron-bombardment-induced conductivity along CdS and CdSe crystals.

535.215:546.48'221 161
Photoconduction in Cadmium Sulphide—J. F. Duncan and D. N. S. Rao. (*Brit. J. Appl. Phys.*, vol. 12, pp. 511–513; September, 1961.) Discrete voltage pulses, of amplitude depending on the energy of incident radiation, have been observed in the photoconduction. Between 4800 and 8000 Å the pulse height and the photocurrent vary in a similar manner with λ.

535.215:546.48'221 162
Effect of Current Injection on Properties of Cadmium Sulphide Crystals—H. Mitsuhashi. (*J. Phys. Soc. Japan*, vol. 16, pp. 1258–1259; June, 1961.) After excitation at 10^{-2} mm Hg by a Tesla coil, conductivity of the crystals increased by several orders of magnitude and photoconductivity became less important than the dark current. An explanation is given.

535.215:546.48'221 163
High Electric Fields in Cadmium Sulphide: Field Effect Constriction of Current Flow and Dielectric Breakdown—R. Williams. (*Phys. Rev.*, vol. 123, pp. 1645–1651; September 1, 1961.) Field-effect measurements are used to study constriction of current flow in very thin single crystals of conductivity CdS. A complete "pinch-off" of current gives 1.8×10^6 v/cm for the breakdown field strength at 25°C.

535.215:546.48'221 164
The Problem of Infrared Quenching of Photoconductivity in CdS Single Crystals—V. V. Serdyuk. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1229–1231; April, 1961.) CdS crystals with a main band of photosensitivity in the region of 520 mμ show an additional band with λ_{max} = 700 mμ after heating in air or vacuum at 300°C. Infrared quenching in this region is less than in the main band.

535.215:546.48'221:534.14–8 165
Ultrasonic Amplification in CdS—Hutson, McFee, and White. (Sec 1.)

535.215:546.817'221 166
Investigation of the Surface of Lead Sulphide by Adsorption Methods—I. A. Berezhnaya and L. N. Kurbatov. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1038–1043; April, 1961.) Results obtained by the two methods described show that highly sensitive photoconductive layers of lead sulphide have a relatively small specific surface area.

535.215:546.817'231 167
The Electrical Conductivity of Vapour-Deposited Lead Selenide Films—H. Gobrecht, F. Niemeck, and K. E. Boeters. (*Z. Phys.*, vol. 162, pp. 337–346; April 14, 1961.) Measurements of conductivity as a function of temperature and Se vapor pressure were carried out on films prepared by the method given in 2244 of 1961.

535.37 168
Two-Photon Excitation in $CaF_2:Eu^{3+}$—W. Kaiser and C. G. B. Garrett (*Phys. Rev. Lett.*, vol. 7, pp. 229–231; September, 15, 1961.)

535.37:546.47–31 169
The Relation between Electrical Dark Conductivity and Green Luminescence of ZnO Crystals—E. Mollwo. (*Z. Phys.*, vol. 162, pp. 557–569; April 27, 1961.) Measurements were made on single crystals prepared by the method given in 878 of 1959 (Bogner and Mollwo).

535.376:546.47'221:539.23 170
Thin-Film Elucidation of the Electroluminescence Process—W. A. Thornton. (*Phys. Rev.*, vol. 123, pp. 1583–1586; September 1, 1961.) "Very thin sulphide films emit light only on alternate half-cycles of the voltage sine wave, whereas the emission of thicker films and phosphor powders is quite symmetrical with polarity. This asymmetry of emission, together with clipping or dc bias of the applied voltage, is used to confirm unambiguously that the excitation and recombination steps in electroluminescence are separable and occur sequentially and under different field configurations, and that the recombination is field-driven."

535.376:546.47'48'221 171
Positive-Hole Motion and Photovoltaic Effects in Zinc Cadmium Sulphide Phosphors—H. Kallmann, B. Kramer, F. Spagnolo, and G. M. Spruch. (*Phys. Rev.*, vol. 123, pp. 1661–1665; September 1, 1961.)

537.226:538.566 172
Reciprocal Relations in an *N*-Slab Dielectric—R. Hollis. (PROC. IRE, vol. 49, p. 1579; October, 1961.) A proof is given showing that when radio waves are propagated through a dielectric consisting of parallel homogeneous slabs, the transmission coefficients are identical for both directions of transmission but the reflection coefficients are not.

537.226:546.42'87'824–31 173
Investigation of the Dielectric Properties of Strontium-Bismuth Titanates at Low Temperatures—A. N. Gubkin, A. M. Kashtanova, and G. I. Skanavi. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1110–1116; April, 1961.)

537.226:621.396.677.85 174
Permittivity and Permeability of a Medium containing Ellipsoidal Metal Particles in a Cubical Array—Krishnaji and S. Swarup. (*J. Inst. Telecommun. Engrs., India*, vol. 7, pp. 117–128; May, 1961.) Expressions for the permittivity and permeability are derived and results are presented for a typical case of a metal-flake artificial dielectric.

537.227 175
Ferroelectric Polarization Reversal by Linear Fields—H. Shibata, H. Toyoda, and S. Waku. (*Rev. Elec. Commun. Lab., Japan*, vol. 9, pp. 279–286; May/June, 1961.) Results are given for triglycine sulphate, and pure and impurity-doped $BaTiO_3$ crystals subjected to linearly rising field pulses.

537.227 176
Ferroelectricity in (K,NH_4), (K,Rb) and (K,Tl) Ferrocyanide Mixed Crystals—S. Waku, K. Masuno, and T. Tanaka. (*J. Phys. Soc. Japan*, vol. 16, pp. 1158–1161; June, 1961.) Substitution of other monovalent ions for the potassium ion in KCN causes the peak of the dielectric-constant/temperature curve to occur at lower temperatures. Compounds with relatively low substitution show ferroelectric behavior similar to KCN trihydrate.

537.227:546.431'824–31 177
Surface Layer in $BaTiO_3$ Single Crystals—E. Fatuzzo and W. J. Merz. (*J. Appl. Phys.*, vol. 32, pp. 1685–1687; September, 1961.) The dependence of switching time and coercive field on crystal thickness is assumed to be due to a thin layer of electrically biased, mechanically strained $BaTiO_3$ on the surface. All experimental features cannot be explained either by this model or by that given by Drougard and Landauer (923 of 1960).

537.227:547.476.3 178
Differing Ferroelectric Properties in Different Regions of Rochelle Salt Crystals—H. E. Müser and H. G. Unruh. (*Z. Phys.*, vol. 162, pp. 313–328; April 5, 1961.) The differences observed are attributed to dislocations and impurities. See also 2250 of July (Müser).

537.228.1:534.286–8 179
Absorption of Sound in Insulators—T. O. Woodruff and H. Ehrenreich. (*Phys. Rev.*, vol. 123, pp. 1553–1559; September 1, 1961.) The theory of attenuation of acoustic waves in crystals is extended and applied to experiments with waves of frequency 1–24 Gc in quartz; see also 3135 of 1960 (Bömmel and Dransfeld).

537.311.3 180
Electron Transport at High Temperatures in the Presence of Impurities—H. L. Frisch and J. L. Lebowitz. (*Phys. Rev.*, vol. 123, pp. 1542–1549; September 1, 1961.) A theoretical treatment of electron scattering by phonons and impurities in solids, using Maxwellian statistics.

537.311.3:539.23:538.632 181
A Device to Determine the Electrical Conductivity and Hall Constant of Thin Metallic and Semiconducting Films in a Vacuum—S. Martinuzzi. (*J. Phys. Radium*, vol. 22, pp. 247–248; April, 1961.)

537.311.33 182
Influence of Electric Field on Electron Temperature, Electrical Conductivity and Thermionic Emission of Semiconductors: Part 2—Effect of Plasma Oscillations and Short-Range Collisions—P. M. Tomchuk. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1019–1030; April, 1961.) Theoretical treatment based on the distribution function for an electron gas in a constant field. Part 1: *Fiz. Tverdogo Tela.*, vol. 2, pp. 2228–2239; September, 1960 (Dykman and Tomchuk).

537.311.33 183
Theory of the Field Effect at Low Temperature—Yu. I. Gorkun. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1061–1065; April, 1961.) The space charge and conductivity variation at the surface of a semiconductor are calculated as a function of zone inflection for an arbitrary filling of local bands.

537.311.33 184
Statistics of Electrons and Holes in Semiconductors with Dislocations—Yu. V. Gulyaev. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1094–1100; April, 1961.) The degree of filling of dislocations as a function of temperature and of position of the Fermi level at low temperatures is determined on the basis of a general consideration of the energy spectrum of charge carriers.

537.311.33 185
Variational Method of Determining Electrical Conductivity taking into Account the Coulomb Interaction between Carriers—F. M. Tomchuk. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1258–1267; April, 1961.) An expression is derived for calculating the conductivity of an electron gas in semiconductors and in plasma, which does not vary with changes in the asymmetric portion of the distribution function.

537.311.33 186
The Properties of Semiconductors with Extremum Loops: Part 3—Behaviour in a Magnetic Field Parallel to the Plane of the Loop—É. I. Rashba and I. I. Boĭko. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1277–1289; April, 1961.) Quasi-classical wave functions are developed for the parallel-field case and resonance effects and oscillations of magnetic susceptibility are considered. Part 1: 1209 of 1961 (Rashba). Part 2: *Fiz. Tverdogo Tela.*, vol. 2, pp. 1874–1883; August, 1960 (Boĭko and Rashba).

537.311.33 187
The Theory of Zone Melting of Semiconductors in the Presence of a Gaseous Phase—O. Madelung. (*Z. Phys.*, vol. 162, pp. 508–515; April 27, 1961.) Zone purification of elements containing volatile impurities is considered under assumptions differing from those used by van den Boomgaard (819 of 1956).

537.311.33 188
Resonance Transfer of Ionization Energy in Semiconductors—S. Koshino and T. Ando. (*J. Phys. Soc. Japan*, vol. 16, pp. 1151–1157; June, 1961.) Thermal conductivity due to resonance transfer is calculated for the extrinsic, exhaustion and intrinsic ranges.

537.311.33 189
Hot-Electron Behaviour at High Pressures—L. Gold. (*J. Phys. Soc. Japan*, vol. 16, p. 1269; June, 1961.) A note on the application of the theory developed in 3038 of September.

537.311.33 190
Description of Impurity Ionization in Semiconductors by Chemical Thermodynamics—W. W. Harvey. (*Phys. Rev.*, vol. 123, pp. 1666–1673; September 1, 1961.)

537.311.33 191
Low Temperature Electric Field Effects in Semiconductors—V. V. Paranjape. (*Proc. Phys. Soc.* (*London*), vol. 78, pp. 516–528; October 1, 1961.) A theoretical investigation of the rise in conductivity of Ge at liquid He temperature at about 2 v/cm. The variation of electron density with field is calculated considering excitation of donors into the conduction band by impact ionization and phonon absorption and also the reverse processes. An expression for current density as an explicit function of field strength is obtained and good agreement with experiment found.

537.311.33:535.215 192
Determination of certain Semiconductor Parameters from the Spectral Distribution of Photoconductivity—V. A. Petrusevich. (*Fiz. Tverdogo Tela.*, vol. 3, pp. 1268–1271; April, 1961.) Development of a method of determining recombination constants and absorption spectra.

537.311.33:546.26–1 193
The Magnetoresistance of *p*-Type Semiconducting Diamond—P. J. Kemmey and E. W. J. Mitchell. (*Proc. Roy. Soc.* (*London*) *A*, vol. 263, pp. 420–432; September 19, 1961.) Measurements on three oriented diamonds at room temperature are described and results are compared with those of calculations.

537.311.33:546.26–1 194
Carrier Densities and Mobilities in Pyrolytic Graphite—C. A. Klein and W. D. Straub. (*Phys. Rev.*, vol. 123, pp. 1581–1583; September 1, 1961.)

537.311.33:[546.28+546.289 195
Experiments on the Variation in Surface Conductivity of Germanium and Silicon—M. L. Ioselevich and V. I. Fistul'. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1132–1136; April, 1961.) An experimental investigation of *n*- and *p*-type Ge and Si single crystals coated with oxidizing and reducing films of $KMnO_4$ and $SnCl_2$ types.

537.311.33:546.28 196
Distribution Coefficient of Antimony in Silicon from Solvent Evaporation Experiments—F. A. Trumbore, P. E. Freeland and R. A. Logan. (*J. Electrochem. Soc.*, vol. 108, pp. 458–460; May, 1961.) Distribution coefficients were calculated from Hall-effect measurements on heavily-doped Si crystals grown from melts containing Sb. Nearly a twofold increase in the coefficient was observed in the concentration range investigated.

537.311.33:546.28 197
Slip Patterns on Boron-Doped Silicon Surfaces—H. J. Queisser. (*J. Appl. Phys.*, vol. 32, pp. 1776–1780; September, 1961.)

537.311.33:546.28 198
The Display of *p-n* Junctions and Boundary Layers—E. Mollwo and C. Solbrig. (*Naturwiss.*, vol. 48, pp. 397–398; May, 1961.) A mixture of sulphur and red-lead dust is allowed to settle on the junction. The constituents of the mixture separate and a dust-free gap of width depending on the potential applied to the junction is formed between the *p* and *n* regions. Results of tests on a Si diode are illustrated.

537.311.33:546.289 199
The Effect of Surface States on the Hall Effect and Magnetoresistance of Germanium—T. N. Sytenko and O. N. Koshel'. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1079–1084; April, 1961.) Results of measurements of the variation of Hall EMF and magnetoresistance with external electric field are in agreement with the theory of Petritz (3497 of 1958).

537.311.33:546.289 200
Effect of Gaseous Ambients upon 1/f Noise in Germanium Filaments—V. E. Noble and J. E. Thomas, Jr. (*J. Appl. Phys.*, vol. 32, pp. 1709–1714; September, 1961.) Gaseous ambients were used to change the surface potential of samples 4-μ thick to find the 1/f noise variation. A minimum was found when surface potential corresponded to conductance minimum. The amplitude variations of accumulation noise and inversion noise were measured.

537.311.33:546.289 201
A Diffusion Mask for Germanium—E. L. Jordan. (*J. Electrochem. Soc.*, vol. 108, pp. 478–481; May, 1961.) A mask of SiO_2 completely screens against diffusion of *n*-type impurities under all conditions encountered in device fabrication. It is easy to apply and may be removed by etches which do not affect Ge.

537.311.33:546.289:535.215 202
The Effect of Surface Treatment on Photoemission from Germanium—Z. P. But. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1137–1143; April, 1961.)

537.311.33:546.289:538.569.4 203
Multicomponent Magnetoplasma Resonance in *p*-Type Germanium—R. E. Michel and B. Rosenblum. (*Phys. Rev. Lett.*, vol. 7, pp. 234–236; September 15, 1961.) The position of the resonance line for *p*-type Ge is plotted against carrier concentration. The observed power absorption, for different values of carrier concentration, is compared with theory.

537.311.33:546.49'231:538.63 204
Angular Dependence of Magnetoresistance in HgSe—T. C. Harman. (*J. Appl. Phys.*, vol. 32, pp. 1800–1801; September, 1961.) Measurements of the magnetoresistance show that the conduction band is spherically symmetric.

537.311.33:546.681'19 205
Crystal-Pulling Furnace for Gallium Arsenide—R. Wickham. (*J. Sci. Instr.*, vol. 38, pp. 396–399; October, 1961.)

537.311.33:546.681'19 206
Properties of High-Resistivity Gallium Arsenide Compensated with Diffused Copper—J. Blano, R. H. Bube, and H. E. MacDonald. (*J. Appl. Phys.*, vol. 32, pp. 1666–1679; September, 1961.)

537.311.33:546.681'86 207
On the Secondary Tunnelling Phenomena in a *p-n* Junction of Gallium Arsenide—A. Shibata. (*J. Phys. Soc. Japan*, vol. 16, p. 1261; June, 1961.) Experimental results indicate the existence of impurity bands in the energy gap and distortion of the energy band edges.

537.311.33:[546.682'19+546.682'86 208
Free-Carrier Voigt Effect in Semiconductors—S. Teitler, E. D. Palik, and R. F. Wallis. (*Phys. Rev.*, vol. 123, pp. 1631–1633; September 1, 1961.) Measurements of the Voigt and

the Faraday effects have been made on samples of *n*-type InAs and InSb. The Voigt and Faraday data are used together to obtain values of both the effective mass and concentration of free carriers which are consistent with values given by other methods. An experimental technique for measurement of small phase shifts in the Voigt effect is described.

537.311.33:546.682'19 209
Optical Properties on *n*-Type Indium Arsenide in the Fundamental Absorption Edge Region—J. R. Dixon and J. M. Ellis. (*Phys. Rev.*, vol. 123, pp. 1560–1566; September 1, 1961.)

537.311.33:546.682'19 210
Investigation of the Dependence of the Effective Electron Mass in *n*-Type InAs on Charge-Carrier Concentration—F. M. Gashimzade and F. P. Kesamanly. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1255–1257; April, 1961.) An equation is derived from the thermal EMF taking into account the nonparabolic structure of the conduction band.

537.311.33:546.682'86 211
Dependence of the Effective Electron Mass in *n*-Type InSb on the Charge-Carrier Concentration—O. V. Emel'yanenko, F. P. Kesamanly and D. N. Nasledov. (*Fiz. Tverdogo Tela.*, vol. 3, pp. 1161–1163; April, 1961.) An increase in the effective electron mass has been observed with increase in carrier density.

537.311.33:546.682'86 212
Electrical and Galvanomagnetic Properties on *n*-Type InSb at Low Temperatures—Lyan' Chzhi-Chao and D. N. Nasledov. (*Fiz. Tverdogo Tela.*, vol. 3, pp. 1185–1189; April, 1961.) An investigation of the impurity band and the oscillatory magnetoresistance in specimens with impurity concentration 10^{12}–10^{15} cm^{-3} in the range 4–77°K.

537.311.33:546.873'241 213
Effect of Pressure on the Energy Gap in Bi_2Te_3—C. Y. Li, A. L. Ruoff, and C. W. Spencer. (*J. Appl. Phys.*, vol. 32, pp. 1733–1735; September, 1961.)

537.311.33:548.5 214
The Form of the Crystallization Front in Semiconductor Single Crystals Grown from the Melt by the Czochralski Method—M. G. Mil'vidskiĭ and B. I. Golovin. (*Fiz. Tverdogo Tela.*, vol. 3, pp. 1015–1018; April, 1961.) Description of a method for controlling crystal pulling in accordance with a calculated growth-velocity program.

537.312.62 215
Enhancement of Superconductivity by Extraction of Normal Carriers—R. H. Parmenter. (*Phys. Rev. Lett.*, vol. 7, pp. 274–277; October 1, 1961.) Electrical extraction of thermally generated normal carriers from a superconductor should maintain the metal superconducting above its usual transition temperature. Conversely, injection of normal carriers should lower the transition temperature.

537.323:546.3–1'48'46 216
The Thermoelectric Power of Magnesium-Cadmium Alloys—S. Noguchi. (*J. Phys. Soc. Japan*, vol. 16, pp. 1145–1150; June, 1961.) Measurements were made over a wide range of composition and the data analyzed to obtain information about the band structure of the alloys in the Mg-Cd system.

537.581 217
Thermionic Emission of UC-Nb—R. H. Abrams, Jr., and F. E. Jamerson. (*J. Appl. Phys.*, vol. 32, pp. 1783–1784; September, 1961.)

537.581 218
Thermionic Emission from a Tantalum Crystal in Cesium or Rubidium Vapour—H. F. Webster. (*J. Appl. Phys.*, vol. 32, pp. 1802–1803; September, 1961.)

538:061.3 219
Proceedings of the Sixth Symposium on Magnetism and Magnetic Materials—(*J. Appl. Phys.*, vol. 32, Supplement, pp. 2S-395S; March, 1961.) The text is given of 149 papers presented at the conference held in New York, November 14–17, 1960.

538.221 220
Antinucleation Processes in Ferromagnetics—A. Jaśkiewicz. (*Phil. Mag.*, vol. 6, pp. 1057–1059; August, 1961.)

538.221 221
Permanent-Magnet Properties of Malleable Fe-Mn-Ti Alloys—G. Rassmann and O. Henkel. (*Z. angew. Phys.*, vol. 13, pp. 169–170; April, 1961.)

538.221 222
Square-Loop Permalloy Alloys—F. Pfeifer. (*Z. angew. Phys.*, vol. 13, pp. 177–180; April, 1961.) Investigation of the properties of permalloy-type materials subjected to various forms of heat treatment. See also 3080 of 1961.

538.221 223
The Influence of Cold Forming and Heat Treatment on the Electrical and Magnetic Properties of Fe-Al Alloys—G. Rassmann and H. Wich. (*Z. angew. Phys.*, vol. 13, pp. 172–174; April, 1961.)

538.221:537.581 224
The Ferromagnetic Anomaly in the Work Function of Nickel—A. Gel'berg, B. Iosifesku, and G. Komsha. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1074–1078; April, 1961.)

538.221:538.632 225
The Magnetostriction of Ferromagnetic Crystals below Saturation—R. R. Birss and E. W. Lee. (*Proc. Roy. Soc. (London), A*, vol. 263, pp. 387–394; September 19, 1961.)

538.221:539.23 226
The Theory of Magnetization Reversal of Thin Films: Part 1—Nucleation in Homogeneous Material—W. Andrä. (*Mber. Disch. Akad. Wiss. Berlin*, vol. 2, no. 7, pp. 398–402; 1960.) Simplified treatment on the basis of domain theory to investigate the possibility of inhomogeneous rotation of magnetization and the nucleation process. See also 4217 of 1961.

538.221:539.23 227
Angular Dependence of Torque in Anisotropic Permalloy Films—W. D. Doyle, J. E. Rudisill, and S. Shtrikman. (*J. Appl. Phys.*, vol. 32, pp. 1785–1787; September, 1961.)

538.221:621.318.018.756 228
How Magnetic Materials Behave at Nanosecond Pulse Widths—G. A. Reeser. (*Electronics*, vol. 34, pp. 72–75; September 8, 1961.) The results of performance tests on tape-wound and ferrite cores at 30-nsec pulse width are illustrated. The reliability of the data is limited by the tolerances of the magnetic properties of the material.

538.221:621.318.124 229
Magnetic Circuits employing Ceramic Magnets—J. C. Helmer. (PROC. IRE, vol. 49, pp. 1528–1537; October, 1961.) The properties of brick-shaped ceramic magnets are discussed and configurations of ceramic-iron and ceramic-steel magnet systems for producing magnetic fields of a desired distribution are considered.

538.221:621.318.134 230
The Solubility of Nickel in Nickel Ferrite—R. Parker and M. S. Smith. (*J. Phys. Chem. Solids*, vol. 21, pp. 76–80; October, 1961.) Electrical measurements indicate that excess Ni dissolves in the ferrite lattice up to a Ni/Fe ratio of 1.029:2.

538.221:621.318.134 231
Some Methods and Results in the Study of Cubic Ferrites—K. Hoselitz. (*Z. angew. Phys.*, vol. 13, pp. 182–184; April, 1961. In English.) Methods of single-crystal preparation are described and the results of measurements on Co in Mn ferrite and on garnets are mentioned.

538.221:621.318.134 232
Magnetization Reversal of Square-Loop Ferrites—G. Vogler. (*Z. angew. Phys.*, vol. 13, pp. 185–187; April, 1961.) A discussion of reversal mechanism on the basis of a simplified model.

538.221:621.318.134 233
Orientation Superstructure in Fe-Ni Ferrite Single Crystals—R. Wagner. (*Z. angew. Phys.*, vol. 13, pp. 187–189; April, 1961.) Report of investigations of anisotrophy changes as a function of orientation and time.

538.221:621.318.134 234
Sublattice Magnetic Moments in Manganese Ferrite—W. P. Osmond. (*Proc. phys. Soc. (London)*, vol. 78, pp. 620–622; October 1, 1961.) The difference between the observed and theoretical moments may be due to the existence of canted spins. A way in which these could arise is suggested.

538.221:621.318.134 235
Thermodynamic and Magnetic Properties of Yttrium Iron Garnet—J. W. Henderson and R. L. White. (*Phys. Rev.*, vol. 123, pp. 1627–1630; September 1, 1961.)

538.221:621.318.134 236
Zero-Field Splitting of the Cr^{3+} Ground State in Y-Ga and Y-Al Garnet—J. W. Carson and R. L. White. (*J. Appl. Phys.*, vol. 32, p. 1787; September, 1961.)

538.221:621.318.134:538.569.4 237
Study of Saturation and Measurement of Relaxation Time in Ferromagnetic Resonance of Yttrium Garnet—M. Sauzade. (*Ann. Phys. (Paris)*, vol. 6, pp. 595–642; May/June, 1961.) 41 references.

621.315.612.6 238
On the Dielectric Losses and Permittivity of Glasses at Ultra High Frequencies—M. D. Mashkovich. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1105–1109; April, 1961.) A report of measurements on seven glasses in the frequency range 100 cps–37.5 Gc at temperatures from 20–400°C.

MATHEMATICS

512.831 239
Spin Matrix Exponentials and Transmission Matrices—L. Young. (*Quart. Appl. Math.*, vol. 19, pp. 25–30; April, 1961.)

517.564 240
Addition Theorems for Spherical Wave Functions—S. Stein. (*Quart. Appl. Math.*, vol. 19, pp. 15–24; April, 1961.)

517.94 241
The Mathematical Treatment of Linear Systems with Time-Variable Parameters—G. Wunsch. (*Nachrtech.*, vol. 11, pp. 221–222; May, 1961.) Application of the treatment of time-invariant linear systems to time-variable systems for the solution of Duhamel's integral.

517.949:621.375.9:621.372.44 242
Solution of a Difference Equation Pertinent to Linear, Parametric Electric Networks—J. J. Leon. (*Quart. Appl. Math.*, vol. 19, pp. 63–68; April, 1961.) An algorithm is presented for finding the fundamental system of solutions of a class of linear difference equations.

519.24 243
Moments of the Generalized Rayleigh Distribution—J. H. Park, Jr. (*Quart. Appl. Math.*, vol. 19, pp. 45–49; April, 1961.)

MEASUREMENTS AND TEST GEAR

621.317.326:621.374 244
Pulse Voltage Comparator measures Height of Positive or Negative Pulses—O. B. Laug. (*Electronics*, vol. 34, pp. 70–71; September 8, 1961.) The circuit responds to pulses of width 50 nsec and duty cycle 10^{-7}.

621.317.331 245
Measurement of Electrical Resistivity by a Mutual Inductance Method—R. G. Chambers and J. G. Park. (*Brit. J. Appl. Phys.*, vol. 12, pp. 507–510; September, 1961.) Simple equipment for measuring mutual inductance over a range of frequencies can be applied to the measurement of resistivity from 2×10^{-9} Ω cm. The interpretation of results is discussed.

621.317.34:621.372.3 246
A New Method of Measuring Very Small Standing Waves in Waveguides—S. Hari-Haran and W. S. Stuart. (*J. Inst. Telecommun. Engrs., India*, vol. 7, pp. 149–152; May, 1961.) The method is based on inserting into the system a short length of waveguide provided with a nonradiating slot and a small probe.

621.317.361:621.317.733 247
Sensitivity Error in Frequency Measurements with the Wien Bridge—H. H. Wolff. (*Rev. Sci. Instr.*, vol. 32, pp. 902–905; August, 1961.)

621.317.73:621.385.1 248
An Input Admittance Meter for Electron Tubes—T. Kojima and H. Hara. (*Rev. Elec. Commun. Lab., Japan*, vol. 9, pp. 309–318; May/June, 1961.) A description of a measuring instrument for use at 70 Mc with details of its performance and experimental results.

621.317.733 249
Calibration of Inductance Standards in the Maxwell-Wien Bridge Circuit—T. L. Zapf. (*J. Res. NBS.*, vol. 65C, pp. 183–188; July–September, 1961.) Residual errors in the circuit are discussed and their effects on inductor measurements considered in a bridge without a Wagner earth.

621.317.755 250
Sampling Techniques Applied to High-Speed Pulse Oscillography—B. Gilbert. (*Mullard Tech. Comman.*, vol. 5, pp. 317–329; June, 1961.) Gain-bandwidth products are 100 times greater than those of conventional oscilloscopes. Display at full brilliance is possible even on large cathode-ray tubes, or an *X-Y* graph plotter may be operated.

621.317.755:621.374 251
Distortion of Steep Pulse Edges due to Finite Electron Transit Time between Deflection Plates—H Lotsch. (*Frequenz*, vol. 15, pp. 163–169; May, 1961.) The distortion of pulse edges in CRO displays is calculated for inclined and parallel deflection plates. See also 2743 of 1961.

621.317.772:621.373.52 252
Measuring Phase with Transistor Flip-Flops—J. R. Woodbury. (*Electronics*, vol. 34, p. 56; September 22, 1961.) A circuit is given for measurement of phase difference to within about 1° in the range 0.2–20 kc.

621.317.799:621.397.6 253
An Instrument for Measuring Television Signal-to-Noise Ratio—S. M. Edwardson. (*B.B.C. Engrg. Div. Monographs*, pp. 5–23; June, 1961.) A range of SNR from 9–51 db can be measured at any grey level with an accuracy of ±0.5 db.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

551.510.62:621.372.413 254
A Battery-Driven Microwave Refractometer of Small Size—W. Steffen. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 47–54; May, 1961.) The refractometer described in 993 of 1960 (Schünemann and Steffen) has been reduced in size by replacing tubes with transistors, and its performance has been improved.

621.362 255
Thermionic Generation of Electricity—M. A. Cayless. (*Brit. J. Appl. Phys.*, vol. 12, pp. 433–442; September, 1961.) A review of recent work and possible future trends from the standpoint of physical processes involved. Over 80 references.

621.375.9:621.372.44:621.313.13 256
Tunnel-Diode Superregenerative Parametric Motor—H. E. Stockman. (PROC. IRE, vol. 49, pp. 1586–1587; October, 1961.) A scheme for applying tunnel-diode "stimulance" to produce rotary motion is described, enabling a dc motor to be designed which is suitable for use in space vehicles. See also 2751 of 1961.

621.385.83:538.691 257
A New Experiment for the Visual Display of Magnetic Field Lines for Measurement Purposes—G. Petter. (*Naturwiss.*, vol. 48, pp. 398–399; May, 1961.) A pencil-beam electron-tube arrangement is described which enables photographs to be taken of electron trajectories in a magnetic field.

621.385.833 258
A Method of Numerical Integration and Marginal Trajectories in Electron Optics—R. Lapeyre and M. Laudet. (*C. R. Acad. Sci., (Paris)*, vol. 251, pp. 1874–1876; October 31, 1960.) An examination of the conditions for convergence of an iterative method for integration. A new formula of the Runge-Kutta type permits approximations of 5th and 4th orders, respectively, for the function and its derivative.

621.385.833:621.317.42 259
Development of an Electron-Optical Method into a Method of Magnetic Measurement—H. Murrmann and C. Schwink. (*Z. angew. Phys.*, vol. 13, pp. 189–193; April, 1961.)

Part 1: Experimental Arrangement and Measurements (pp. 189–191)—The application of the electron-optical shadow technique is considered. See also 3441 of 1959 (Schwink and Murrmann) and back references.

Part 2: Theoretical Considerations (pp. 192–193.)

621.56/.57:537.322.15 260
The Frigistor Diagram—W. Dörr. (*Elektron. Rundschau*, vol. 15, pp. 156–158; April, 1961.) A diagram is given facilitating the design of thermoelectric cooling devices of semiconductor material (see 4272 of 1961).

PROPAGATION OF WAVES

621.391.812.4:621.396.93 261
The Detection of Seasonal Delay Fluctuations of Waves in the 100-kc/s Band by means of an Evaluation of Observations carried out over Many Years on the German Decca Chain—W. Feyer. (*Nachrtech. Z.*, vol. 14, pp. 183–196; April, 1961.) Detailed analysis of observations made during the period 1952–1959 with the aim of determining the errors introduced by seasonal fluctuations to improve the accuracy of the system.

621.391.812.6:537.56 262
Reflection of Electromagnetic Waves at Electron Density Ramps—L. S. Taylor. (*J. Appl. Phys.*, vol. 32, pp. 1796–1797; September, 1961.) The problem is analyzed mathematically. See also 1619 of 1961 (Albini and Jahn).

621.391.812.62 263
On the Theory of Wave Propagation Through a Concentrically Stratified Troposphere with a Smooth Profile: Part 1—Discussion of the Extended W.K.B. Approximation—H. Bremmer. (*J. Res. NBS*, vol. 64D, pp. 467–482; September/October, 1960.)

621.391.812.62.029.63:621.396.65 264
The Influence of the Equivalent Gradient and Atmospheric Stability on Transhorizon Links in the Sahara and the Congo—P. Misme. (*Ann. Télécommun.*, vol. 16, pp. 110–116; May/June, 1961.) A variation of 20 db in a link of 170 km at 471 Mc could not be explained by the use of classical radio-meteorological parameters. The atmospheric stability was shown to have an effect on transmission comparable with that of the gradient of the refractive index.

621.391.812.623 265
V.H.F. and U.H.F. Signal Characteristics Observed on a Long Knife-Edge Diffraction Path—A. P. Barsis and R. S. Kirby. (*J. Res. NBS*, vol. 65D, pp. 437–448; September/October, 1961.) The long-term fading range is substantially less than expected for tropospheric scatter paths of comparable length. Transmission-loss levels agreed generally with predicted knife-edge diffraction propagation with allowance for rounding of the knife edge. A technique for estimating long-term fading ranges is presented and results agree well with observations.

621.391.812.623.029.63 266
Multipath Properties of Tropospheric Propagation of Very Short Radio Waves Beyond the Horizon—M. Hirai. (*J. Radio Res. Labs, Japan* vol. 8, pp. 147–174; May, 1961.) Theory is derived relating the arrival-angle distribution, space-diversity characteristics, antenna-beam pattern and antenna-to-medium coupling loss. Experimental data obtained at 600 and 2120 Mc agree with the theory.

621.391.812.63 267
The Properties of Low-Frequency Radio Waves Reflected from the Ionosphere, Calculated by the Phase-Integral Method—E. Cooper. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 122–141.) Calculations of equivalent height and absorption of radio echoes from model ionospheres are made. Simple ray theory is elaborated to give good results down to frequencies of about 200 kc.

621.391.812.63 268
A Study of the Upper Frequency Limits for Reflection by the Ionosphere in Europe—P. Halley, J. Duboc, and J. Thuillier. (*Ann. Télécommun.*, vol. 16, pp. 117–124; May/June, 1961.)

621.391.812.63 269
A Theory of Incoherent Scattering of Radio Waves by a Plasma: Part 2—Scattering in a Magnetic Field—D. T. Farley, J. P. Dougherty and D. W. Barron. (*Proc. Roy. Soc. (London) A*, vol. 263, pp. 238–258; September 5, 1961.)

A general expression for the frequency spectrum of radio waves scattered by the random thermal fluctuations of electron density in a plasma in a magnetic field is derived from the generalized Nyquist noise theorem used in Part 1 [2345 of 1961 (Dougherty and Farley)]. The magnetic field has a significant effect on the shape of the spectrum only for an incident radio beam nearly orthogonal to the field.

621.391.812.63 270

The Spectrum and the Total Intensity of Electromagnetic Waves Scattered from an Ionized Gas in Thermal Equilibrium in the Presence of a Static Quasi-Uniform Magnetic Field—J. Renau, H. Camnitz, and W. Flood. (*J. Geophys. Res.*, vol. 66, pp. 2703–2732; September, 1961.) The spectrum of em waves scattered from thermal electron density fluctuations has been derived with the aid of Nyquist's noise theorem applied by Dougherty and Farley (2345 of 1961) to the no-field case. General results of spectrum analysis are given and graphs are presented for radar backscatter from the exosphere.

621.391.812.63 271

Nondeviative Absorption at Night—N. Wakai. (*J. Radio Res. Labs., Japan*, vol. 8, pp. 213–218; May, 1961.) The field strengths of standard-frequency stations in Japan and South Africa were measured during voyages from Japan to the Antarctic. For oblique incidence paths, nondeviative absorption at night is large enough to be important.

621.391.812.63 272

The Effect of Small Values of $Z=\nu/\omega$ on the Complex Refractive Index in the Magneto-ionic Theory—H. Unz. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 100–104; October, 1961.) The first three terms of the Maclaurin expansion of the general Appleton-Hartree formula, with Z as the variable, are calculated. For small values of Z the refractive index varies as Z^2 while the absorption coefficient varies as Z.

621.391.812.63.029.4/.5 273

On the Analysis of L.F. Ionospheric Radio Propagation Phenomena—J. R. Johler. (*J. Res. NBS*, vol. 65D, pp. 507–529; September/October, 1961.) Recent theoretical work and extensions of the classical geometric-optical theory of propagation are used to analyze LF propagation data. The results illustrate a practical model of the ionosphere.

621.391.812.63.029.45 274

Resonance of the Space between Earth and Ionosphere—H. Poeverlein. (*J. Res. NBS*, vol. 65D, pp. 465–473; September/October, 1961.) Maximum field strength of the standing wave, at a few kc, produced in the space between the earth and the ionosphere, is derived for the resonance frequencies of this space. The earth and the ionosphere are considered as two reflectors, the lower ionosphere being only a partial reflector allowing radiation to enter and leave the resonance space. See also 3631 of 1960 (Wait).

621.391.812.63.029.45 275

Observed Attenuation Rate of E.L.F. Radio Waves—A. G. Jean, A. C. Murphy, J. R. Wait, and D. F. Wasmundt. (*J. Res. NBS*, vol. 65D, pp. 475–479; September/October, 1961.) Propagation attenuation rates for frequencies below 1 kc were calculated from the spectra of atmospherics observed at widely-spaced stations. Observed rates seem to be consistent with a two-layered ionosphere model with its lower region 90 km above the earth. See 4365 of 1960 (Wait.)

621.391.812.63.029.45 276

A Note concerning the Excitation of E.L.F. Electromagnetic Waves—J. R. Wait. (*J. Res. NBS*, vol. 65D, pp. 481–484; September/October, 1961.) "Previous solutions for the ELF mode series are discussed briefly. Particular attention is paid to the height-gain functions. The excitation of the modes for vertical and horizontal dipole excitation is also considered." See 4365 of 1960.

621.391.812.63.029.45 277

Reflection from a Sharply Bounded Ionosphere for V.L.F. Propagation Perpendicular to the Magnetic Meridian—D. D. Crombie. (*J. Res. NBS*, vol. 65D, pp. 455–463; September/October, 1961.) Formulas are derived for the reflection coefficient of a sharply bounded ionosphere for plane waves incident from the east or west at any magnetic latitude. In general the reflection coefficients for the directions east-west and west-east are not equal. See also 4004 of 1960 (Taylor).

621.391.812.63.029.45 278

V.L.F. Ionospheric Reflection Coefficients—Derivation from Impedance Concepts and Values for some Model Ionospheres—E. C. Field and P. Tamarkin. (*J. Geophys. Res.*, vol. 66, pp. 2737–2750; September, 1961.) A 16-kc vertically-polarized plane wave, propagated in a direction perpendicular to the geomagnetic meridian, is assumed to be incident obliquely upon a sharply bounded, magnetically anisotropic, absorptive ionosphere. The calculated results for eight model ionospheres (four daytime, four nighttime) are plotted. Agreement with experiment is close except in the four low-density models.

621.391.812.63.029.45:523.75 279

Prediction of V.L.F. Diurnal Phase Changes and Solar Flare Effects—W. D. Westfall. (*J. Geophys. Res.*, vol. 66, pp. 2733–2736; September, 1961.) Expressions are derived for calculating phase changes over long (>5000 km) and short paths. For long paths, results of measurements agree closely with those calculated. Further multimode experimental data are needed for short paths in the range 2000–5000 km. Expressions are also given which relate phase changes to height changes caused by solar flares.

621.391.812.63.029.62:551.510.535 280

Frequency Dependence of D-Region Scattering at V.H.F.—J. C. Blair, R. M. Davis, Jr., and R. C. Kirby. (*J. Res. NBS*, vol. 65D, pp. 417–425; September/October, 1961.) Results are given of simultaneous observations for one year at five frequencies from 30 to 108 Mc using narrow-beam antennas scaled in dimension and height according to wavelength. Hourly values of system loss closely follow the law $P_t/P_a \propto f^n$, where P_t is power to transmitter antenna, P_a is available power at receiving antenna, f is frequency. n varies with time of day and season, with 90 per cent of observations lying in the range 7–9.5.

RECEPTION

621.391.812:[550.385+523.75 281

Some Relationships between Short-Wave Fade-Outs, Magnetic Crochets and Solar Flares—Acton. (See 109.)

621.391.812.6 282

Characteristics of Fading due to Ducts and Quantitative Estimation of Fading—S. Ugai. (*Rev. Elec. Comm. Lab., Japan*, vol. 9, pp. 319–360; May/June, 1961.) A detailed study of fading at frequencies between 0.2 and 6.72 Gc over paths of 48, 97 and 147 km, and the statistical correlation with meteorological parameters.

621.391.82:621.376.3 283

F.M. Reception under Conditions of Strong Interference—J. van Slooten. (*Philips Tech. Rev.*, vol. 22, pp. 352–360; July 27, 1961.) The effects observed in FM signals with strong interference are clarified using simple theory.

621.391.822.029.64 284

The Effect of Rain and Water Vapour on Sky Noise at Centimetre Wavelengths—D. C. Hogg and R. A. Semplak. (*Bell Sys. Tech. J.*, vol. 40, pp. 1331–1348; September, 1961.) Measurements at 6 Gc show that for an antenna-beam position 5° above the horizon, the sky-noise temperature increases about 15°K from winter to summer due to humidity changes. With the antenna pointed to the zenith, temperatures of 130°K have been observed, compared with the usual background value of 3°K due to oxygen and water vapor.

STATIONS AND COMMUNICATION SYSTEMS

621.376.2/.3:621.395.665.1 285

New F.M.-A.M. Method of Compression, Expansion and Multiplication—W. R. Aiken and C. Süsskind. (PROC. IRE, vol. 49, pp. 1550–1552; October, 1961.) A method based on the combination of two signals has been developed for magnitude compression or expansion of AF or video signals, and for multiplication or other uses involving modulation with separated components. The multiplication process is suitable for analog computers.

621.376.56:621.391.82 286

Compression, Filtering, and Signal-to-Noise Ratio in a Pulse-Modulated System—E. A. Marcatili. (*Bell Sys. Tech. J.*, vol. 40, pp. 1421–1453; September, 1961.) A discussion of the transmission of Gaussian pulses in a frequency-division multiplex system, with examples of design data. See also 3145 of 1961.

621.391.1:621.376.5 287

Transmission Systems with Ideal Pulse Response—E. Kettel. (*Arch. elekt. Übertragung*, vol. 15, pp. 207–214; May, 1961.) The ideal pulse transmission system is defined; it is represented by a low-pass filter followed by an equalizer to produce the ideal pulse response. Limiting conditions and possibilities of realization of such a system are discussed.

621.391.1:621.391.827 288

Mutual Interference between Surface and Satellite Communication Systems—W. J. Hartman and M. T. Decker. (*J. Res. NBS*, vol. 65D, pp. 433–436; September/October, 1961.) Estimates of interference are presented in a form suitable for engineering applications. The systems can share the same frequencies under suitable conditions.

621.391.1:621.396.41 289

The Transmission Properties of Networks with a Periodically Operated Switch—H. Kaden. (*Frequenz*, vol. 15, pp. 141–155; May, 1961.) The problem of reducing the power loss in periodically-gated transmission networks, such as time-division multiplex systems, is considered. Steady-rate transmission can be approximated by means of a tuned circuit providing an inertia effect. A detailed theoretical treatment of such a circuit is given.

621.396.216 290

Compatible Single Sideband—L. R. Kahn. (PROC. IRE, vol. 49, pp. 1503–1527; October, 1961.) A description is given of a new type of modulated wave which can be received by existing AM receivers. This compatible SSB system (see also 971 of 1959) achieves a desired/undesired sideband ratio better than 30 db under normal modulation conditions. The main advantage of the technique is to reduce co-channel and adjacent-channel effects and to produce a signal of higher fidelity.

621.396.41:621.372.823 291
Interference Waves in Waveguide Transmission and Intermodulation Noise in F.M. System for Waveguide Communication—K. Miyauchi. (*Rev. Elec. Comm. Lab., Japan*, vol. 9, pp. 361–393; May/June, 1961.) A theoretical discussion of the transmission in a frequency-division multiplex system.

621.396.43:523.3 292
Bandwidth of a Moon Communication Circuit—J. V. Evans. (*Brit. J. Appl. Phys.*, vol. 12, pp. 406–409; August, 1961.) The measured demodulation of an AF AM signal at 162 Mc shows that reflection takes place from a small part of the moon's disk and that the AF bandwidth is adequate for speech communication but not sufficient for good-quality music.

621.396.43:551.507.362.2 293
Courier Communication Satellite—P. W. Siglin and G. Senn. (*J. IEE*, vol. 7, pp. 504–508; August, 1961.) A description of the satellite and its associated equipment and the performance of the system during its operational life, October 4–21, 1960.

621.396.73 294
A 150-Mc/s Personal Radio Signalling System—D. Mitchell and K. G. Van Wynen. (*Bell Sys. Tech. J.*, vol. 40, pp. 1239–1257; September, 1961.) A description of the system, installed in New York City in October, 1960, which comprises pocket receivers with tuned reeds and a central office for direct dialing of subscribers.

621.396.945 295
Underground Radio Communication and its Application for Use in Mine Emergencies—D. J. Vermeulen and P. J. Blignaut. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 52, pp. 94–105, Discussion, pp. 105–109; April, 1961.) Reliable communication at 335 kc can be achieved up to distances of about 2000 ft in South African gold mines. In other mining conditions, distances in excess of 800 ft may be obtained and where insulated conductors are present a considerable improvement is likely.

621.396.946 296
Communications at Megamile Ranges—R. C. Hansen and R. G. Stephenson. (*J. Brit. IRE*, vol. 22, pp. 329–344, Discussion, p. 345; October, 1961.) A survey is made of the design of systems suitable for communications over many millions of miles. A description of Pioneer V and the results obtained up to loss of contact at about 23,000,000 nautical miles are given. 45 references.

SUBSIDIARY APPARATUS

621–526 297
A Nonlinear Integral Equation from the Theory of Servomechanisms—V. E. Beneš. (*Bell Sys. Tech. J.*, vol. 40, pp. 1309–1321; September, 1961.)

621.314.58 298
Improved High-Power D.C. Converters—J. R. Nowicki. (*Electronic Eng.*, vol. 33, pp. 637–641; October, 1961.) Design and performance details of several new circuits are given.

621.316.722.1.078 299
Cathode-Coupled Cascode Stabilizer Circuit—N. Bett: S. G. Jones and R. S. McClean. (*Electronic Eng.*, vol. 33, p. 674; October 1961.) Comment on 3948 of 1961 and authors' reply.

TELEVISION AND PHOTOTELEGRAPHY

621.397.13 300
Twenty-Five Years of B.B.C. Television—E. Bishop. (*B.B.C. Engrg. Div. Monographs*, pp. 5–41; October, 1961.) Review of main engineering developments and future plans.

621.397.132 301
Modified N.T.S.C. Colour TV Signal—J. Davidse. (*Electronic Tech.*, vol. 38, pp. 388–392; November, 1961.) A practical modification of the N.T.S.C. signal is suggested to adapt it to the needs of a single-gun display system.

621.397.331.22:621.391.822 302
Analysis of Noise in the Image Orthicon—B. H. Vine. (*J. Soc. Mot. Pic. Telev. Engrs.*, vol. 70, pp. 432–435; June, 1961.) Any increase in photocathode sensitivity or mesh transmission permits a corresponding reduction in the light required for equal performance, but a large increase in the secondary emission of the target produces only a limited improvement.

621.397.331.222 303
The High-Beam-Velocity Vidicon—J. Dresner. (*RCA Rev.*, vol. 22, pp. 305–324; June, 1961.) The experimental tube described incorporates a glassy Sb_2S_3 photoconductive layer; it has a fast response and provides adequate image quality.

621.397.331.24:621.385.832 304
Measurement of Gas Pressure in Picture Tubes—Ehlbeck, Ritter, and Schütze. (See 340.)

621.397.6:621.317.799 305
An Instrument for Measuring Television Signal-to-Noise Ratio—Edwardson. (See 253.)

621.397.62 306
Optimum Band Shape for Television Intermediate-Frequency Amplifier—T. Murakami. (*RCA Rev.*, vol. 22, pp. 245–279; June, 1961.) On a linear-phase IF amplifier the optimum band shape is parabolic.

621.397.62:621.396.67 307
Receiving Antenna and Distribution Systems for Operation of Multiple Television Receivers—F. R. Lackey. (*Proc. IRE (Australia)*, vol. 22, pp. 429–441; July, 1961.)

621.397.7:551.517.362.2 308
Television Communications using Earth Satellite Vehicles—L. F. Mathews. (*J. Brit. IRE*, vol. 22, pp. 349–359; October, 1961.) Types of orbits, operational frequencies, modulation methods and repeater equipment design are discussed. Relative advantages of satellite video recordings and video tape recordings transported by air are considered.

621.397.743:621.372.2 309
Practical Experience with Single-Wire Transmission Line (Goubau Line) for the Distribution of Programs to Television Subscribers—J. Kornfeld. (*Radioschau.*, vol. 11, pp. 168–172; May, 1961.) A Goubau-line installation in an Austrian mountain valley is described, including details of the method used to couple the receivers to the line. For a description of an earlier installation see 2923 of 1960.

TRANSMISSION

621.396.72:621.396.74.029.62 310
Small Transmitters in the Swiss F.M.-V.H.F. Broadcasting Network—M. Bischhausen. (*Tech. Mitt. PTT*, vol. 39, pp. 150–154; May, 1961. In German and French.) The equipment described is designed for outdoor installation and provides an output of 6 or 60 w. The antenna system can be arranged to suit local conditions so that reception in remote areas is improved.

TUBES AND THERMIONICS

621.382.2 311
Measurement of Semiconductor Diode Transient Characteristics—D. F. Grollet. (*Mullard Tech. Commun.*, vol. 5, pp. 385–391· August, 1961.) Forward and reverse recovery times, stored charge, recombination time, and associated characteristics are defined and methods of measuring them are outlined.

621.382.23 312
Measurement of Tunnel-Diode Conductance Parameters—B. Christensen. (PROC. IRE, vol. 49, p. 1581; October, 1961.) A brief description is given of a method of determining the instantaneous incremental conductance of a tunnel diode.

621.382.23 313
On the Temperature Dependence of the Conductivity of *p-n* Junctions in the Reverse Direction—V. I. Fistul' and O. B. Orzhevskiĭ. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1158–1160; April, 1961.) Results of measurements on a number of Ge and Si diodes and Se cells are interpreted in terms of an equation derived earlier (*Fiz. Tverdogo Tela.*, vol. 2, pp. 2214–2217; September, 1960) assuming that the activation energy is decreased by the electric field.

621.382.23 314
Observations on the Cut-Off Frequency of the Tunnel Diode—M. Müller. (*Nachrtech. Z.*, vol. 14, pp. 165–168; April, 1961.) A qualitative and quantitative assessment of the factors setting the upper frequency limit in tunnel diodes. Optimum conditions are determined; it seems unlikely that a cutoff frequency of 10 Gc could be reached.

621.382.23 315
***p-n* Junctions at Very Low Temperatures**—A. K. Jonscher. (*Brit. J. Appl. Phys.*, vol. 12, pp. 363–371; August, 1961.) Properties of Si and Ge at liquid-He temperatures are reviewed and experimental results are given. The characteristics of Ge devices are similar to those at room temperature provided that ohmic contacts are adequate. Si diodes show anomalies.

621.382.23 316
Electronic Processes and Excess Currents in Gold-Doped Narrow Silicon Junctions—C. T. Sah. (*Phys. Rev.*, vol. 123, pp. 1594–1612; September 1, 1961.) Large amounts of excess current in Au-doped Si tunnel junctions are observed and interpreted as due to transition processes with the two Au energy levels in the forbidden gas of Si as intermediate states. Experimental results are compared with predictions involving Hall-Shockley-Read steps and electron tunneling between the states.

621.382.23 317
Electronic Determination of the *I*, *G* and *I/G* Parameters of a Tunnel Diode—C. R. Gneiting. (*Rev. Sci. Instr.*, vol. 32, pp. 907–908; August, 1961.) A method of obtaining *X-Y* recordings of the I/V, G/V and $(I/G)/V$ curves is described.

621.382.233:621.318.57 318
Silicon *p-n-p-n* Switch: Part 1—A. Bardsley and J. Walker. (*Electronic Tech.*, vol. 38, pp. 406–410; November, 1961.)

621.382.233:621.373.44 319
Silicon Four-Layer Devices as High-Power Pulse Generators—R. Thompson: R. P. F. Lauder. (*Electronic Eng.*, vol. 33, p. 674; October, 1961.) A comment on 3573 of 1961 and author's reply.

621.382.3:621.391.822 320
The Influence of Nonuniform Base Width on the Noise of Transistors—A. Baelde. (*Philips Res. Repts.*, vol. 16, pp. 225–236; June, 1961.) The discrepancy between experimentally observed and predicted values of the noise resistance is due to the fact that a region

of smaller base width will carry a greater than average share of the total current. This results in a greater contribution of this region to the noise resistance of the transistor.

621.382.333 321
The Base Resistance of the Transistor in the Saturated Region—O. Nakahara. (*J. Phys. Soc. Japan*, vol. 16, pp. 1260–1261; June, 1961.) Base resistance is heavily modulated by the collector voltage in the saturation region because of conductivity modulation induced by minority-carrier injection from the collector junction.

621.382.333 322
Nonlinear Behaviour of Junction Transistors at Higher Frequencies—R. Paul. (*Nachrtech.*, vol. 11, pp. 213–220; May, 1961.) Large-signal conditions are considered.

621.383.032.217.2 323
Spectral Response of Antimony-Caesium Photocathodes—J. Birks and I. H. Munro. (*Brit. J. Appl. Phys.*, vol. 12, pp. 519–522; September, 1961.)

621.383.292 324
Electron-Optical Conditions at Venetian-Blind-Type Dynodes and their Effects in Photomultipliers—P. Görlich, A. Krohs, and H. J. Pohl. (*Brit. J. Appl. Phys.*, vol. 12, pp. 525–526; September, 1961.) The suggestion of Jennings and Misso (964 of 1957) for the improvement of focusing by correct design of the slat structure is investigated.

621.385.032.213.23 325
Emission of Oxide Cathodes during Activation and Operating Life of Thermionic Valves—F. Dlouhy and H. J. Schütze. (*Telefunken-Röhre*, no. 39, pp. 121–157; May, 1961.) The emission quality of oxide-coated cathodes is assessed on the basis of shot-noise tests. 32 references.

621.385.13:531.788.7 326
Determination of Pressure in Thermionic Valves whose Ion Collector Emits Thermal Electrons—H. J. Schütze and H. W. Ehlbeck. (*Telefunken-Röhre*, no. 39, pp. 61–78; May, 1961.) The lower limit of internal pressure measurable with a tube operated as ionization gauge (see 327 below) rises under high power loads because of thermionic emission of the ion collector. A method of compensating for this emission, and its application to tube tests are described.

621.385.13:531.788.7 327
Measurement of Residual Gas Pressures in Indirectly Heated Thermionic Valves—H. J. Schütze. (*Telefunken-Rohre*, no. 39, pp. 5–60; May, 1961.) The gas pressure can be measured by operating the tube as an ionization gauge. Experimental results are given and compared with those determined theoretically. 69 references.

621.385.13.032.94 328
Degassing in Thermionic Valves during Getter-Operation—H. J. Schütze and H. W. Ehlbeck. (*Telefunken-Röhre*, no. 39, pp. 91–120; May, 1961.) The effects on residual gas pressure of various pumping procedures and getter heat treatments are investigated.

621.385.3.029.64:621.391.822 329
The Measurement of the Noise Quantities of the EC 56 at 1400 Mc/s—P.A.H. Hart and F. L. van der Vinne. (*Philips Res. Repts.*, vol. 16, pp. 237–244; June, 1961.) A method of measuring four characteristic noise quantities of a microwave triode is described. The results obtained are used to determine the minimum noise factor of a traveling-wave tube with the same type of beam as the triode. Comparison is made with results of measurements at 4 Gc and with theory [1682 of 1961 (Vlaardingerbroek)].

621.385.6:537.533 330
Quasi-Brillouin Electron Streams—M. H. Miller. (*J. Appl. Phys.*, vol. 32, pp. 1791–1793; September, 1961.) The assumption of laminar Brillouin flow is discarded in the electron-stream model discussed. This helps to explain observed discrepancies; other significant results are noted.

621.385.6:621.375.9:621.372.44 331
Generalized Coupled Mode Theory—M. C. Pease. (*J. Appl. Phys.*, vol. 32, pp. 1736–1743; September, 1961.) The pair-wise coupling of modes in distributed systems, such as the traveling-wave parametric amplifier and other electron-beam devices, is shown to permit only two types of interaction.

621.385.6:621.384.622.1 332
Fields in Cavity-Excited Accelerators—E. G. Cristal and J. Van Bladel. (*J. Appl. Phys.*, vol. 32, pp. 1715–1724; September, 1961.) Field configuration and resonance frequency are determined for the lowest mode of a coaxial cavity surrounding a circular tube. The calculation is repeated for a parallel-plane configuration which gives similar results.

621.385.623.5:621.318.57 333
Reflex Klystron as a High-Speed Microwave Switch—B. G. Whitford. (*Rev. Sci. Instr.*, vol. 32, pp. 919–921; August, 1961.) At *X*-band frequencies flat-topped pulses were produced with rise time of 1.5 nsec and decay time of 6 nsec. A ratio of "on"/"off" attenuation of 30 db was obtained with a reflector voltage swing of 10 v.

621.385.624:621.316.726 334
Arrangement for the Correction of Frequency Drifts in the Microwave Range—J. Windel. (*Nachrtech.*, vol. 11, pp. 229–231; May, 1961.) A frequency-stabilizing system for klystrons is described in which the reflector voltage is regulated by a discriminator circuit consisting of a diaphragm filter or resonator.

621.385.63 335
Suppression and Limiting of Undesired Signals in Travelling-Wave-Tube Amplifiers—H. J. Wolkstein. (*RCA Rev.*, vol. 22, pp. 280–291; June, 1961.) Saturation by RF signal may cause a decrease in the suppression of a simultaneous low-level signal. Saturation characteristics in the presence of overdriving signals are considered.

621.385.63 336
Effect of Distributed-Loss Noise Generators on Travelling-Wave-Tube Noise Factor—S. Bloom. (*RCA Rev.*, vol. 22, pp. 347–349; June, 1961.) The contribution of Nyquist noise generators to noise factor is about 0.6 db and cannot be neglected in tubes of modern design.

621.385.633 337
A Resonant-Type Backward-Wave Oscillator for the Frequency Range 30–38 Gc/s—F. Gross. (*Arch. elekt. Übertragung*, vol. 15, pp. 227–233; May, 1961.) The principle of operation and the design of a resonant-type backward-wave oscillator [see also 3596 of 1961 (Heynisch)] are described. Experimental results are given of an oscillator capable of producing a CW output of over 200 mw with an efficiency up to 1.5 per cent.

621.385.633.1 338
A Backward-Wave Oscillator for the Frequency Range 10–20 Gc/s with a Delay Line consisting of Annular Fingers—W. Wendrich. (*Nachrtech. Z.*, vol. 14, pp. 203–211; April, 1961.) The system described provides a stronger stray field for better beam interaction. The delay line is terminated by a short but effective field absorber of graphite.

621.385.832 339
On the Geometric Transformations at the Foundation of all Cathode-Ray Tube Design: Part 1—Transformations which are Conformal with respect to Space Charge—H. Moss. (*J. Brit. IRE*, vol. 22, pp. 313–327; October, 1961.)

621.385.832:621.397.331.24 340
Measurement of Gas Pressure in Picture Tubes—H. W. Ehlbeck, J. Ritter, and H. J. Schütze. (*Telefunken-Röhre*, no. 39, pp. 79–90; May, 1961.) Pressure measurements during processing and the lifetime of picture tubes are carried out by using the electrode system as ionization gauge. See also 327 above.

621.385.832:681.142 341
The Barrier-Grid Storage Tube—D. J. Gibbons. (*Electronic Eng.*, vol. 33, pp. 603–636; October, 1961.) The principle of operation and practical applications are discussed.

621.387 342
Space-Charge-Limited Current Relation in High-Pressure Gas Diodes—R. Forman. (*Phys. Rev.*, vol. 123, pp. 1537–1541; September 1, 1961.) For applied voltage V and gas pressure p, the current is shown theoretically to vary as $V^{3/2}$, instead of V^2 as generally assumed, and as $p^{-1/2}$. Corroborative experimental data are presented.

MISCELLANEOUS

621.395/.396].002.2 343
Progress in the Manufacture and Testing of Telecommunication Equipment—G. Weinmann. (*Elektrotech. Z.*, vol. 13, pp. 425–429; August 7, 1961.) Measures taken to increase the quality and reliability of components are discussed, and the prediction of equipment life is considered.

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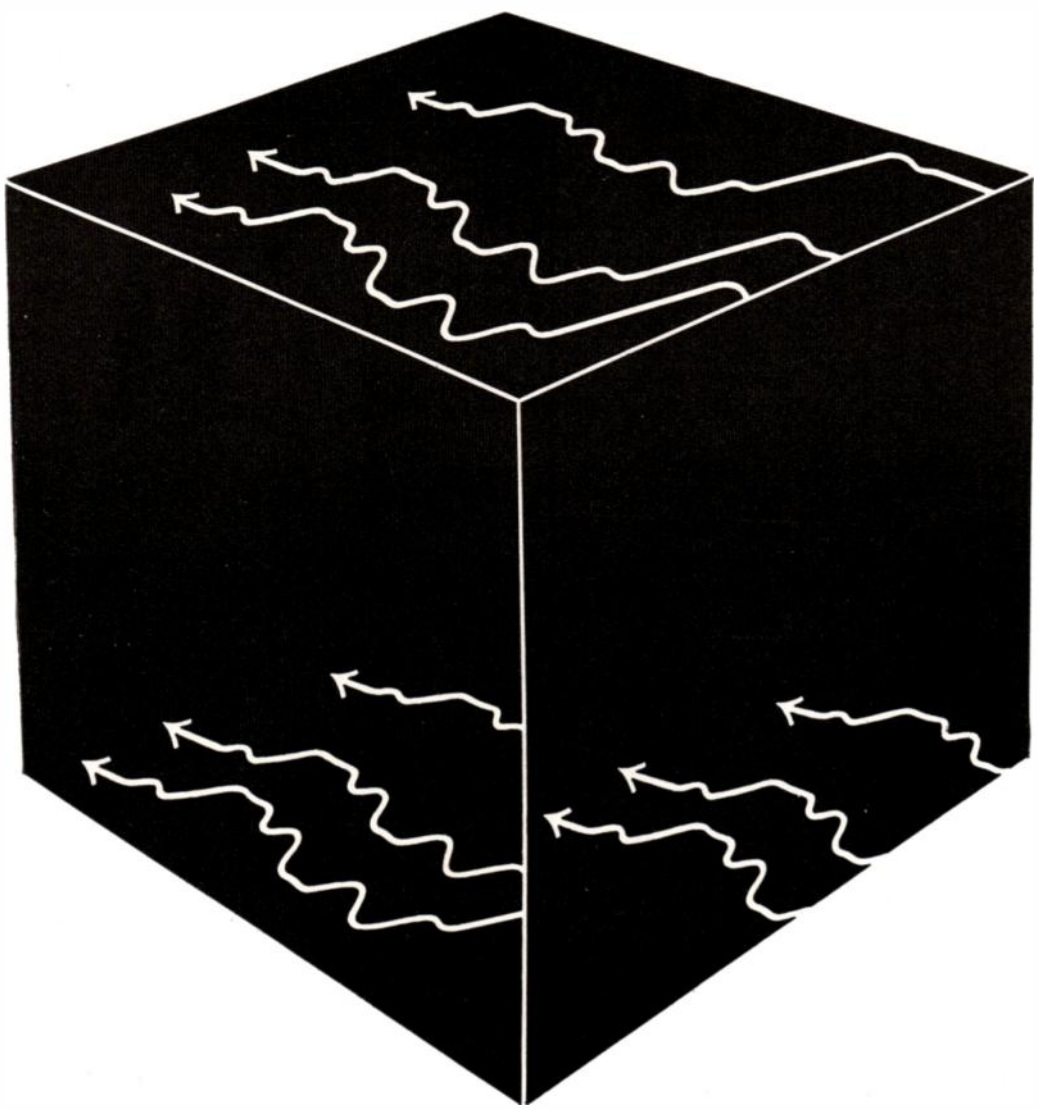
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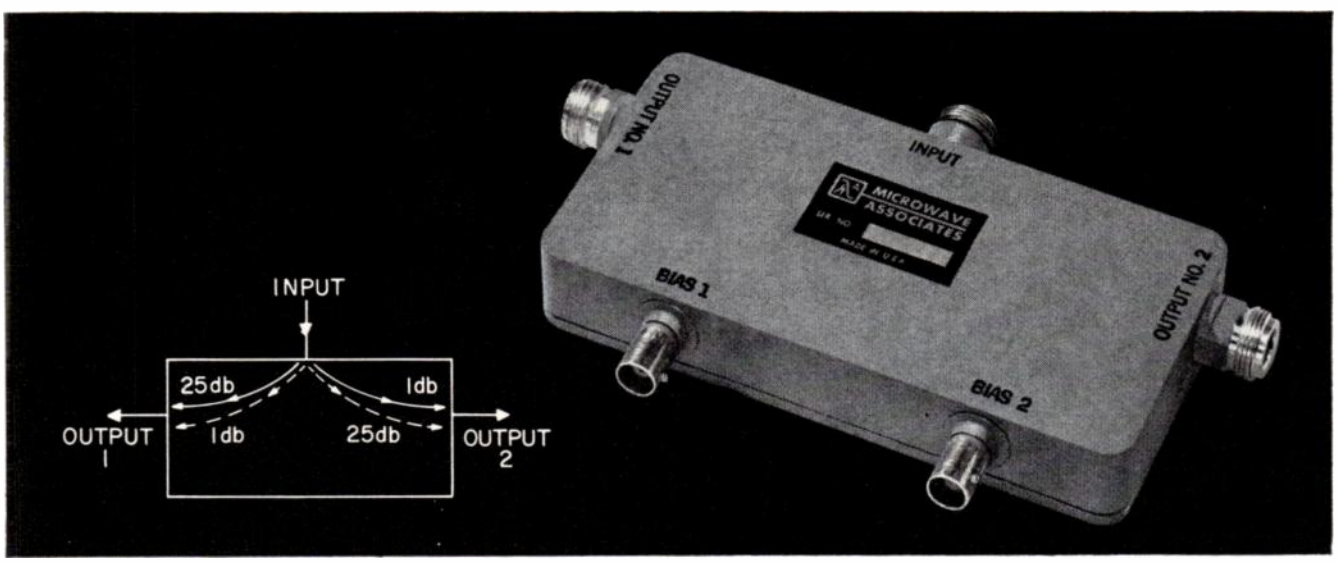
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7. Power source 115 v, 60 cps. . .	approximately 6 watts
8. Connectors	type C

Professional Group Meetings

(Continued from page 70A)

Washington, D. C.—November 7

"How Smart Can You Be—Is Electronic Sophistication Debasing Our Defense Posture?" T. Clark, Westinghouse Electric Corp.

Nuclear Science

Albuquerque-Los Alamos—October 18

"The Mass Spectrometer and Its Application to Nuclear Science," J. W. Guthrie, Sandia Corp., Albuquerque, N. M.

Dayton—October 20

"Solid-State Radiation Effects," Dr. B. Kulp, USAF, Aero Research, Wright-Patterson AFB.

Product Engineering and Production

Boston—November 21

"Audio—Visual Concepts in Production," J. Perry, Litton Systems Inc., Applied Comm. Systems Div.

Dayton—March 9

"Gamma Radiation Effects," V. J. DiRito, Wright-Patterson AFB.

Radio Frequency Interference

Fort Worth—November 14

"Selective Signaling," D. W. Land, Motorola Comm. and Elec., Inc.

Los Angeles—October 19

"Purposes and Philosophy of the PGRFI," H. L. Malone, Genistron Inc., Los Angeles.

"Systems Approach to Interference Control," K. R. Schuette, Filtron Inc., Culver City.

San Francisco—July 18

"Microwave Interference Problems and Solutions," R. H. Stone, GE Microwave, Palo Alto.

Washington, D. C.—October 17

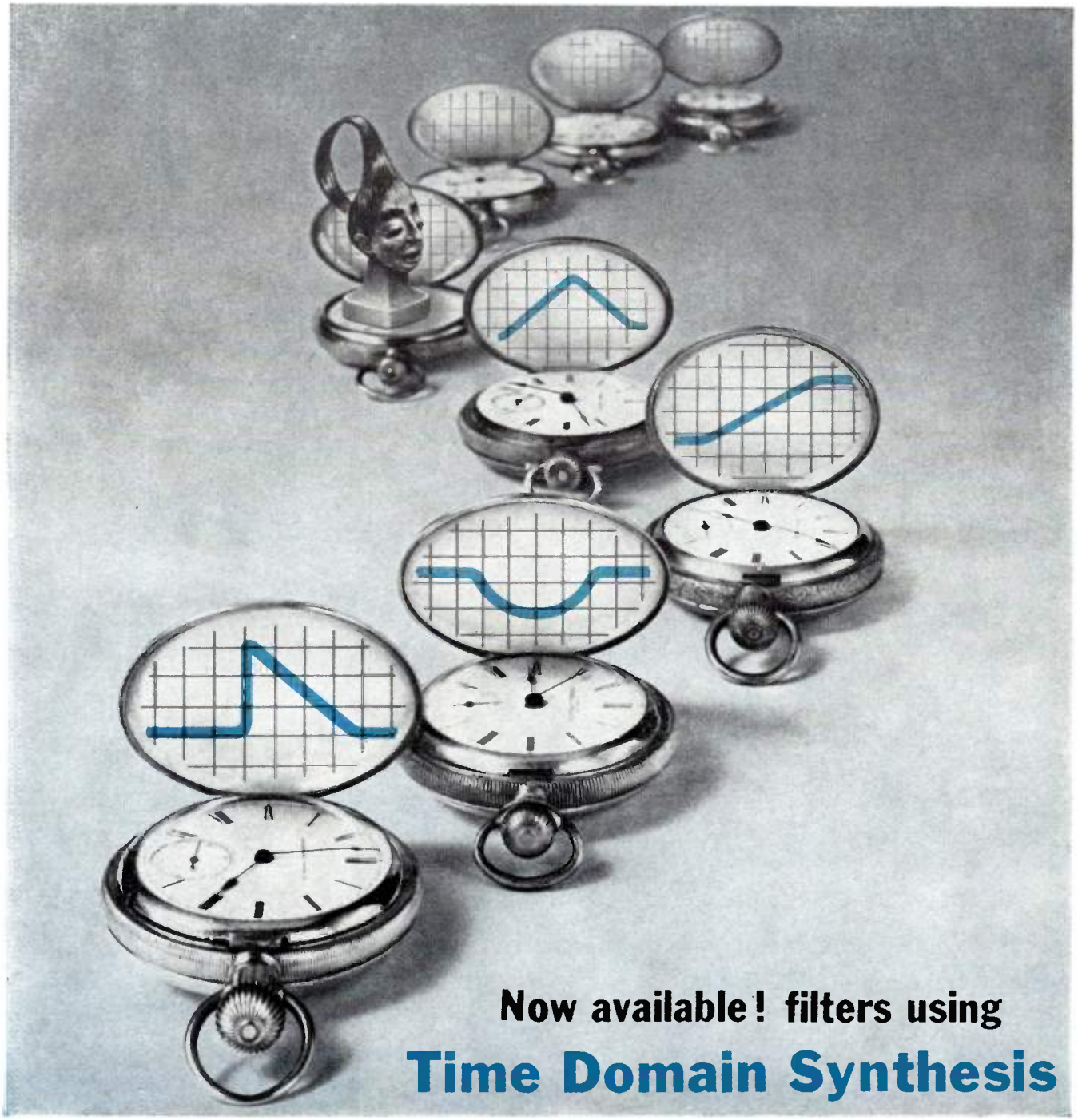
"Notes on Spectrum Signature Measurements," A. H. Sullivan Jr. and J. J. Dozier, Frederick Research Corp., Wheaton, Md. and J. W. Savage, Jansky and Bailey, Washington, D. C.

Reliability and Quality Control

Boston—October 19

"A Matrix Explanation of Reliability Management," A. A. Rothstein, Avco Corp.

(Continued on page 78A)



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Professional Group Meetings

(Continued from page 76A)

"Management of a Reliability Program," R. Jacobs, Sylvania Corp.

Discussion of Chapter's participation in NEREM.

Discussion of future Chapter programs.

Chicago—November 16

"The AGREE Report," R. Herst and M. Friedman, Motorola.

Chicago—October 18

"Maintainability Prediction," E. F. O'Connell, Federal Electric Co.

Chicago—September 21

"Improved Reliability of Silicon Diodes Through Burn-In Technology," D. Cowan, Continental Devices Corp.

Philadelphia—November 28

"Auditing Vendors: Reliability and W. C.—A Simplified Approach," T. K. Brown, Westinghouse Electric Corp., Baltimore.

"The Vendor Survey," H. B. Reed, Philco Corp., Philadelphia.

Washington D. C.—October 25–26

Symposium: "Conference on Reliability Assurance Techniques for Semiconductor Specifications."

Space Electronics and Telemetry

Orlando—September 27

"Optimum Code Patterns for PCM Synchronization," M. W. Willard, Dynatronics, Inc., Orlando, Fla.

San Francisco—November 21

"The State of the Art in High-Power Ground-to-Space Transmitters," E. R. Altschul, Radiation at Stanford, Palo Alto, Calif.

Vehicular Communications

Chicago—November 10

"Personal Signalling," R. E. Sefton, Illinois Bell Telephone Co.

Los Angeles—November 16

"Using Diplexing Antenna Techniques for Reducing Intermodulation Distortion," T. J. McMullin, Decibel Products Inc., Dallas, Tex.

Los Angeles—October 19

"The Motrac Extender, History, Operation and Use," E. Scott, Motorola Communications and Electronics Inc., No. Hollywood.

Los Angeles—September 21

"The Message Master," R. M. McIntosh, Hallett Mfg. Co., Los Angeles.

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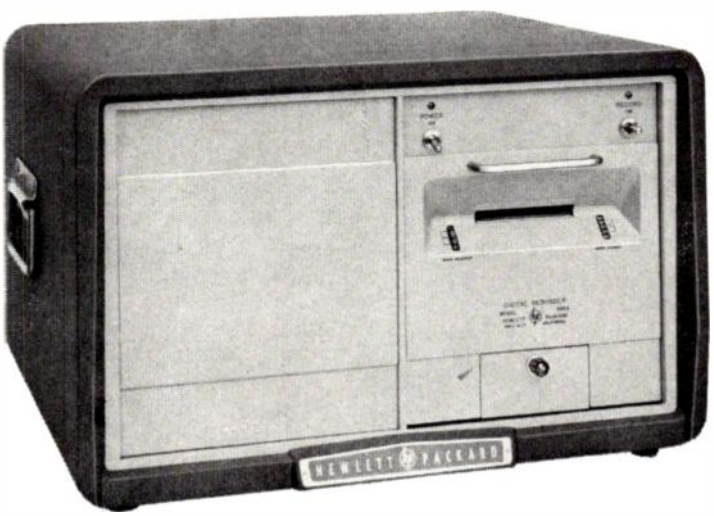
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SPECIFICATIONS

Printing Rate: 5 lines/sec. maximum

Column Capacity: To 11 columns (12 available on special order)

Print Wheels: 12-position, 0 through 9, a minus and a blank (Many special character wheels available from stock)

Driving Source: Parallel-entry 4-line BCD, 1-2-2-4. Other codes available on plug-in cards. Source reference voltages establish "0" and "1" states, which may be as much as 100 v above or below ground. "1" state 4 to 75 v above "0" reference. Driving power approx. 30 μa into 270,000 ohms.

Print Command: ± pulse, 20 μsec or greater in width, 6 to 20 v.

Hold Signal: (Available for each data source) —7 v to + 15 v and + 15 v to —7 v.

Transfer Time: 2 msec

Paper Required: Standard 3" roll or folded

Line Spacing: Single or double, adjustable

Size: Cabinet, 20¾" x 12½" x 18½"; Rack, 19" x 10½" x 16⅞" deep behind panel.

Price: hp 562A (cabinet) or hp 562AR (rack mount) $1,600.00 to $2,200.00 depending on options. hp 580A Digital-Analog Converter, price on request.

New, solid state hp 562A Digital Recorder prints digital data on 3" paper as fast as 5 lines per second, each line containing up to 12 digits. The instrument incorporates a unique data storage unit for each digit column that allows the data source to transfer data to the recorder in just 2 milliseconds, after which the source is free to collect new data.

Besides the standard parallel-entry 4-line BCD code (1-2-2-4), you can easily use other 4-line codes just by substituting plug-in column cards. Ten-line code operation (without data storage feature) is also available with plug-in cards.

Further, hp 562A accepts dual input (optional) and prints data simultaneously from two unsynchronized sources. A "patch panel" permits programming these two separate, unsynchronous inputs (even if coded differently) in any manner. Combinations of plug-in column code cards and "patch panel" column programming give complete flexibility in both dual-source data acquisition and data print positioning.

Analog output for high-resolution strip chart and X-Y recording is available as an extra-cost built-in feature of the 562A or through the new hp 580A Digital-Analog Converter, a separate solid state, high-precision instrument.

Designed for use with solid state and vacuum tube counters, Model 562A is ideal for a wide variety of individual and system applications. Call your hp representative today.

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World War II and the advent of radar gave a tremendous boost to this upward climb through the frequency domain, so that today the microwave field has blossomed to a position of prominence—almost predominance—in the radio engineering art.

On March 7, 1952, this thriving field of activity received another important boost when the IRE formed a Professional Group on Microwave Theory and Techniques. For the first time, this important branch of the field was provided with its own organization for channeling and spreading vital, specialized knowledge to its own members and for stimulating a planned program of service tailored to the needs of the field. The value of the services being performed by the Group is attested to by the fact that more than 5500 engineers have joined and paid the modest $3 assessment fee.

The most important service provided by the Group is its TRANSACTIONS, which is issued bi-monthly to all Group members. Its pages carry over 100 articles and letters annually, providing microwave engineers an unparalleled opportunity to keep abreast of the rapid progress of this swiftly moving field.

The Group has also been very active in sponsoring conferences throughout the year. In addition, local meetings are held by Chapters of the Group in 18 cities in the United States and in Tokyo.

If the activities of the PGMTT are not history-making they are assuredly history-changing, for already these activities have altered the course and speed of progress in this field.

Ernst Weber

Chairman, Professional Groups Committee

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Fox, E. D., Basking Ridge, N. J.
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(Continued on page 82A)



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SM 36-5M	0–36	0–5	SM 36-15MX
SM 75-2M	0–75	0–2	SM 75-2MX
SM 160-1M	0–160	0–1	SM 160-1MX
SM 325-0.5M	0–325	0–0.5	SM 325-0.5MX

5¼" PANEL HEIGHT

SM 14-15M	0–14	0–15	SM 14-15MX
SM 36-10M	0–36	0–10	SM 36-10MX
SM 75-5M	0–75	0–5	SM 75-5MX
SM 160-2M	0–160	0–2	SM 160-2MX
SM 325-1M	0–325	0–1	SM 325-1MX

8¾" PANEL HEIGHT

SM 14-30M	0–14	0–30	SM 14-30MX
SM 36-15M	0–36	0–15	SM 36-5MX
SM 75-8M	0–75	0–8	SM 75-8MX
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0.32 ampere 10 H — 100 H.
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Type 2030 Portable Megohmmeter — battery operated, 500 volt test potential. Range 1 megohm to 10 million megohms.

Send for NEW 48 page transformer catalog. Also ask for complete laboratory test instrument catalog.

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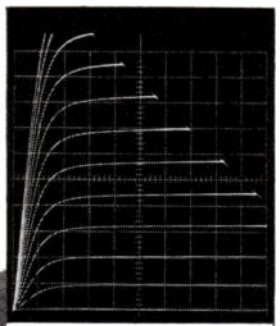
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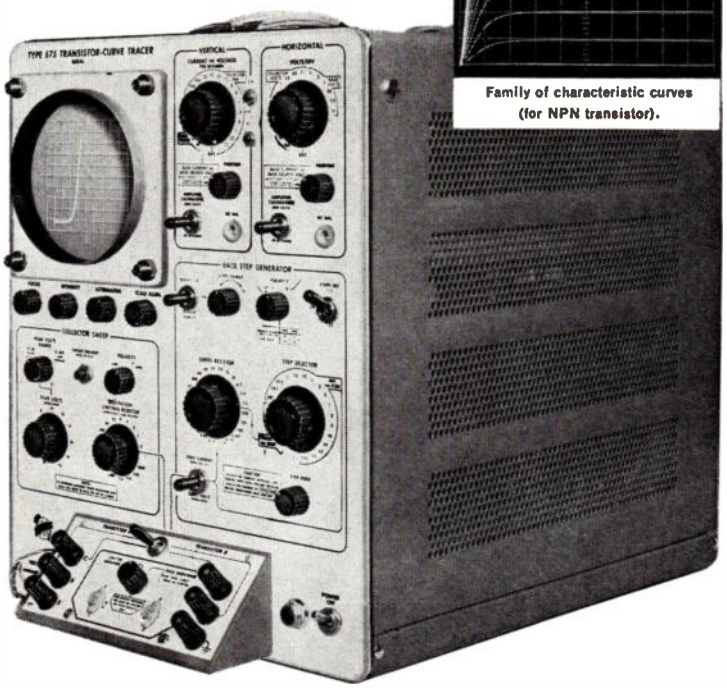
(Continued on page 84A)

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INVALUABLE TOOL FOR EVALUATING SEMICONDUCTOR DEVICES



Family of characteristic curves (for NPN transistor).



The Type 575 provides 20-ampere collector displays (10-ampere average supply current), two ranges of collector supply (0 to 20 volts, 0 to 200 volts), and 2.4-ampere base supply (positive or negative base stepping).

With a Type 575, you can plot and measure 7 different transistor characteristics. You can display 4 to 12 curves per family—with input current from 1 microampere/step to 200 milliamperes/step or input voltage from 10 millivolts/step to 200 millivolts/step—in repetitive or single-family presentations. You can select either common-emitter or common-base configurations.

Add a Type 175 Adapter and you extend the range of collector displays 10 times and the range of base supply 5 times.

You can also test diodes under a wide variety of conditions and observe waveform characteristics on the 5-inch crt with a high degree of accuracy.

Type 575 Calibrated Displays

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Horizontal Axis—Collector Voltage, 11 steps from 0.01 v/div to 20 v/div.

Both Axes—Base Voltage, 6 steps from 0.01 v/div to 0.5 v/div. Base Current, 17 steps from 0.001 ma/div to 200 ma/div. Base Source Voltage, 5 steps from 0.01 v/div to 0.2 v/div.

Type 575 Transistor-Curve Tracer . . . $1075

U.S. Sales Prices f. o. b. Beaverton, Oregon

HIGH-CURRENT ADAPTER

For measuring high-powered semiconductor devices which exceed the current capabilities of a Type 575, ask your Tektronix Field Engineer about the Type 175 High-Current Adapter. Not intended for separate use, the Type 175 depends upon the circuitry and crt of a Type 575 to provide *200-ampere collector displays, three ranges of collector supply, and 12-ampere base supply*—for calibrated displays with Collector Current on the Vertical Axis and either Collector Voltage or Base Voltage on the Horizontal Axis.

Type 175 Transistor-Curve Tracer High-Current Adapter $1475



HIGH-VOLTAGE TYPE 575

Supplied on order from your Tektronix Field Engineer is a special model of the Type 575 Transistor Curve Tracer. Although similar to the Type 575, the special model provides much higher diode breakdown test voltage (*variable from zero to 1500 volts* at a maximum current of 1 milliampere) and also much higher Collector Supply (*up to 400 volts*, at 0.5 ampere).

For complete specifications of this special model—call your Tektronix Field Engineer.

Type 575 Mod 122C $1275

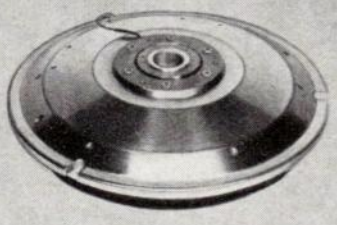
. . . for more information about evaluating semiconductor devices with a Type 575 or other Tektronix test equipment, call your Tektronix Field Engineer. He will be glad to assist you.

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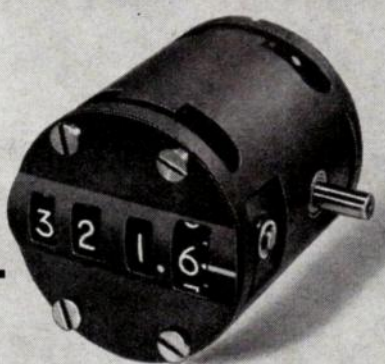
SPECIFICATIONS

Typical Part Numbers	Function	Excitation	Accuracy
3Z6390-001	Resolver Transmitter	20v 900 cps	3 min.
CZ06360-002	Resolver Transmitter	115v 800 cps	20 sec.
5Z6360-009	Resolver	115v 800 cps	3 or 5 min.
3Z6210-006	Synchro	As req'd	2 min.
1/3Z6360-006	Resolver	115v 800 cps	20 sec.

For complete data write Kearfott Division, General Precision, Inc., Little Falls, New Jersey.

KEARFOTT

KEARFOTT NAVIGATIONAL COUNTERS



Direct visual counter display with large numerals enables lower valued increments of angles to be easily read in terms of minutes and seconds of arc. Precision design, rugged construction, durable materials. With odometer or Geneva drives. Temperature range −65°C to +90°C; meet military requirements for shock, vibration, case size.

CHARACTERISTICS

Part Number	Maximum Count	Counts Per Rev.	Maximum Speed RPM	Break-Away Torque IN. OZ.
C160432-( )	179° 59′	20′	1200	.75
C160431-( )	179° 59.9′	2′	1200	1.0
C160435-( )	179.9°	2°	1200	.75
C160433-( )	99.9	2.0°	1200	.75
C160434-( )	999.9	2.0°	1200	.75
MK 3 MOD 1	359.	2.0°	1800	.35
C160407004	359.9	2.0°	1800	.35

For complete data write Kearfott Division, General Precision, Inc., Little Falls, New Jersey.

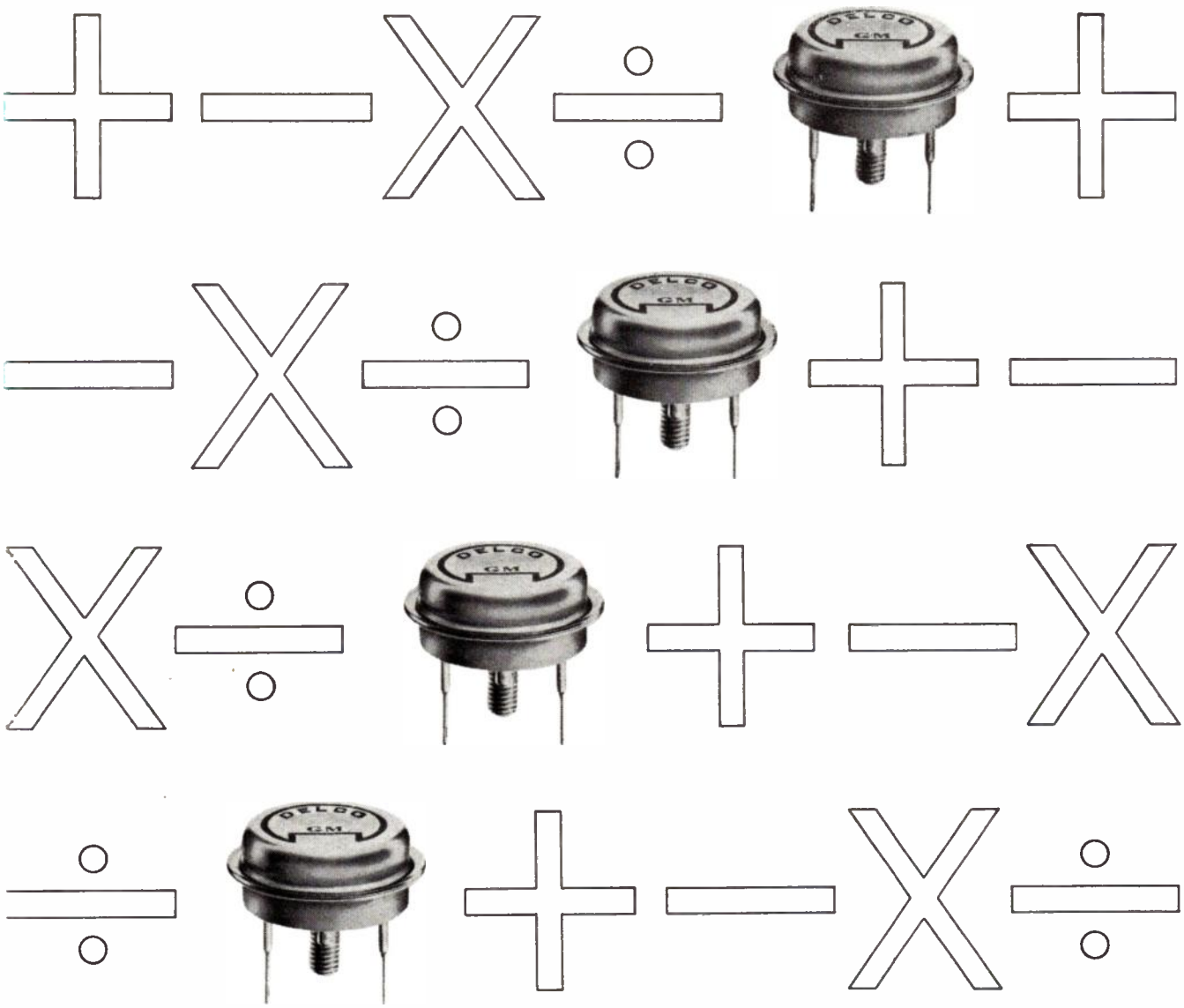
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(Continued on page 86A)



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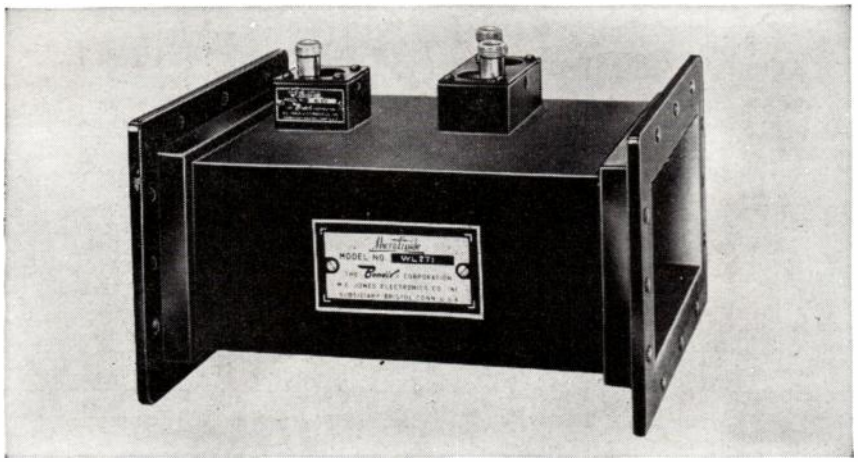
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The model WL271, illustrated, is an example of a standard model modified to meet a specific customer requirement: L Band; 1100-1700 MCs.; 2RF sampling probes 30 and 72 db below main line Incident Power, and 1 probe 53 db below main line Reflected Power; directivity 35 db minimum; 150 KW average; 30 megawatts peak power. *All this in a package 1/10th the size of a conventional waveguide coupler.*

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(Continued on page 88A)

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PocketPot*

SPECIFICATIONS

ACCURACY: ± .05% of reading or ± .5 mv., whichever is greater.

RANGES: 0-5. 100 v. When used with PC-S 0—500 v. and 0—1 amp.

SENSITIVITY: Infinite resistance at null. When used with PC-S, 2,000 Ω/v.

RESOLUTION: Continuous, 1 mv. divisions on slide wire.

BATTERY OPERATED: Does not require an external power source.

READOUT: Direct "in line."

SIZE: 9" x 4¼" x 1¾". **WT:** 3 lbs.

PRICE: Model PC, $450.00; Model PC-S $175.00. F.O.B., New Rochelle, N.Y.

PocketPot*

Write for additional information or contact your nearest SRIC sales representative.

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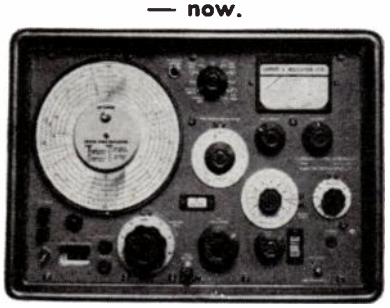
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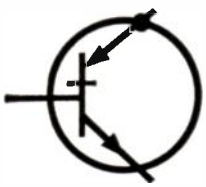
to those engineers and scientists who will be attending the IRE International Convention in New York during March 26-29. Products displayed by Litton's various operations will include microwave tubes and display devices, microwave ferrite components and hardware, computer components, audio-visual teaching machines, and a variety of other advanced electronic systems and hardware. *Booths 1608-16 and 1709-17.*

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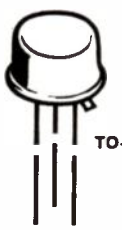
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Kilowatt Switching Capability—100 Nanoseconds

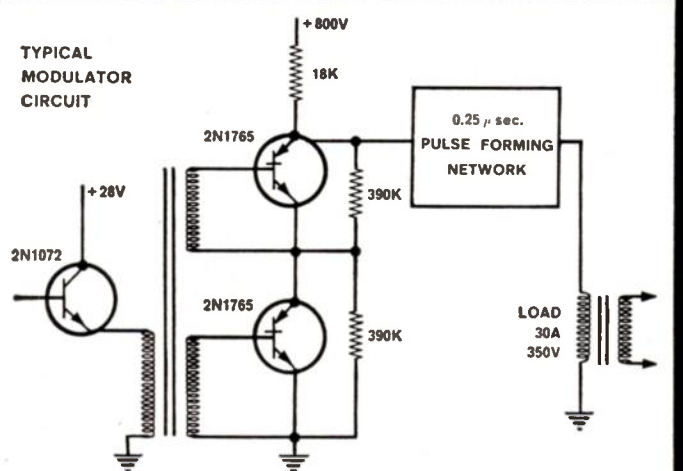


2N1765



TO-5 PACKAGE

The PNPN 2N1765 is a 3-junction, 3-terminal diffused silicon transistor in a vacuum-tight enclosure suitable for fast, high-current, pulse switching. This device's rise time to 30 amperes is less than 100 nanoseconds. Turn off time is approximately 1 microsecond. It is operational at emitter currents to 50a with gate drives from 5ma to 2a.



MAXIMUM RATINGS AT 25° C

$I_{(on)}$	50a
P (avg)	500mW

ELECTRICAL CHARACTERISTICS

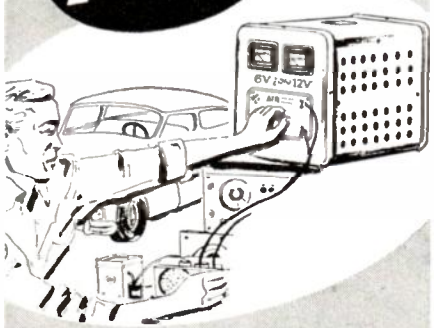
$BV_{(off)}$ (min)	400V dc
V_F (max) (@ 200 nsec)	50V @ 30a
$t_d + t_r$ (max) (@ i_F=50a)	125 nsec
I_H (min)	5mA

The 2N1765 transistor may be purchased in quantity from Western Electric's Laureldale Plant. For technical information, price, and delivery, please address your request to Sales Department, Room 106, Western Electric Company, Incorporated, Laureldale Plant, Laureldale, Pa. Telephone—Area Code 215—WAlker 9-9411.

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TYPE	INPUT A.C. Volts 60 Cycles	D.C. OUTPUT VOLTS	D.C. OUTPUT AMPERES Cont.	D.C. OUTPUT AMPERES Int.	SHIP. WT.	USER PRICE
610C-ELIF	110	6 -or- 12	10 6	20 12	22	$49.95
620C-ELIT	110	6 -or- 12	20 10	40 20	33	66.95

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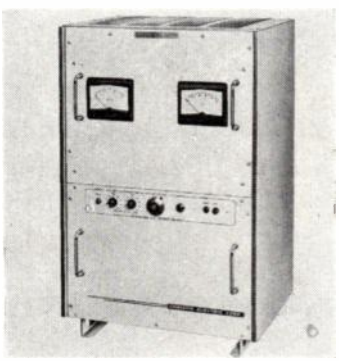
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 39A)

Precision DC Power Supplies

Christie Electric Corp., 3410 W. 67th St., Los Angeles 43, Calif., has just introduced a line of "Transient-Free" high current power supplies. The units are rated 100, 200, 250, 400 and 600 amperes, respectively, with the dc voltage adjustable up to 36 volts. The following typical specifications are achieved through use of advanced magnetic amplifier-transistor circuitry: Dynamic Regulation—0.5 volts; Voltage Ripple—1 millivolt rms; Recovery Time—50 microseconds; and Output Impedance is 1 milliohm.



With these power supplies the over and undershoot is negligible even at the end of a long cable, and transistorized loads are fully protected.

Other features include continuously adjustable current limiting; dual input voltage; remote sensing; simplified servicing through use of plug-in module assemblies; parallel or series operation; ultra-fast overvoltage and over-current protection. Several cabinet styles are available such as the Lab Cabinet (pictured), which is easily converted into 19" Rack, and two mobile models. Militarized units are also available. For details write to the firm.

Low Weight Antenna Mount

Weighing less than 1,000 lbs., a new Antenna Mount Model 28 produced by **Temec, Inc.**, 7833 Haskell Ave., Van Nuys, Calif., is a servo controlled pedestal for shipboard and ground based applications. It is designed for use in telemetry reception and the tracking of aircraft missiles and satellites.

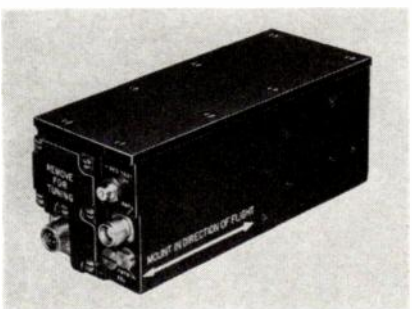


Special features include gyro stabilization for shipboard use, a complete solid state AC servo system and automatic beam crossover switching from 3 db to 9 db. Current production model employs a 5 ft. reflector for use in the 5 kmc range, but other reflectors and frequencies are available on order. The current system has slew rates of 36°/second, and accelerations of 130°/second/second.

Remote stowing is accomplished by electromechanical brakes located on the pedestal. One-quarter hp servo motors are used for this purpose but current designs permit 1 hp motors to be employed, if desired.

An azimuth rotation of 710°, with hydraulic stops, is currently available, but slip rings may be provided to facilitate continuous rotation, if desired. Dynamic braking is provided automatically to protect the mechanical structure from shock loading should a mechanical limit be reached.

C-Band Radar Beacon



The RB-200 is a lightweight, solid-state subminiature C-Band Beacon designed by **General Instrument Corp.**, Defense and Engineering Products Group, Andrews Rd., Hicksville, L. I., N. Y., for operation as a tracking device to assist radar operators to follow and identify "in-flight" missiles, drones, aircraft or space probes at distances greatly exceeding the normal reflective path of a ground radar.

The beacon consists of a single package measuring $2\frac{1}{2}'' \times 2\frac{1}{2}'' \times 7''$ and weighing less than four pounds. The unit contains a three-cavity preselector which incorporates an integral duplexer and video detector, a broadband video amplifier, decoder, modulator and power supply. The total volume of the unit is 44 cubic inches. Silicon solid state active elements are utilized throughout with the exception of a planar triode transmitting tube.

The principal electrical sections of the beacon include a transmitter capable of emitting a power output well in excess of 10 watts, and a receiver which has a sensitivity which exceeds −45 dbm. The beacon is capable of operation over the frequency range of 5400–5900 mc. Beacon delay variation with signal level is extremely low thus permitting very accurate range tracking.

Power for the beacon may be switched from an external power source to an internal 6.5–7.5 volt battery by a self-contained relay. The operational life of the internal battery exceeds 30 minutes.

The unit is constructed to withstand the extreme environmental conditions encountered during air and space operation.

(Continued on page 118A)

NEW!



Model CFI 1,000 to 10,000 mc



Model CFI 950 to 11,260 mc

TRANSISTORIZED CALIBRATED FIELD INTENSITY RECEIVER...

This is the Polarad Model CFI—the latest in field intensity measurement test equipment. It's transistorized for portability...excellent for airborne applications. The built-in impulse calibrator enables RFI measurements in accordance with latest military requirements. Plug-in tuning heads under development, will extend the frequency range beyond the present 1,000 to 10,000 mc capability.

FUNCTIONS AS A MULTI-PURPOSE MICROWAVE RECEIVER, TOO!

As an all-purpose receiver, the CFI offers AM, FM, CW and Pulse capability. These features make this the most versatile receiver you've ever used: 3 impulse bandwidths; 70 db dynamic range; sensitivity —90 dbm; direct reading meter circuits. You can use the CFI for all general laboratory and field work. Call your Polarad representative for a demonstration, or mail the card.

SPECIFICATIONS

FREQUENCY: 1,000 to 10,000 mc in four plug-in tuning units (950 to 11,260 mc as receiver)
SENSITIVITY: to —90 dbm
FREQUENCY DIAL ACCURACY: ±1%
IMPULSE CALIBRATOR includes built-in impulse generator, RF attenuator (—60 db), IF attenuator (0—20 db), in 1 db steps
ANTENNA EQUIPMENT: 4 directive and 1 omni-directional; mounting tripod
OUTPUTS: Audio, Video and Recorder
METER CIRCUITS: Average and slideback peak, direct-reading peak and quasi-peak
INTERNAL CALIBRATION SIGNAL: Impulse type; 1 to 10 gc ±0.5 db flat output
IMPULSE BANDWIDTHS: 1 mc, 5 mc, and 8 mc
VIDEO BANDWIDTH: 3.5 mc
IMAGE REJECTION: 60 db
POWER INPUT: 12 volts DC; 115 volts AC; 50 to 420 cps

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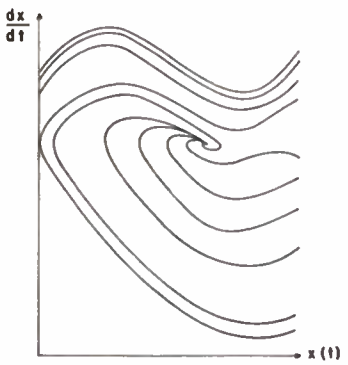
Please send me information and specifications on:
☐ Model CFI Calibrated Microwave Field Intensity Receiver

MY APPLICATION IS_____
Name_____
Title_____ Mail Station_____ Dept._____
Company_____
Address_____
City_____ Zone_____ State_____

NON-LINEAR DIFFERENTIAL EQUATIONS

The non-linear differential equation shown above the phase-plane plot governs the behavior of one form of the well-known electronic circuit used for the reception of weak communication signals. The circuit, known as the phase-lock loop, has a mechanical analog consisting of a mass subjected simultaneously to non-linear restoring and resistive forces.

$$\frac{d^2x}{dt^2} + (\alpha + \kappa \cos x)\frac{dx}{dt} + \gamma \sin x = \beta$$



INTEGRAL CURVES IN THE PHASE PLANE

Simpler equations of this basic type have been studied for a number of years, and computers have produced many phase-plane plots for special values of the parameters, such as shown above. For fuller understanding of the circuit operation, however, new analytical methods are being investigated at Philco which may give an overall picture of the solutions of this, and even more general, equations.

This study is typical of the Philco Scientific Laboratory's activity in applied mathematics. Philco offers attractive staff positions in this field and in a number of others such as electrodynamic theory, optical communications, pattern recognition, and solid state electronics. Applications and inquiries may be addressed to: Mr. W. P. Diefenbach, Personnel Manager, Philco Scientific Laboratory, Blue Bell, Pennsylvania.

PHILCO SCIENTIFIC LABORATORY

PHILCO

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MICROELECTRONICS SPECIALIST

MELPAR offers excellent opportunity for a semiconductor device design and solid state circuit specialist to technically direct a program in microcrystalline microelectronics in a newly organized solid state physics facility. Must have had prior experience in this field and/or in the design of monocrystalline semiconductor devices. This is an excellent growth opportunity for creative work in this field, under ideal conditions, in which publication is encouraged. Salary open and dependent on prior experience.

Write in confidence to:
John Haverfield, *Employment Manager*

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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.

ENGINEER

$115 week. Approx. College grad. with a major in either electrical or mechanical engineering. Work involves investigating major appliances and building products in laboratory of national magazine. Prefer some experience but will consider recent college graduate. Veteran or Draft Deferred. Send typed resume to: Hearst Magazines, Personnel Office, 309 West 56 Street, New York 19, N. Y.

TELEVISION BROADCAST TECHNICIAN

Opportunities for video technician with group television stations. Write G. G. Jacobs, Corinthian Broadcasting Corp., 302 South Frankfort, Tulsa, Okla.

BIO-MEDICAL RESEARCH

The Department of Physiology and Pharmacology seeks a young man interested in bio-medical research. Prefer a recent graduate with extensive background in circuit theory to assist in developing instrumentation for bio-medical research. Interested persons should communicate with: Albert F. Kelso, Ph.D., Dept. of Physiology and Pharmacology, c/o Chicago College of Osteopathy, 1122 East 53 Street, Chicago 15, Ill.

BROADCAST ENGINEERS

The United States Information Agency seeks Electronics Engineers with Communications or Broadcasting Systems experience for its Voice of America. The positions are located in Washington, D. C. and abroad. Domestic assignments are mainly concerned with the design and planning of broadcasting systems and transmitting plant up to 1000 kilowatts.

Foreign positions are of a construction management type.

(Continued on page 96A)



A panoramic concept on space... you find it at Lockheed

Long experience and proven capability give Lockheed a great advantage: perspective on space. This results in programs that cover the broad spectrum of aerospace dynamics; that present the most far-reaching technological problems; that offer endless creative challenge to Scientists and Engineers; that give them new concepts to explore—new goals to reach.

This far-ahead program in the concepts and vehicles of space creates a climate most favorable to advancement in the state of the art as well as in professional status.

Scientists and Engineers with an eye to a secure and rewarding future will find these openings at Lockheed worthy of serious consideration: Aerodynamics; thermodynamics; dynamics; electronic research; servosystems; electronic systems; physics (theoretical, infrared, plasma, high energy, solid state, optics); electrical and electronic design; structural design (wing, empennage, fuselage); human engineering; reliability; wind tunnel design. Write today to Mr. E. W. Des Lauriers, Manager Professional Placement Staff, Dept. 1802, 2402 N. Hollywood Way, Burbank, California. An equal opportunity employer.

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A DIVISION OF LOCKHEED CORPORATION

ENGINEERS SCIENTISTS MANAGERS

New long term diversified development and design contracts create unusually attractive opportunities at the Link Division of General Precision, Inc. Qualified men, proficient in broad systems and equipments engineering, will be interested in these commercial and military projects. Both aircraft and space vehicle systems are involved. Excellent salaries and ideal living in the Binghamton, New York area will attract the qualified professional man or manager seeking advancement opportunity and challenging work.

PROJECT MANAGER—laboratory precision measurement, visual displays and special projects including G.S.E., simulators and checkout equipment. PROJECT MANAGER—electro-optical precision measurement systems. MANAGER PRODUCT SUPPORT—12-15 years mechanical design, electronic packaging, model shop construction, and department supervision in military programs. Direct product design department, model shop, product programs and advanced development.

SENIOR & JUNIOR ENGINEERS

DEVELOPMENT · DESIGN · ANALYSIS

PRINCIPAL ENGINEER—visual systems projects. OPTICS—optical, electro-optical measuring, inspection and scanning systems. OPTICS—periscopic, projection and relaying lens, analysis of optical problems, laboratory proposals. DIGITAL—transistor circuits, switching circuits, computer logic, "NOR" logic, direct-coupled transistor logic, proposals. VISUAL DISPLAYS—electronic systems project responsibility. Supervise design, test and verification of prototype model and preparation of engineering data. SIMULATION—analog computing devices, audio systems, transistorized amplifiers, radio aids, radio navigation, aircraft communications. SIMULATION—program digital computers, digital systems design. SIMULATION—define problems, program and solve equations of flight simulation, specify components, initiate requirements for design and configuration of electronic and electro-mechanical systems. MECHANICAL—systems design of servos, hydraulics, missiles, life support systems. MECHANICAL—systems design of mechanisms, structures, hydraulics, electro-mechanical packaging, materials, plastics. AERONAUTICAL—simulation, project responsibility for computation of equations, defining motion and engine performance of aircraft, specification interpretation, concept determination, data search and liaison, data analysis and processing, test guide inception and computation. AERONAUTICAL—process raw aerodynamic coefficient data and engine data into equation form for electronic simulation, engine performance calculations, test guide to check simulator. ELECTRONIC SCIENTIST—airframe and spacecraft performance, stability and control, engine performance, analog simulation and digital computations.

All positions require an appropriate degree—advanced degree is highly desirable for managerial and senior positions. Minimum experience required is 4 years—an additional 4 years experience is required for managerial and senior positions.

Qualified men are invited to phone collect (RAymond 3-9311) or write Mr. James T. Gibbons. An equal opportunity employer.

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The expansion of our Electronic Tube Division to meet the growing demands of the commercial, industrial and military markets has created several outstanding opportunities for experienced ENGINEERS. Salaries are open, depending on experience and qualifications. The plant is located in Elmira, N.Y. . . . a delightful community for work and recreation in the Finger Lakes region of upstate New York.

Experience in any of the following areas is desirable:

- **MICROWAVE TUBES & DEVICES**

  Senior Engineers with experience in design or manufacture of MAGNETRONS, KLYSTRONS, TRAVELING WAVE TUBES and other microwave tubes and devices.

- **IMAGE TUBES**

  Design or manufacture of STORAGE TUBES, CAMERA TUBES and other imaging devices.

- **CATHODE RAY TUBES**

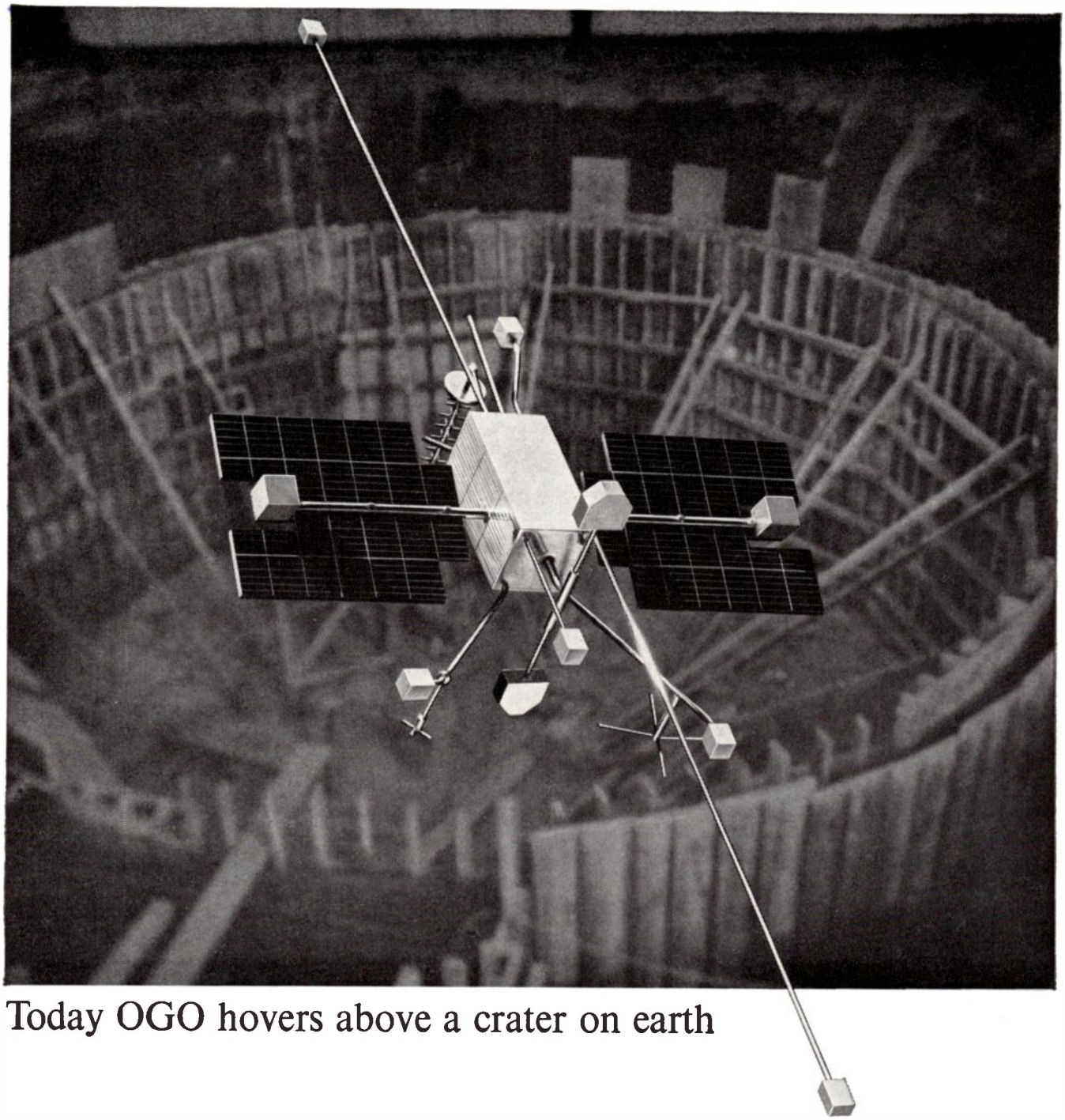
  Design or manufacture of black and white PICTURE TUBES, and INDUSTRIAL and MILITARY TUBES.

Write or send resume to:

William Kacala,
Technical Recruiting,
P.O. Box 284, Elmira, N.Y.
or phone collect
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ELECTRONIC TUBE DIVISION ELMIRA, NEW YORK



Today OGO hovers above a crater on earth

Soon a new space chamber 30 feet in diameter will fill this deepening bowl of earth. Here OGO (NASA's Orbiting Geophysical Observatory) will be subjected to conditions of solar heating, vacuum, and vehicle radiation to the cold of outer space. The new space chamber will be the sixth at STL. It will enable engineers and scientists working on OGO, Vela Hotel and other STL projects to test large, complete spacecraft as well as major subsystems. And along with other advanced facilities at STL's Space Technology Center, it will provide unusual scope for engineers and scientists to verify and apply new techniques in design, development and fabrication of spacecraft. STL's expanding space programs have created new opportunities for engineers and scientists in the following fields: Aerodynamics, spacecraft heat transfer; Communication Systems; Electronic Ground Systems; Power Systems; Propellant Utilization; Propulsion Controls; Reentry Body Evaluation; Systems Analysis; Thermal Radiation; and Trajectory Analysis. All qualified applicants are invited to write Dr. R. C. Potter, Manager of Professional Placement and Development, for opportunities with STL in Southern California or at Cape Canaveral. STL is an equal opportunity employer.



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ENGINEERS

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RADAR & COMMUNICATIONS SYSTEMS ANALYSIS

SPECTRUM SIGNATURE MEASUREMENT

PROPAGATION RESEARCH

ANTENNA SYSTEMS

SIGNAL ENVIRONMENT ANALYSIS

The ARMOUR RESEARCH FOUNDATION has immediate staff openings at the newly established ELECTROMAGNETIC COMPATIBILITY ANALYSIS CENTER in Annapolis, Maryland. Engineers interested in the challenging and vital field of Electronic System Compatibility are invited to inquire about positions in the above areas:

These positions require a minimum of a B.S. degree.

We offer excellent salaries and related benefits; an opportunity for increased professional growth as a member of an outstanding research staff; up to FOUR WEEKS VACATION; and TUITION-FREE GRADUATE STUDY at nearby colleges or universities.

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ARMOUR RESEARCH FOUNDATION

ELECTROMAGNETIC COMPATIBILITY ANALYSIS CENTER
USNEES, ANNAPOLIS, MARYLAND

Positions Open

(Continued from page 92A)

Broadcasting and Communications Systems cover frequency ranges in the medium and high frequency bands.

Foreign Service appointments will be made at the entrance level of $10.785 per annum and range to $12,025 plus overseas allowances. Appointment levels for domestic service: $8955, $10,635, $12,210, $13,730 and range to $15,030 per annum commensurate with background and experience.

Relocation expenses paid in accordance with government regulations.

Applicants *must* be AMERICAN citizens.

Please send background resume or application for Federal Employment (SF-57) to: Employment Branch, U. S. Information Agency, Washington 25, D. C.

SENIOR ELECTRONICS ENGINEER

Excellent opportunity exists for professional advancement with small research organization. Seeking B.S.E.E. or M.S.E.E. with 4 to 10 years experience including design work in R. F. and microwave fields in the 30 to 3000 MC regions. Interest in Project Management responsibilities desired. Contact Richard A. Johnson, Vice President, Systems Research Laboratories, Inc., 500 Woods Drive, Dayton 32, Ohio. Phone 244-4051.

MAGNETIC CORE COMPONENTS

Electrical Engineer with design experience involving components for pulse circuitry with emphasis on filters, R. F. coils and pulse transformers. A background involving the uses of ferrite and powered iron ores is highly desirable. Send complete resume and salary requirements with first letter to Paul E. Dicker, Chief Engineer, Aladdin Electronics, Nashville 10, Tennessee.

ELECTRONIC MUSICAL INSTRUMENT ENGINEER

Graduate Electrical Engineer to join research group on electronic musical instruments. One to three years electronic experience required, with some musical background desirable. Write Albert Meyer, Engineering & Research Dept., The Baldwin Piano Co., 1801 Gilbert Ave., Cincinnati 2, Ohio.

EDUCATION AND RESEARCH IN BIO-MEDICAL ENGINEERING

An educational program designed to prepare life and physical scientists, doctors of medicine, and engineers for research in life systems, through the current period of transition of biological sciences from the natural history stage of inquiry by inductive empiricism to that of inquiry by deductively formulated theory. An intensive four-term (12 month) preparation in methods of logical thought, mathematics, physics, and electrical theory for doctors of medicine and biologists, and one-term (3-month) preparation in life sciences for engineers, makes possible joint course work in the final three terms (nine months). Thesis research on biological systems is conducted at the Presbyterian Hospital by teams drawn from both disciplines. The program leads to a graduate degree at Drexel. Limited stipend and scholarship funds are available. Assistance with fellowship application to National Institutes of Health is offered. For inforamtion, James W. Dow, M. D. Dept. of Bio-medical Engineering, Drexel Institute of Technology, 15 North 32 St., Philadelphia 4, Pa.

(Continued on page 100A)



ENGINEERS
SYSTEMS-DESIGN
DEVELOPMENT
PHYSICISTS

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AS PARTNER IN AN OUTSTANDING

OPPORTUNITY

Vitro Laboratories has been awarded significant and extensive new responsibilities in systems engineering, research and development. We are a young and vigorous organization . . . we are expanding . . . dynamically! Important work lies ahead. We invite you to share in this opportunity.

Vitro is a good place to work. Here you will find a professional atmosphere conducive to creative and imaginative thinking. You will advance as fast as your talents allow. Skilled clerical and technical support frees you to concentrate on the important aspects of your assignment, and salaries are at a level worthy of your attention.

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RESEARCH & STUDY
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Basic and applied research
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- Operations Research
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- Analog Computers
- Mechanical Design
- Field Engineering

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Furthering your education? Sylvania encourages, sponsors graduate study at Stanford and other nearby institutions.

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Government Systems Management
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GENERAL SYSTEM

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To meet the increasing need for advanced knowledge in the communications area, Bell Aerosystems Company is expanding, creating new positions for Research Scientists. Programs will be undertaken in research and advanced development in the general technical areas of modulation and demodulation, information theory and communication in the presence of noise and interference as applied to:

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MODERN COMMUNICATION SYSTEMS

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Permanent positions are immediately available at all professional levels of communication engineering for Research Engineers with PhD, MS or BS and some experience related to the various areas of study.

Inquiries are invited.
Please address Mr. George Klock.

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By maintaining a reputation for technical engineering excellence and a position of leadership in research and development, Electro-Mechanical Research, Inc. (e-m-r) has become the largest of all companies specializing in telemetry systems. Today, with the space age bringing new and vital significance to the field of electronics, e-m-r affords mature engineers an environment for maximum responsibility and personal growth.

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Minimum of 3 years experience in detailed design in such areas as digital circuits — using semiconductors and magnetic elements — transistor implemented low level amplifiers, ferrite core memories, logic design of digital computer, analog to digital converters, feed-back amplifiers, servomechanisms, and RF transmitters and receivers — 100 to 1000 megacycle range.

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Meet One Manager of Engineering Who's Happy About Losing His View!

(An unusual bit of evidence to support our claim of "outstanding engineering opportunities")

AS MANAGER OF ENGINEERING, and a key contributor to successful and far-reaching expansion plans, Dick Gifford feels the loss of one set of windows is a small personal price to pay for more than 22,000 square feet of new space for expanding Engineering operations.

But this is only about 20% of all the plant expansion now underway. A second wing will further increase total plant floorspace owned and leased from 400,000 to over 500,000 square feet by March, 1962.

In recent years, the Communication Products Department has built a healthy mix of industrial, commercial and military programs. Products represent a spectrum of communication systems—from tiny, personal two-way radios, to continent-spanning microwave radio relay networks. Now, by continuing to grow along virtually this entire front, the professional opportunities newly created carry a hard-to-find combination of two key career factors: rapid growth potential, *and near-optimum stability.*

High-Priority Programs for Defense; Highly Competitive Products for Industry—

In the military sphere, we recently added a new program of great importance to the nation's security—developing the intersite communications link for Titan II operational bases... a system which must be *survivable and reliable.*

In our commercial areas, a number of new opportunities have been created by expansion of such programs as: a 600 channel, fully transistorized and miniaturized multiplex telephone carrier system; a 600 channel microwave radio relay system (also fully transistorized) embodying a heterodyne repeater, parametric and traveling wave tube amplifiers; compact, transistorized two-way FM mobile radio; highly miniaturized personal communication equipment; tropospheric scatter equipment with modulation techniques, designed specifically for thin route applications.

Immediate Openings In These Areas:

Military Communication Systems Engineering
Military Communication Equipment Design
Telephone Carrier Circuit Design
RF and Video Circuit Design
Communication Equipment Advance Development
Solid State Receiver Design
2-Way Mobile Transmitter Design
2-Way Mobile Systems & Reliability Engineering
2-Way Radio Mechanical Design

■ Your letter or resume will receive careful attention and a prompt, personal reply. Write in confidence to Mr. W. J. Kelly, Div. 53-MB, Communication Products Department, General Electric, Mountain View Road, Lynchburg, Virginia.

• Lynchburg offers the advantages of a modern, progressive community located in the beautiful Blue Ridge area of Virginia, only four easy driving hours from Washington, D. C.

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ELECTRONICS ENGINEERS

YOU can Contribute Now to Many Advanced Projects in SSB Communications

Over the past several years, most of the military HF, SSB development contracts have gone to General Dynamics/Electronics. This response to the Company's broad technical competence has resulted in steady, solid growth, creating important positions for talented engineers in advanced design, development and production of mobile communications equipment. The current press of recently accelerated military procurements has further increased staff requirements in the following areas.

SYNTHESIZERS

Working knowledge of SSB techniques is preferred, but FM and AM experience with sufficient background in synthesizer-oscillators will be considered. Must be capable of leading work in advanced development and sophisticated frequency standard designs and able to conceive and develop novel and unconventional ideas.

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Requires strong background in transistor circuit design and in arrangement of circuits into complete communications systems. Will be responsible for the design of advanced SSB receivers.

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6 to 10 years electronics experience required with emphasis on high power vacuum tube amplifier design. Experience in SSB linear amplifier techniques is desirable.

RELIABILITY

To make reliability analyses of circuits and components for military electronic equipment. Requires BSEE with 4 years experience in engineering design plus 4 years in reliability.

■ If you would like to know how contributions to General Dynamics/Electronics' unique capabilities can advance *your* future, submit a detailed resume to Mr. M. J. Downey.

GENERAL DYNAMICS | ELECTRONICS

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Positions Open

(Continued from page 96A)

ANTENNA ENGINEERS

Experienced Antenna Engineers. Physicists. B.S., M.S., Ph.D. Citizens of U. S. A. Very challenging theoretical, applied research and engineering problems on ground, missle, satellite and deep space probe antenna systems. Write H. W. Haas, Physical Science Laboratory, New Mexico State University, University Park, New Mexico.

OVERSEAS

Robert College, in Istanbul, Turkey, presents a challenge in education where East meets West. An opportunity to contribute significantly to the development of a young republic is available to specialists in engineering, business administration and economics, the sciences, and the humanities. Graduate degrees required. Adress inquiries to Dr. Howard P. Hall, Vice President, Robert College, Bebek Post Box 8, Istanbul, Turkey; with copy to the Near East College Association, 548 Fifth Avenue, New York 36, N.Y.

AUDIO ENGINEER

BSEE or BS Physics. 3 to 5 years experience in the design and development of moving coil microphones and/or miniature dynamic speakers. Manufacturing experience desirable. Progressive company manufacturing electroacoustic transducers in the Metropolitan area. Salary competitive. Send complete resume to Box 2062.

(Continued on page 104A)

Are You "Going Someplace?"



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PROJECT ENGR.—R&D heavy exp. Aero & fluid dynamics.

GENERAL MGR.—Industrial TV, Instrum., Mobile Comm'cations background Top Administrator, Technically Fortified.

PHD'S—Electro., Mech. & Nuclear Many openings. "To Sky and Beyond."

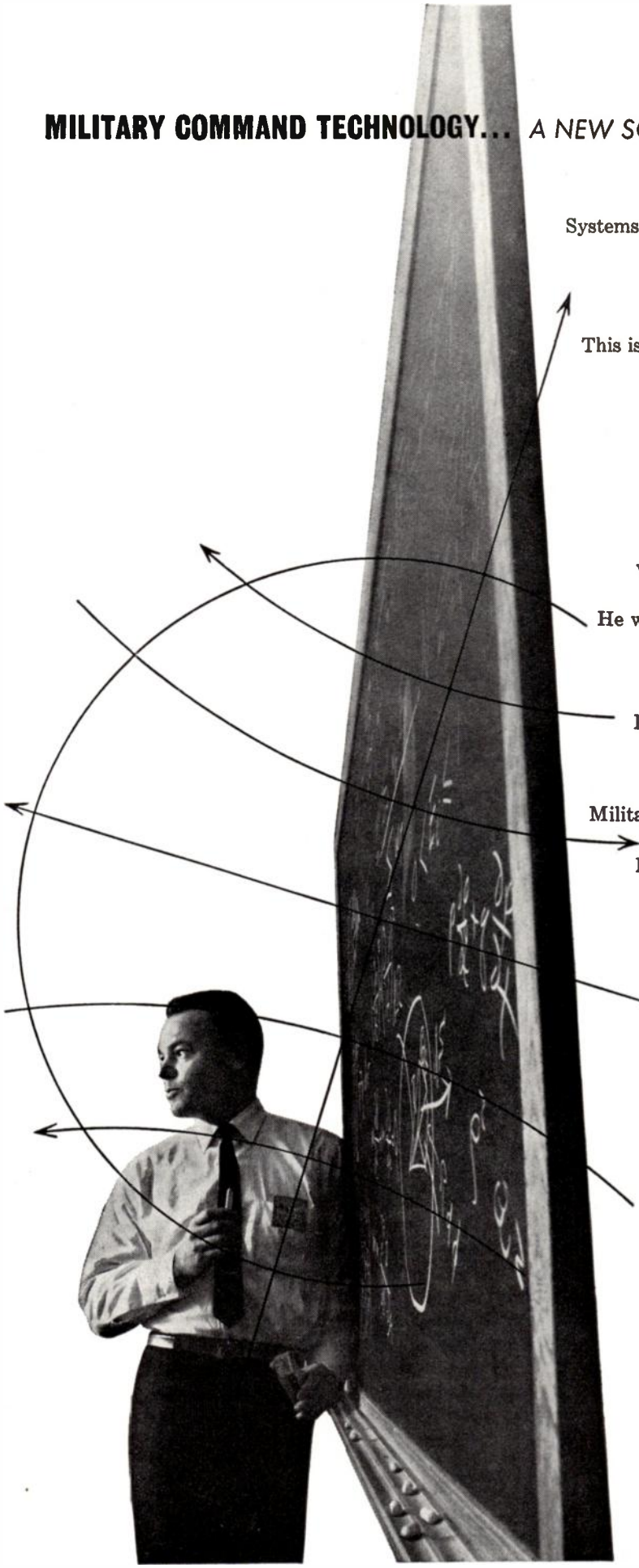
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MILITARY COMMAND TECHNOLOGY... *A NEW SCIENCE FOR NATIONAL DEFENSE*



Systems that instantly provide the military Commander with the information necessary for decision. Systems to enable the Commander to control all his forces under any conditions.

This is the purpose of Military Command Technology. It is the work carried out at MITRE. It includes command systems, control systems, intelligence systems, warning systems, and support systems. It encompasses a vast network of interrelated, constantly evolving systems that protect our country.

The designer of these systems must be able to visualize how war would be fought. He will work closely with the nation's top policy makers. He will help solve the problems of military command — nature, deployment, and use of weapons; war-plans; control of forces; missions; logistics; support and intelligence operations. But, most important — he must be able to apply existing and predictable technology to the abstract problems of future military command.

Military Command Technology, in short, is a systems engineering task of overwhelming importance. MITRE has men who can get the job done. And there is room at MITRE for more such men — top professionals who feel they want to serve their country in a vital area. There are key assignments available in system analysis and planning; intersystem integration; general system engineering; initial system design; and research and experimentation.

If you feel you can advance this new science, you are urged to write Mr. R. R. Everett, Vice President — Technical Operations, The MITRE Corporation, Box 208, Dept. MB4, Bedford, Mass.

THE MITRE CORPORATION

MITRE is an independent, nonprofit corporation working with — not in competition with — industry. Formed under the sponsorship of the Massachusetts Institute of Technology, MITRE serves as Technical Advisor to the Air Force Electronic Systems Division and is chartered to work for such other Government agencies as FAA.

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- OPTICAL SPECIALISTS
- LOGIC DESIGNERS
- PROJECT ENGINEERS
- ADVANCED ELECTRONICS

Scientists and engineers needed with experience in all phases of digital computer and control system design.

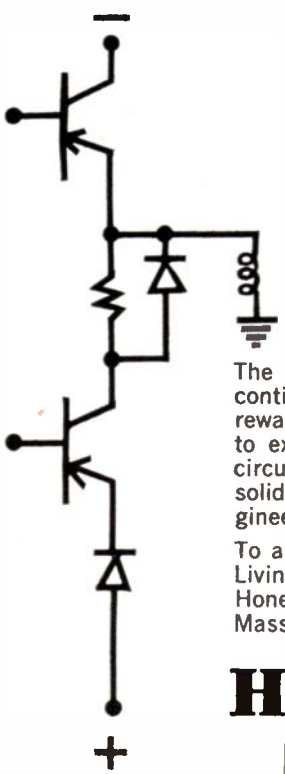
Systems organization, logic design, magnetic core and drum memories, dynamic analysis, and electro-optical correlation devices. Also advanced areas such as high-speed tunnel-diode techniques, thin films, and hybrid analog-digital techniques. Applications include airborne digital equipment, numerical machine control, photogrammetric equipment, and special-purpose control computers. Both commercial and military programs, emphasizing advanced development and research. We think you will find this work unusually stimulating and satisfying. Comfortable and pleasant surroundings in suburban Detroit.

If interested, please write or wire A. Capsalis, Research Laboratories Division, The Bendix Corporation, Southfield, Michigan.

Research Laboratories Division

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CIRCUIT DESIGN ENGINEERS
MANUFACTURING ENGINEERS
SOLID STATE ENGINEERS
ELECTRO-MECHANICAL DESIGN ENGINEERS

The Boston Division of Minneapolis-Honeywell offers continued growth and stability within a dynamic and rewarding atmosphere. Qualified engineers are invited to explore the opportunities available in the fields of circuit design, electro-mechanical component design, solid state device engineering, and manufacturing engineering with members of our technical management.

To arrange an interview please contact: Mr. G. Ronald Livingstone, Dept. B, Boston Division, Minneapolis-Honeywell, 1400 Soldiers Field Road, Boston 35, Massachusetts.

Honeywell

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Opportunities also exist in other Honeywell facilities coast to coast. Send resume to H. B. Eckstrom, Minneapolis-Honeywell, Minneapolis 8, Minnesota.

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Create Unusual Assignments at Various Levels of Responsibility in Research, Development and Engineering

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For creative staff engineering in digital electronics, air data systems, pressure sensors and transducers, optical electronic systems, digital computers and sensors, solid state physics, logical design, systems analysis and synthesis, infrared and visual scanning techniques, information theory, systems development for navigation and guidance of aircraft, missiles and space vehicles, heat transfer analyses, and reliability.

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For project engineering assignments on automatic navigation, air data and flight control systems and devices. Also assignments in ground support equipment in the above areas, as well as in our new ordnance equipment group. Experience in electro-mechanical/optical systems desirable.

Please forward resume, in confidence, to Mr. Donald Svenson

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(Advanced Inertial Guidance Systems and Components)

Systems Engineers and Mathematicians
Electronic Circuit Engineers
Mechanical Design Engineers
Instrument Engineers
Electromagnetic Engineers
Radar Systems Engineers

Los Angeles Research and Development Laboratory

(Advanced Inertial Guidance Systems and Airborne Digital Computers)

Digital Computer Development Engineers
Research and Development Engineers
Transistor Circuit Design Engineers
Systems Engineers

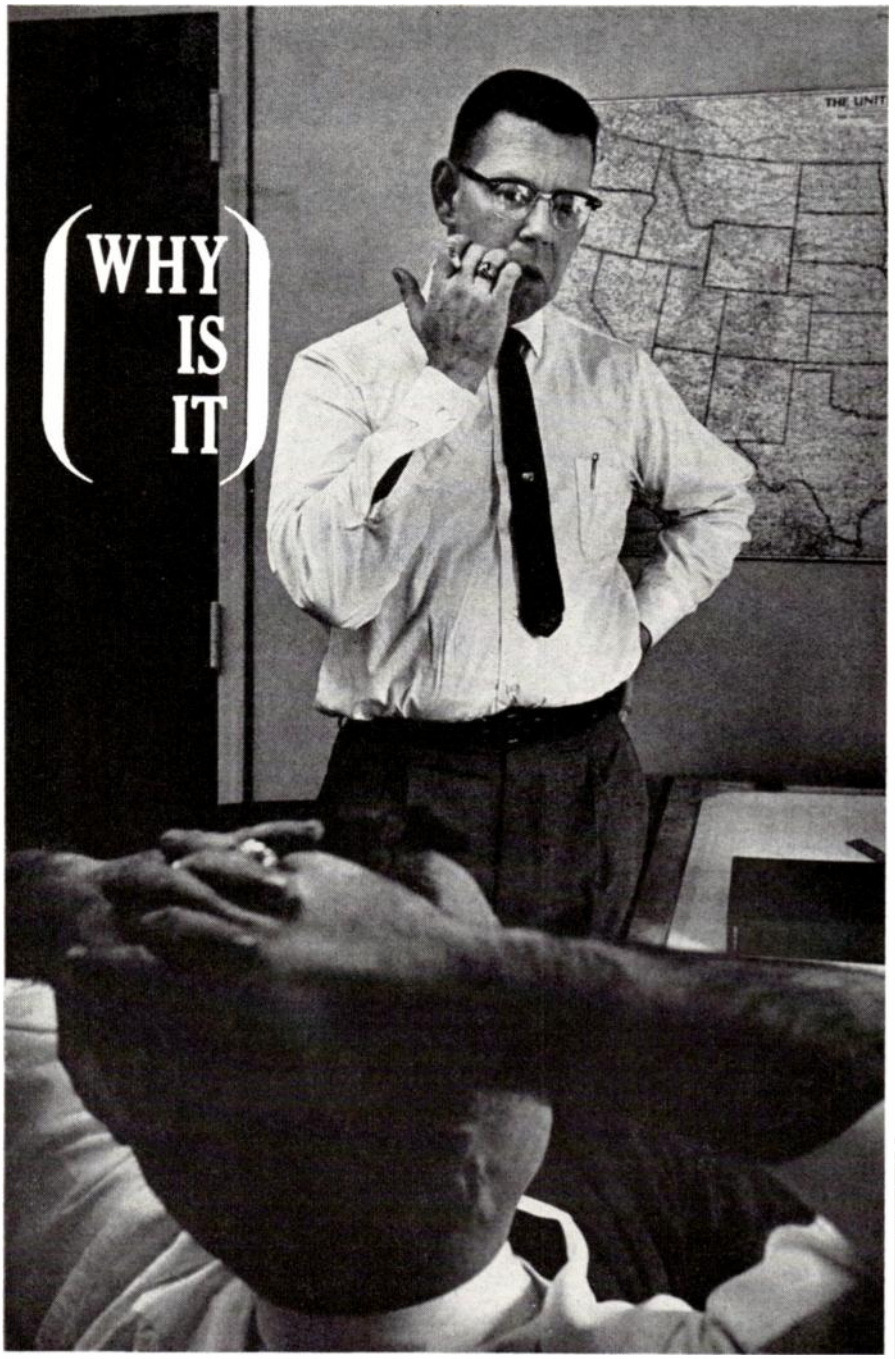
Milwaukee

Military Defense Systems, Program and Project Engineers
Radar Design and Development Engineers
Radar Systems Engineers
Radar Test Engineers
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If you are experienced and interested in any one of these positions, please send your resumé to Mr. G. F. Raasch, Director of Scientific and Professional Employment, Dept. 5753, 7929 S. Howell, Milwaukee 1, Wisconsin. We are an Equal Opportunity Employer.

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ORI

Positions Open

(Continued from page 100A)

PROFESSORS

Ph.D.'s required in electrical engineering (both new and those established in their field) for a rapidly expanding and state supported urban University with a strong Graduate School. Outstanding opportunities in growth and salary—all located in an excellent total environment involving a research-based community in New York State. Box 2064.

PROFESSOR

Ph.D. wanted to head E. E. Department in metropolitan New York school, September 1962. Educational & industrial experience desired. Excellent opportunity to participate in teaching, research & growth of Department. Write box 2065.

TEACHING-ELECTRICAL ENGINEERING

Electrical Engineering staff openings in an expanding undergraduate program. Candidates should possess graduate degrees (M.S. or Ph.D.). Encouragement is given to private consultation and research. Rank and salary commensurate with academic attainment and experience. Apply: Wm. Shewan, Chairman, Dept. of E. E., College of Engineering, Valparaiso University, Valparaiso, Indiana.

SCIENTISTS AND ENGINEERS

For JPL's Lunar and Planetary Programs with experience in any of the following areas: Celestial Mechanics, Orbit Determination, Space Navigation Theory, Trajectory Studies, and Systems Analysis. A new technology, based on celestial mechanics, mathematical statistics, and engineering systems analysis, is evolving in the space exploration program. Both theoretical and project positions are available for experienced people in this expanding field. Send resume to P. I. Locke, Jet Propulsion Laboratory, California Institute of Technology, 4800 Oak Grove Drive, Pasadena, California. "An equal opportunity employer."

ENGINEERS

The Federal Aviation Agency is filling three engineer positions:

1. General Engineer (Aircraft Electrical/Electronic) GS-13, $10,635 per annum.
2. Aeronautical Design Evaluation Engineer GS-12, $8955 per annum.
3. Aeronautical Power Plant Design Evaluation Engineer (Engines) GS-12, $8955 per annum.

Incumbents will serve as technical specialists in the Engineering and Manufacturing Branch, Flight Standards Division. Some travel is required. Call Jamaica, N. Y. OL 9-7000 ext. 209 for further information.

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Permanent positions as radio and electronics instructors for energetic young men up to 30 years old. Excellent and unlimited pay for persons with the right desire and qualifications. College definitely preferred. Excellent advancement opportunities. Send complete resume to Elkins Institute of Radio and Electronics, 2603 Inwood Road, Dallas 35, Texas.

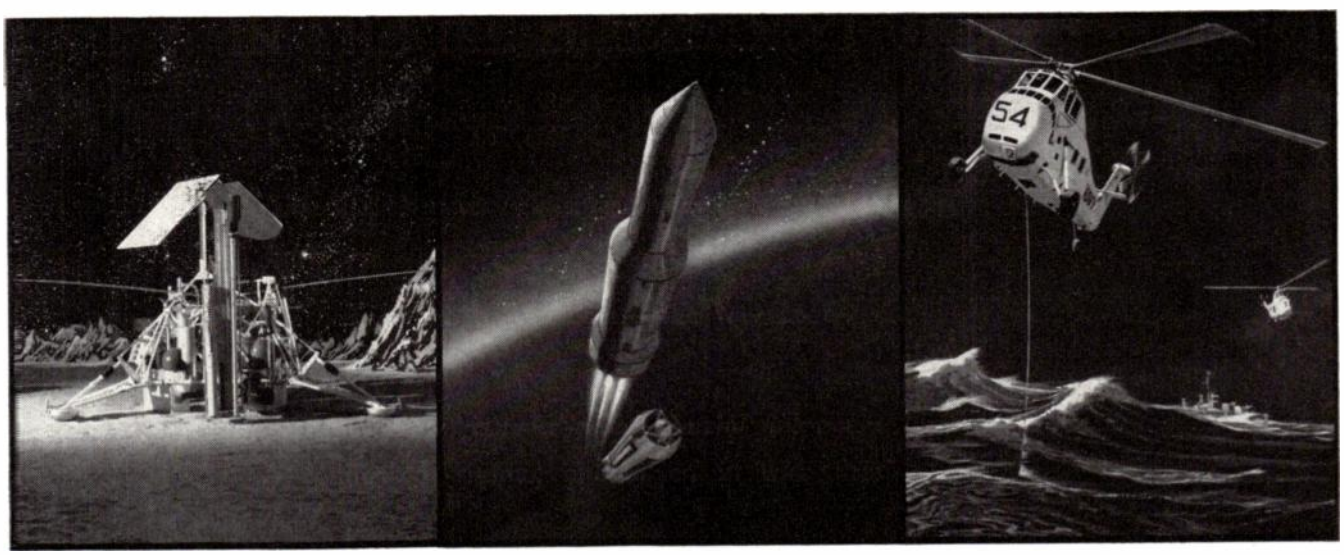
PROFESSOR

Professor of Engineering to teach electronics and work in the development of undergraduate and graduate programs. Ph.D. and teaching ex-

(Continued on page 106A)

ELECTRONICS & SPACE ENGINEERS

New long-range contracts at Ryan have created top-salaried opportunities in sunny San Diego



RYAN ELECTRONICS *was selected by Hughes Aircraft Company, Inc., to design, develop and fabricate the Radar Altimeter and Doppler Velocity Sensor equipments for NASA's Surveyor Lunar Soft Landing Spacecraft.*

SATURN LAUNCH VEHICLE *will employ radar altimeters now under development at Ryan Electronics. Engineers interested in a space age future are invited to get in touch with fast-moving Ryan Electronics right now.*

WORLD LEADER *in the design and production of Doppler Navigators for all types of aircraft now flying or projected, Ryan Electronics seeks engineers for challenging opportunities in sunny San Diego.*

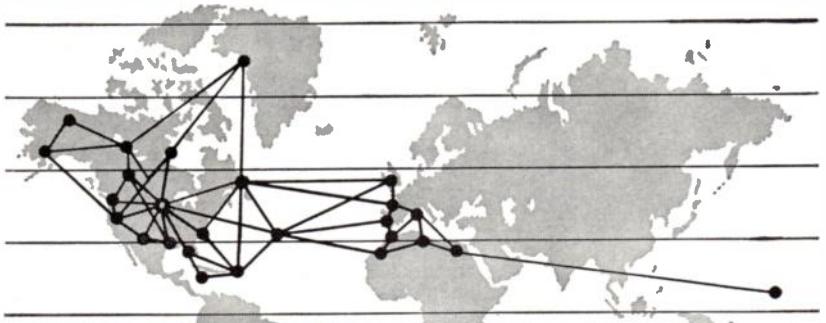
Ryan is on the move! New contracts call for **years** of research, design and development work on such space age projects as radar altimeters for the Surveyor lunar soft landing vehicle and the Saturn launch vehicle. Ryan Electronics is also the world leader in Doppler navigation systems, as well as in the field of space guidance and control. If you are a career engineer interested in a top-salaried job with stable, diversified Ryan... if you would like to live in cool, sunny San Diego on the blue Pacific where the living is America's finest... **Send resume in complete confidence to: JAMES C. KERNS, Ryan Aeronautical Company, 5650 Kearny Mesa Rd., San Diego 12, California.**

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Senior Systems Scientists, Engineers: Opportunities NOW on World-Wide Digital Communications & Control System — 465-L



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These assignments with the Communications Systems Division are under a *prime ITT contract.* The scope of the work is indicated by the scope of the program.

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- airborne strength, location, fuel radius and target accessibility of each aircraft
- number of missiles on ready
- crew availability for aircraft and missiles
- number of aircraft in maintenance and maintainance time estimates
- force dispositions and target destruction after strike

COMMAND DECISIONS, based on this real-time portrayal of relevant data, will be transmitted instantly to the most remote bases over 465-L's global network.

OPENINGS IN CHICAGO, LOS ANGELES, & FT. WAYNE (IND.) FOR WORK ON 465-L AND OTHER MILITARY AND COMMERCIAL COMMUNICATIONS & CONTROL SYSTEMS

SENIOR COMMUNICATIONS SYSTEMS SCIENTIST, MS-PhD
Perform fundamental feasibility studies supporting and leading to preliminary systems design.

SENIOR HUMAN FACTORS SCIENTIST, MS-PhD
Establish, direct, evaluate human factors design; manning requirements.

SENIOR SYSTEMS DESIGN ENGINEERS, BS-MSEE
Semiconductor circuit design, audio frequency communications, logic, digital data processing, teletype, switching, pulse techniques, machine control.

SENIOR CIRCUIT DESIGN ENGINEERS, BS-MSEE, PhD
Develop advanced transistorized digital and/or pulse circuits.

SENIOR SYSTEMS ANALYSTS, MS-PhD
Determine systems requirements, formulate and evaluate functional system specifications.

SENIOR PRODUCT ENGINEERS, BS-MSEE
Evaluate & redesign computers, switching, data handling equipment, etc. during production stage. Establish test procedures. Must know transistor & logic circuit analysis; Mil Specs.

Also, MANUFACTURING ENGINEERS, TEST ENGINEERS, QUALITY CONTROL ENGINEERS, & SENIOR TECH WRITERS.

Qualified Scientists and Engineers are invited to write in confidence to Mr. D. C. Schaller, Dept. 12 B

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COMMUNICATIONS SYSTEMS DIVISION
500 N. Pulaski Road Chicago 24, Illinois

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Positions Open

(Continued from page 104A)

perience in the field are required. Also, Assistant Professor to teach electronics. M. S. required, Ph.D. desirable. Excellent opportunities in a rapidly growing college. Apply to the Head of the Engineering Department, Los Angeles State College, Los Angeles 32, California.

ASSISTANT or ASSOCIATE PROFESSOR

For January or September 1962, assistant or associate professor of Electrical Engineering required to teach undergraduate and graduate courses. Ph.D. required. Specialty networks, computors, or electromagnetic theory. Write to Dr. H. Mahrous, Electrical Engineering Dept., Pratt Institute, Brooklyn 5, New York.

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Cornell Aeronautical Laboratory of Cornell University has several attractive openings for qualified Engineers or Scientists to participate in research in the thermodynamics of advanced electronic devices.

Projects include investigations of the thermal aspects of microminiature and new solid state electronic devices.

Appropriate experience in thermodynamics, cryogenics, solid state physics, and electronics required.

Contact E. P. Rentschler, Cornell Aeronautical Laboratory, Inc., 4455 Genesee Street, Buffalo 21, New York.

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Electronic Engineers needed in the field of research, development and installation of Army Air Defense Systems. Salary to $10,635 per annum. Positions are in the Career Civil Service and are subject to Federal Civil Service benefits and requirements. Apply Civilian Personnel Office, Fort George G. Meade, Maryland.

McMASTER UNIVERSITY

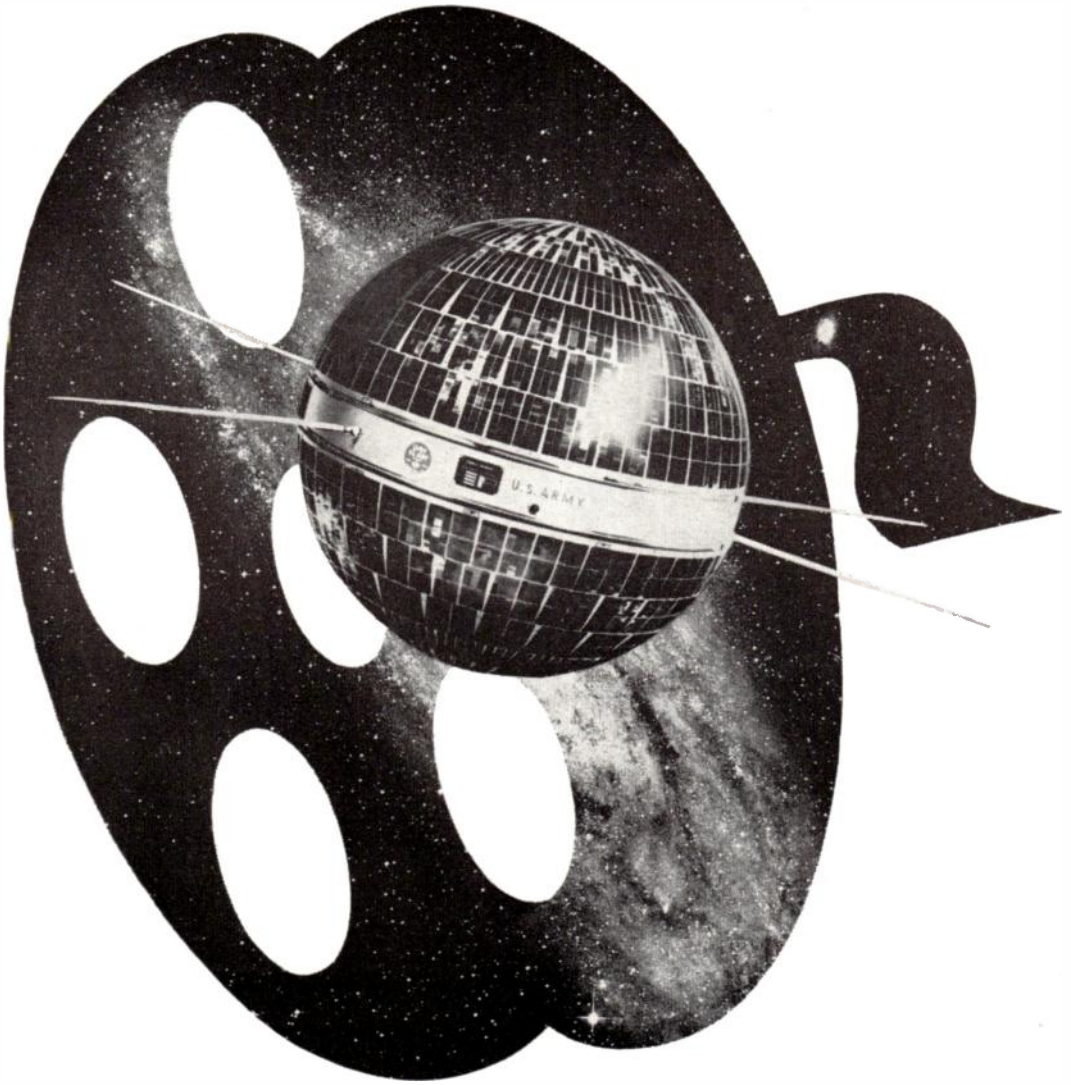
Applications are invited for appointments in the Department of Electrical Engineering. Candidates should be qualified, or have interests, in one or more of the following fields: Circuits, electronics, control systems, communication systems, information theory, computers and digital systems. Applicants should hold a doctorate in Electrical Engineering, or in an appropriate branch of Mechanical Engineering of Physics, or should have equivalent qualifications and experience. Duties include teaching at undergraduate and graduate levels, conducting and supervising research. Salary according to qualifications and experience. Minimum rates: Assistant Professor, $7020.00; Associate Professor, $9,000.00; Professor, $11,700.00. Additional stipends for summer research. Applications and inquiries should be addressed to The Chairman, Department of Electrical Engineering, McMaster University, Hamilton, Ontario.

TEACHERS

The Electrical Engineering Department of the City College of New York has several positions available on the teaching staff beginning September 1962. Rank and Salary commensurate with qualifications and experience. Opportunity for Graduate Study. Applicants must be present residents of the United State. Address inquiry to Prof. H. Taub, Dept. of Electrical Engineering, The City College, Convent Ave. at 139th Street, New York 31, New York.

Challenging opportunities exist in basic phy-

(Continued on page 108A)



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From the earliest plans to invade space, PHILCO Western Development Laboratories has played a vital role in satellite vehicle instrumentation, still but *part* of its contribution to space communications. From this newest electronics center on the San Francisco Peninsula comes a continuing flow of advanced missile tracking, range and data processing instrumentation.

Added research projects and growing programs assure *you* a long and rewarding career as a member of the PHILCO Western Development Laboratories. What you think and what you do can be unhampered and uninhibited. Personal recognition and advancement promptly follow performance, with monetary rewards to match. Northern California provides an affluent climate for living, as PHILCO Western Development Laboratories provides a stimulating climate for working. For information on careers in electronic engineering, please write Mr. W. E. Daly, Dept. R-2.

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin; U. S. citizenship or current transferable Department of Defense clearance required.

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Armour Research Foundation, one of the largest independent research organizations, offers exceptional opportunities in Chicago for qualified Electronics research engineers to work on a variety of industrial and military research projects in the fields of

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MEASUREMENT TECHNIQUES
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COMPONENT DEVELOPMENT

Minimum requirements include a B.S. degree in Electrical Engineering with above average academic grades. Research experience in one of the above fields is highly desirable, though not absolutely necessary.

Our environment combines the better features of academic and industrial research and offers you the opportunity to work on advanced research programs.

Staff members receive attractive salaries, up to 4 weeks vacation, generous insurance and retirement benefits, and tuition paid graduate study. Our Chicago location offers excellent cultural and recreational advantages. Please reply in confidence to Mr. Ron C. Seipp.

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Positions Open

(Continued from page 106A)

sical and electrical measuring techniques applied to nuclear, chemical, physical, and process variables. Static and dynamic conditions are investigated on the highest level, pressing the state-of-the-art. Typical fundamental detecting mechanisms and techniques encountered are: ionization and scintillation effects; thermoelectric and thermoresistive effects for temperature detection; variable reluctance, piezoelectric and magneto-strictive effect for fast response pressure and flow detection; piezoresistive effect for fast response pressure, strain, and acceleration detection; ultrasonic techniques for transient flow and density measurements; and microwave techniques for fluid conductivity and dielectric measurements.

We hold the responsibility for special instrumentation for four large test reactors (MTR, ETR, EOCR, and ATR), the four SPERT reactors, three critical facilities (RMF, ARMF, and ETRC), and the Chemical Processing Plant (CPP).

Openings exist for Electrical Engineers, Physicists, Engineering Physicists, and others who have several years experience and interest in this field.

Write: Personnel Administration, Phillips Petroleum Company, P.O. Box 2067-GH, Idaho Falls, Idaho.

ASSOCIATE PROFESSOR & ASSOCIATE DIRECTOR OF RESEARCH

Associate Professor and Associate Director of Research in Electrical Engineering. To have the

(Continued on page 112A)

'62 and YOU

What will it be like for you in '62? Information supplied by 1215 large and medium sized corporations in 47 states indicates that 26% more engineers will be recruited in '62 than in 1961.

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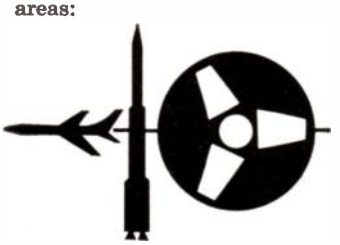
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Advanced development of digital computer techniques and solid state devices on navigational, tactical and training projects for aircraft, missile and aerospace simulator systems.

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Installation, maintenance, operation and instruction on company and competitor electronic equipments, including communications, computers, radar and simulator systems. Assignments in this country and abroad.

Our modern facilities are located in residential suburbs of Washington, D. C. and New York City, affording you a choice of city, suburban or country living. Advanced study under tuition refund may be conducted at several nearby universities.

For details, contact: Mr. A. C. Sugalski, Manager, Professional Staffing, Dept. A-16 at our Riverdale facility.

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No two are the same

And the molecules in any two elements are never the same either. At Cal Tech's Jet Propulsion Laboratory, these molecular differences will soon add up to a better understanding of what the Moon is made of.

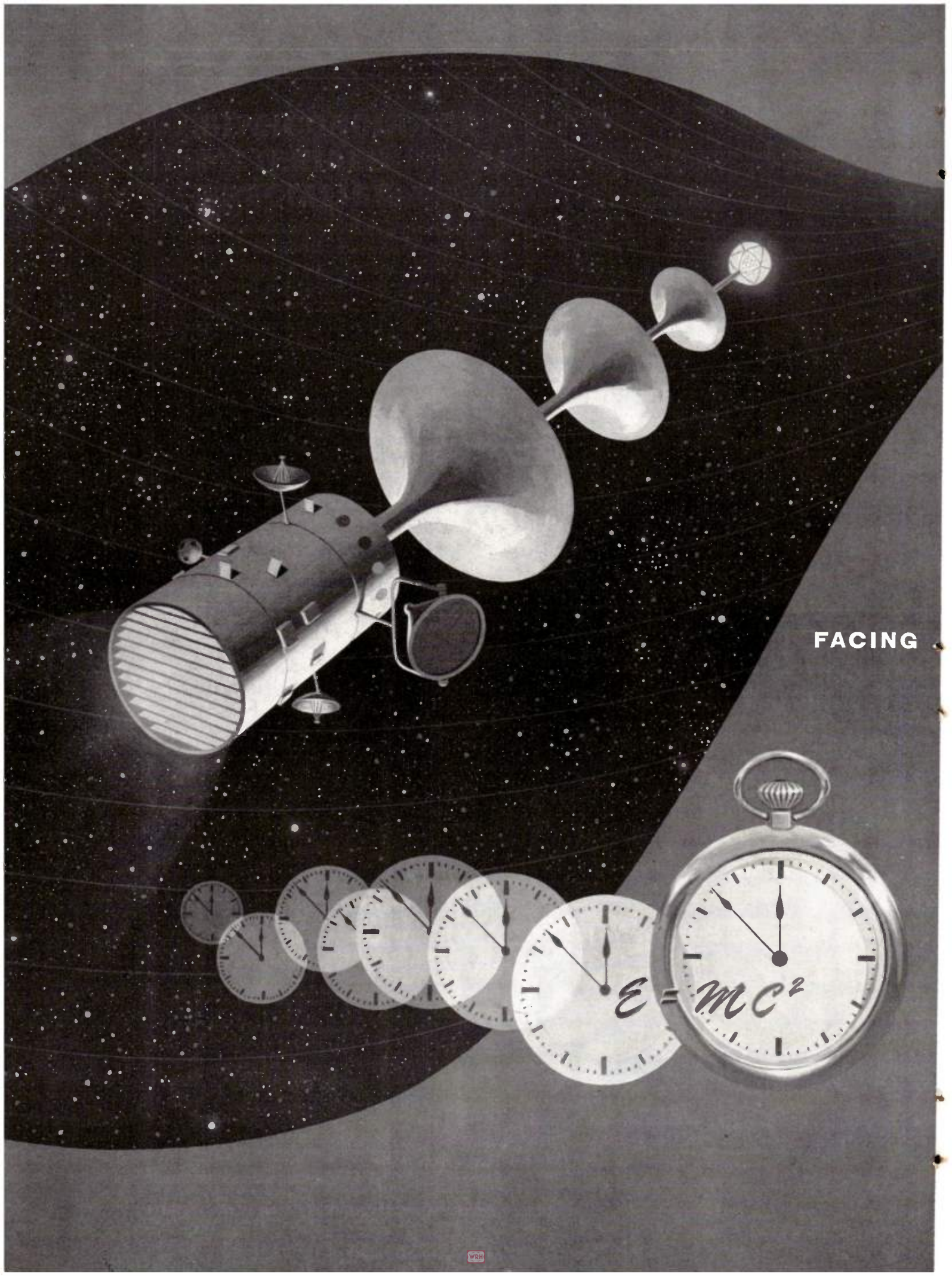
When the first Surveyor spacecraft soft-lands on the Moon in 1963, it will drill into and extract samples from the Moon's surface. After an electric oven decomposes these samples, their molecules in the form of burned-off gas will pass through a column to a detector. The detector response rate — and the amount of molecules — will be recorded and transmitted to earth.

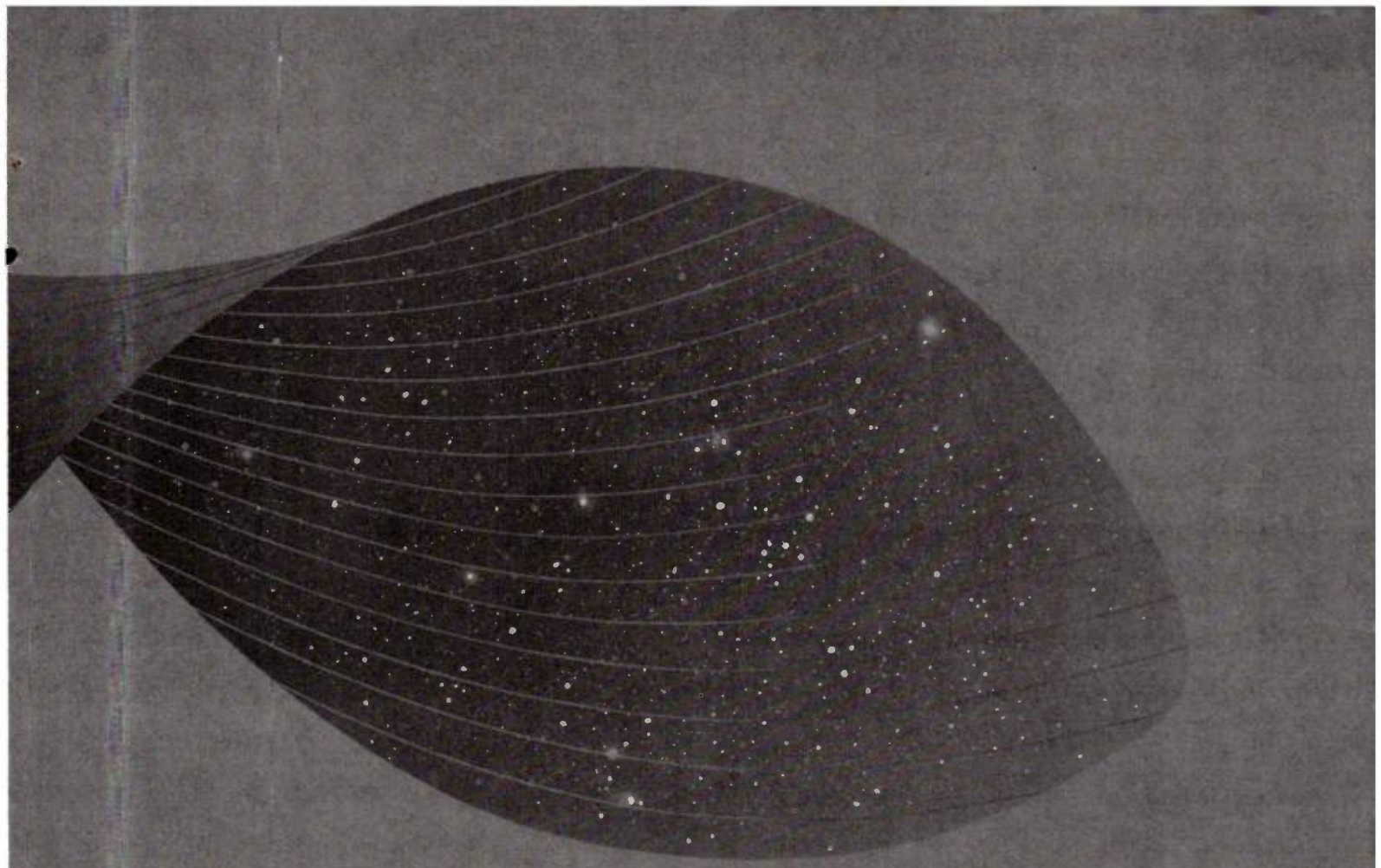
Since JPL scientists and engineers know what molecules cause what kinds of responses (fingerprints), they will then know what elements are on the Moon. Most important, they may detect pre-life molecules on our satellite — a first clue to life in outer space.

At JPL it takes many good, curious, searching, dedicated minds to determine what instruments will measure what on the Moon and planets. Many many minds that work as one. Minds such as yours, perhaps. Will you write us today? If you have a mind to?

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E = mc²



THE FOURTH DIMENSION IN PROPULSION DEVELOPMENT

Whether the universe has a "saddle shape," or any shape at all, is a matter of interesting conjecture. The matter of space travel, however, is the subject of intense experimentation. A nuclear/thermionic/ionic propulsion system, currently being studied at Lockheed Missiles & Space Company, might well become the power source for space vehicles.

Its design incorporates a nuclear reactor only one foot in diameter, generating heat at a temperature of 1850°K. This is transmitted to banks of thermionic generators, converting the heat directly into electrical energy for the ion beam motor which uses cesium vapor as a fuel. The entire system is designed without any moving parts, minimizing the possibility of failure.

Lockheed's investigation of propulsion covers a number of potential systems. They include: plasma, ionic, nuclear, unique concepts in chemical systems involving high-energy solid and liquid propellents, combined solid-liquid chemical systems. The fundamentals of magnetohydrodynamics, as they might eventually apply to propulsion systems, are also being examined. Just as thoroughly, Lockheed probes all missile and space disciplines in depth. The extensive facilities of the research and development laboratories—together with the opportunity of working with men who are acknowledged leaders in their fields—make association with Lockheed truly rewarding and satisfying.

Lockheed Missiles and Space Company in Sunnyvale and Palo Alto, on the beautiful San Francisco Peninsula, is an exciting and challenging place to work. For further information, write Research and Development Staff, Department M-24A, 599 North Mathilda Avenue, Sunnyvale, California. An Equal Opportunity Employer.

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APPLICATION AND DESIGN OF ANALOGUE AND DIGITAL COMPUTERS

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Positions Open

(Continued from page 108A)

responsibility for developing a research program including the securing of research contracts and the directing of research activity. Salary depending on experience and academic background. Write to Head, Electrical Engineering Department, South Dakota School of Mines and Technology, Rapid City, South Dakota.

ELECTRONICS ENGINEER

Ph.D. or equiv. exp.; background in servomechanisms, transistor circuitry, digital computers; to join medical group with aggressive experimental program in medical electronics. Long-term NIH support. N.Y. Area. Write Box 2067.

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Forward-looking faculty members in all phases of Engineering. Unusual opportunity for development of outstanding science-oriented engineering program in small-college atmosphere. New building to be constructed, new curriculum to be developed, desire to develop research program and work towards establishment of graduate degrees. Will establish core engineering program with interdisciplinary outlook but yet opportunity for specialization. Outstanding salaries available for highly qualified professors at all ranks. Ph.D. or top-notch M.S. Submit qualifications to A. T. Murphy, Director of Engineering, Pennsylvania Military College, Chester, Penna.

CHAIRMAN OF ELECTRONIC TECHNOLOGY

The position of Chairman of the Faculty of Electronic Technology is open at Arizona State University. Electronic Technology is devoted to the preparation of Engineering Technicians in a four-year program which leads to a Bachelor of Science degree. Interested persons holding a M.A., M.S., or Ph.D., having teaching, industrial and administrative experience in electronics are invited to correspond with the Division of Industrial Education, Arizona State University, Tempe, Arizona.

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Scholarships of $1,940.00 and assistantships of $600.00 are available for the Session 1962-3. Inquiries should be addressed to; The Chairman, Department of Electrical Engineering, McMaster University, Hamilton, Ontario.

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Unusual opportunity for scientist with background in electrical engineering or physics for teaching and research in solid state devices in eastern university. Opportunity includes possible association with research activities in nearby prominent, research oriented industrial center. Write Box 2068.

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(Continued on page 114A)

NCR

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The National Cash Register Company's Research and Development Division located in Dayton, Ohio has opportunity for experienced Senior Engineers and Scientists for the following positions:

NCR's efforts are expanding now to meet the needs of the future for business and government functions. The stable nature of our Company provides an atmosphere for long range planning of personal goals and scientific achievement.

- **DEVELOPMENT PLANNING STAFF**—to conduct studies of electronic equipment requirements for AERO/SPACE projects with background in Microwave Equipment, Digital Computers, Electro-Optical Devices, Human Engineering of Controls and Displays, or Navigation and Guidance Systems. Prefer MS in EE or Physics and 5-7 years experience in Military Development.
- **ENGINEER ANALYST**—to investigate multipath phenomenon relative high frequency communications. BS. EE and some graduate work in mathematics. Prefer experience in information theory and communications.
- **ELECTRONIC ENGINEERS** — BS, EE, 2-5 years in military electronics and familiarity with ground based or airborne equipment development.
- **CIRCUIT ENGINEERS**—BS, EE, 3-5 years in design of electronic equipment using solid state or vacuum tube circuits.
- **LOGIC ENGINEERS**—BS, EE, 2-4 years in design of digital logic circuits.
- **COMPONENT ENGINEERS** — BS, EE, 1-3 years in testing and evaluation of electronic components.
- **COMMUNICATION SYSTEM ENGINEERS**—BS, EE, 5-7 years in the high frequency communication area. Experience in propagation techniques.

To be considered for one of these challenging positions, send resume to:

Mr. T. F. Wade, Technical Placement, The National Cash Register Company, Main and K Streets, Dayton 9 Ohio.

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I want a "Whatcha-ma-call-it?"

It's a "thinga-ma-jig that . . . . . "

How often have you struggled to remember the name of a component or electronic item. . . . Just could not think quickly what it is called?

YOU CAN FIND IT IN THE IRE DIRECTORY!

because

(1) The IRE Directory classifies products by purpose and use.

(2) Its listings are fundamental—the way an engineer thinks.

(3) "Terminology" is cross indexed in the pink pages —condensed, simple, not mixed in with firm names.

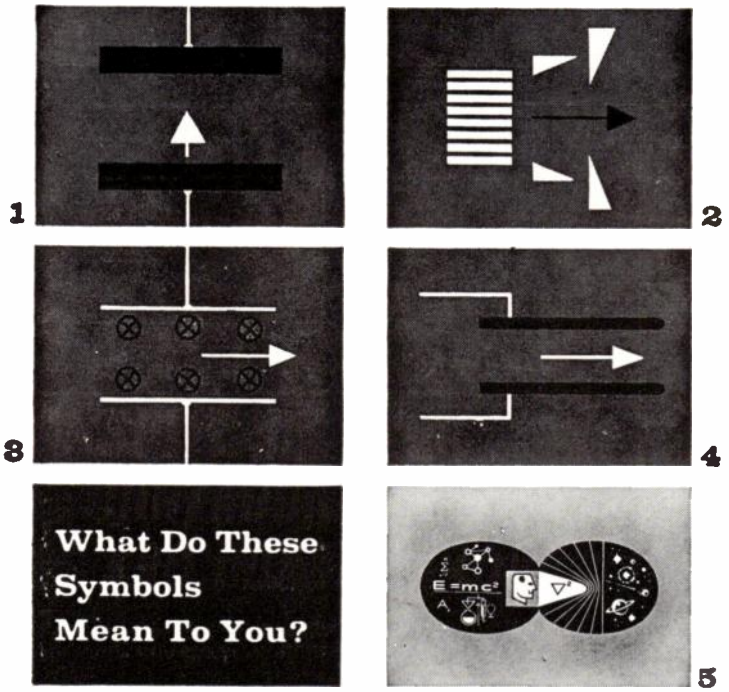
(4) Ads face listings thus helping to identify products by actual pictures.

(5) Product code numbers reduce complex and duplicate listings, saving you "searching" time and effort.

Study it . . . save it . . . work it.

THE INSTITUTE OF RADIO ENGINEERS

1 East 79th Street, New York 21, N.Y.



To the Plasma Physics Group at Allison they represent some of our areas of current, challenging investigation—(1) Thermionic Energy Conversion; (2) Electrostatic Propulsion; (3) MFD Power Generation; (4) Electromagnetic Propulsion; and (5) the guiding theme at Allison, "Energy Conversion Is Our Business."

Long-range objective of this group is the development of practical energy conversion devices in each area, for such applications as direct nuclear conversion systems . . . plasma acceleration for space flight propulsion . . . mobile compact reactors and energy depots.

To implement these expanding projects, we need additional personnel—both theoretical and experimental—in the following areas:

PLASMA THEORISTS (MS or PhD—Physics or equivalent)—Perform theoretical studies of the physical processes present in the formation, confinement, acceleration, and deceleration of plasmas. Strong background in mathematics, classical electricity and magnetism, kinetic theory, atomic behavior and magnetofluidmechanics.

PLASMA EXPERIMENTALISTS (MS or PhD—Physics or EE)—Perform experimental research in the areas of magnetofluidmechanics and thermionic energy conversion. Familiarity with diagnostic tools such as magnetic and electric probes, spectroscope, photomultiplier, etc. Apply pulsed and steady state instrumentation techniques. Ability to interpret experimental data. Industrial experience desirable.

PHYSICAL ELECTRONICS SCIENTISTS (MS or PhD—Physics, ME or EE)—Perform theoretical and experimental studies on thermionic energy conversion. Familiarity with thermionic emission, space charge neutralization, adsorption theory, and high temperature materials. Experience in heat transfer and mechanical design problems as related to converter design desirable. Knowledge of nuclear, solar, and chemical heat sources useful.

RESEARCH ENGINEERS (BS, ME/EE or equivalent experience)—Construct experimental and prototype test devices for energy conversion or propulsion programs. Perform tests and evaluate data on experimental devices. Should have knowledge of high temperature materials and electronic tube fabrication processes.

If you're interested in tackling challenging, new problems in applied plasma physics—and your background and experience qualifies you for a place on the progressive Allison-GM team—let us hear from you. Send your resume, or write: Mr. V. A. Rhodes, Scientific & Professional Placement, Dept. 901, Allison Division, General Motors Corporation, Indianapolis 6, Indiana.

An equal opportunity employer.

ALLISON DIVISION · General Motors Corporation

Indianapolis, Indiana

Positions Open

(Continued from page 112A)

CRITICAL VACANCIES AT GRIFFISS AIR FORCE BASE, N. Y.

Accountant	GS-7	$5355 pa.
Electronic Engineer	GS-7	$6345 pa.
Supply Requirements Offr.	GS-7	$5355 pa.
Cost Accountant	GS-9	$6435 pa.
Management Analyst	GS-9	$6435 pa.
Electrical Engineer	GS-9	$6435 pa.
Research Psychologist (Physiological, Experimental & Engineering)	GS-9	$6435 pa.
General Engineer	GS-11	$7560 pa.
Industrial Engineer	GS-11	$7560 pa.
Electrical Engineer	GS-11	$7560 pa.
Electronic Engineer	GS-12	$8955 pa.
General Engineer (Systems)	GS-12	$8955 pa.
Mechanical Engineer	GS-12	$8955 pa.

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 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.

ENGINEER—MANAGER

Age 37, P.E., 11 yrs. Electro-Mechl. Mfg. as Chief Engineer, Manager of Eng., General Manager, Vice-President, including Overseas Assignments. Air Force Vet. Location open including Overseas. Box 3962 W.

SALES REPRESENTATIVE

Psysicist and Engineer with twelve years experience, four in sales, is desirous of forming sales representative organization selling instruments and components in the Metropolitan New York area. Age 36, married. Would be willing to consider participation by manufacturers who prefer to make an investment. Box 3963 W.

DESIGN ENGINEER—MANAGER

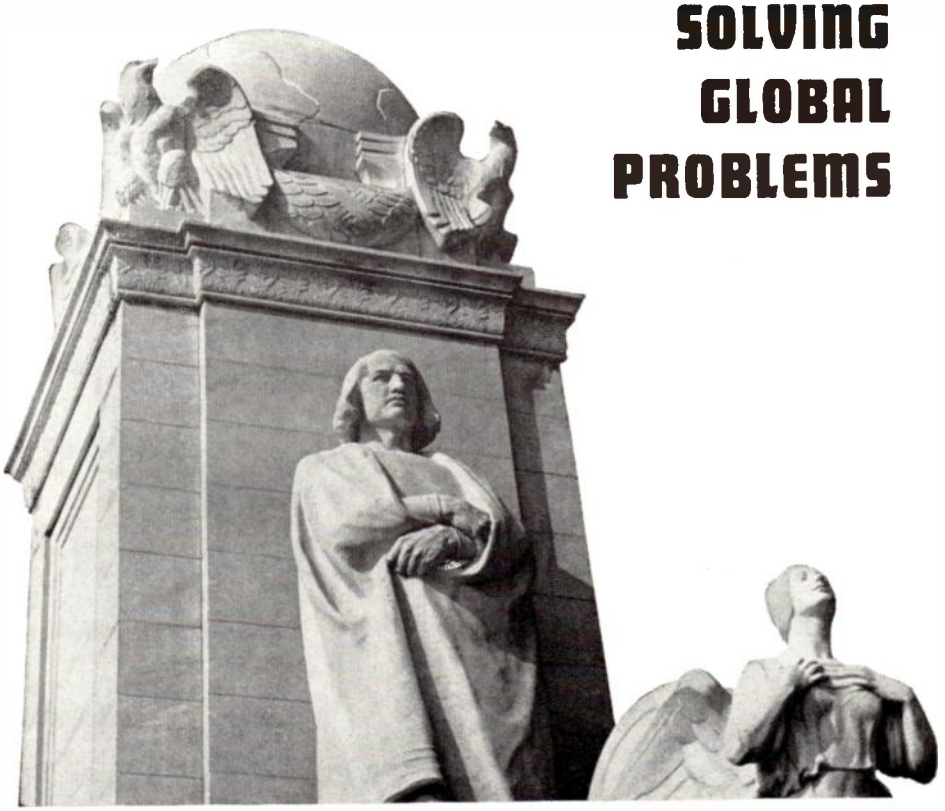
B.S.E.E. in 1941, also graduate work in specialty. 15 years experience in design and management of antennas, microwave assemblies, and military electronics systems, including ECM, aircraft C & N, and radar. Eight years of large corporation management experience. Prefer suburban area in Mid-Atlantic region. Box 3964 W.

SALES MANAGER

B.S.E.E., 8 yrs. experience in Electronic Component sales plus two years experience as

(Continued on page 116A)

SOLVING GLOBAL PROBLEMS



EXTENDING AMERICAN CAPABILITIES FOR THE NATION'S DECISION-MAKERS

Columbus made decisions that changed the world. Today's decision-makers have a responsibility every bit as awesome. The Research Analysis Corporation assists the planners of the United States Army by producing through operations research and systems analysis the recommendations and solutions pertinent to problems requiring major Army decisions.

RAC's expanding multi-disciplinary research programs have created a need for physicists, engineers, mathematicians, computer programmers, political scientists, and other social scientists. Their attributes should include a high degree of creativity, an interest in problem-solving, and a willingness to cross disciplinary boundaries in a mixed-team approach to scientific research.

RAC offers you an exciting professional career and the opportunity to make essential contributions to our national purpose. For additional information, please contact ... John G. Burke, Research Personnel Officer

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Increased technical responsibilities in the field of range measurements have required the creation of new positions at the Lincoln Laboratory. We invite inquiries from senior members of the scientific community interested in participating with us in solving problems of the greatest urgency in the defense of the nation.

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RE-ENTRY PHYSICS

PENETRATION AIDS DEVELOPMENT

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- Strategic Communications
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NEW RADAR TECHNIQUES

SYSTEM ANALYSIS

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Massachusetts Institute of Technology
BOX 16
LEXINGTON 73, MASSACHUSETTS



Positions Wanted

By Armed Forces Veterans

(Continued from page 114A)

Project Engineer for Component Manufacturer, Successful record in increasing sales. Tripled sales in present territory in three years. Box 3965 W.

PRODUCTION MANAGER

Mgt. Executive, 39, M.I.E., B.M.E., 15 years varied experience. E1/Mech Industries, seeking plant production Manager level opportunity with/without investment. Tau Beta Pi; Pi Tau Sigma. Box 3972 W.

PATENT ATTORNEY

Electronics and mechanical arts, 4 years corporate and 1 year law firm experience, writing and prosecuting applications, interference, infringement studies and licensing. 6 years engineering experience including transitor circuitry and aircraft instruments. Interested in position in New York City area. Box 3973 W.

ELECTRONICS TECHNICIAN

Graduate RCA Institute V7 course, Age 24, Single. Willing to relocate in United States or abroad. Member IRE, Resume upon request. Box 3974 W.

FREE LANCE TECHNICAL WRITER

Experience and degrees in electronic engineering and physics. Associate in St. Paul and Milwaukee. Box 3975 W.

FREE LANCE TECHNICAL WRITER

Experience and degrees in electronic engineering and physics. Associate in St. Paul and Milwaukee. Box 3976 W.

ENGINEER/PILOT

BSEE '60. Age 29. Former USAF pilot, presently ANG pilot, hold commercial license (multiengine). 16 months field engineering experience with computer type checkout equipment for missle systems. Speak, write and read Spanish fluently. Seeks challenging position in Spanish speaking country or Europe. Box 3977 W.

SALES REPRESENTATIVE—WESTERN CANADA

Professional Electronics Engineer with 12 years experience in a wide range of Communications and Medical-Industrial Electronics equipment and systems design and good Sales background, seeks appointment as Manufacturers Representative for a good line of electronics equipment or instruments. Box 3983 W.

LINGUIST-ENGINEERING PHYSICIST

Would enjoy coordinating the research efforts of a laboratory with related research done in other countries—to avoid duplication and accelerate results. Training includes: mathematics, management, manufacturing, engineering, graphic arts, Russian, German, French, etc. Interested in automatic language translation. (home: Philadelphia; age: 41) Box 3982 W.

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SOUTH
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If you are interested and believe that you can contribute, make your appointment today.

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Supervisor of Scientific Employment
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We promise you a reply within one week.

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Melpar offers excellent opportunity for semiconductor materials physicist, physical chemist, or semiconductor metallurgist to take charge of programs on new semiconductor compound survey and epitaxial growth techniques. This is an excellent growth opportunity for creative work in this field, under ideal conditions. Publication is encouraged. Salary open and dependent upon prior experience.

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PHILLIPS 66

PHILLIPS PETROLEUM COMPANY
ATOMIC ENERGY DIVISION
Announces
new senior staff positions in the
INSTRUMENT DEVELOPMENT BRANCH
located at the
NATIONAL REACTOR TESTING STATION, IDAHO

OPPORTUNITY

Expansion of the newly organized Instrument Development Branch has created several new senior positions. This branch is responsible for the research and development activities in the field of nuclear instrumentation concepts, systems and components. The present day "state of the art" has not produced the techniques required to solve existing problems. Consequently, additional theoretical and experimental effort must be brought to bear in this technical area.

THE REQUIREMENTS

The senior staff personnel being sought for these positions must meet the following minimum requirements.
- Advanced degree in Electrical Engineering or Physics.
- Five years experience in research and development activities.
- Working knowledge in the fields of data system analysis, statistical information analysis or instrumentation techniques as applied to nuclear and reactor physics.
- A high degree of initiative and imagination.

EMPLOYER

The Atomic Energy Division of Phillips Petroleum Company has been given by the Atomic Energy Commission the operational responsibility for the Materials Testing Reactor (MTR), Engineering Test Reactor (ETR), and the Chemical Processing Plant (CPP), conduction of the Special Power Excursion Reactor Tests (SPERT) programs; the conceptual designs and operational responsibility of the Experimental Organic Cooled Reactor (EOCR) and the proposed forty million dollar Advanced Testing Reactor (ATR); in addition to supporting research and development activities.

Please forward inquiries to:
PHILLIPS PETROLEUM COMPANY
Personnel Administration
P.O. Box 2067-GN, Idaho Falls, Idaho

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ACT NOW! Address your resume to
Mr. Louis A. Kay
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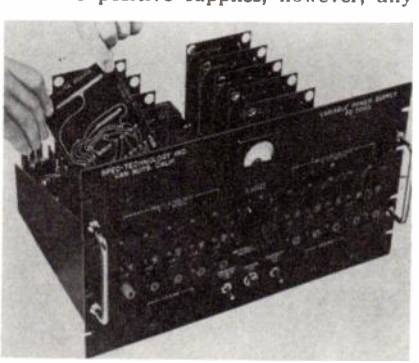
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 90A)

Variable Power Supply

Spec-Technology, Inc., subsidiary of Speck Electronics, Inc., 13901 Saticoy St., Van Nuys, Calif., announces the availability of a new solid state variable DC power supply designated as the PS 5003. This unit consists of ten independent power supplies, each from 0.5 to 30 volts at ½ ampere. All supplies have a common ground, each is fused separately and has its own regulator channel selector which is located centrally on panel front for independent monitoring of voltage and current. The standard PS 5003 has five negative and five positive supplies, however, any



(Continued on page 120A)

BUILDING THE FUTURE IS A BIG JOB!



The radio-electronic engineers who form the membership of the Institute of Radio Engineers must remain years ahead of actual production, in order to pave the way for the products of tomorrow, through research and development today.

The special issues of "Proceedings of the IRE" help these men transform the theory of today into the production lines of tomorrow. Pictured above are the October 1951 Color Television Issue; November 1952 Transistor Issue; and January 1953 UHF Issue (all now out of print).

Special issues on special subjects are more than usually helpful, but every issue of "Proceedings of the IRE" is filled with facts and figures to keep you far ahead of the field!

HIGH STANDARDS



The creative work of research personnel is the basic product of the Electro-Physics Laboratories. Therefore, men willing to devote their full talents and energy to achieve rapid and economical scientific progress are essential. **THE PROBLEM** The useful applications of long range propagation of electromagnetic radiation in the VLF to VHF portion of the spectrum. These applications are concerned with communications and the measurements of natural and man-made disturbances of the geophysical environment. **THE MAN** He is imaginative in the application of classical theory to today's problems. He accepts challenging technical problems with the characteristically positive approach of the successful investigator. He is enthusiastic about the pursuit of useful knowledge along uncharted pathways. If he detests mediocrity and gadgeteering . . . if he seeks the best possible solution to the fundamental problems regardless of the approach . . . then this is our man. Staff appointments are now available for personnel having, or eligible for, security clearances. For additional details, please write to:

This 100-foot high, HF helical antenna, designed and built by the Electro-Physics Laboratories, is one of the many unusual antennas used to increase our capabilities in experimental investigation of ionospheric phenomena.

Mr. William T. Whelan, Director of Research and Development

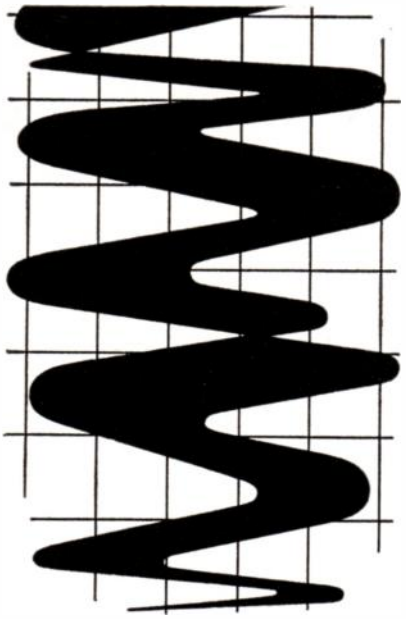
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ACF INDUSTRIES



creative instrument engineers

We are confident the experienced, really creative engineer will like the frequently changing application of his talents to conceiving, designing, and actually building "custom" instruments and "custom" systems for the gamut of basic and applied research experiments underway or planned at Argonne. Flexibility, versatility, a mature and sometimes "way out" approach to problems are essential. We need *experienced men.* A BSEE degree and a minimum of four years experience in the design of electronic instruments in the nuclear fields are required. *Our salaries are competitive, our benefits are unexcelled, our address is...*

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NEWS New Products

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(Continued from page 118A)

combination of negative and positive ratios are available up to ten upon customer specification.

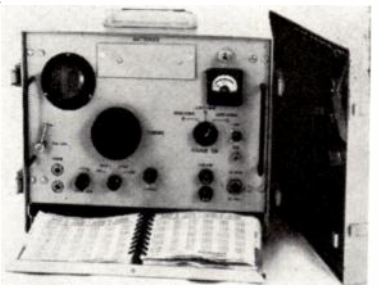
Circuitry consists of printed modules, easily removed for maintenance. PS 5003 is available with 50 millivolts regulation and 2 millivolt peak to peak ripple. Input is 115 vdc.

Unit can be either rack and panel mounted (PS 5003) or bench mounted (PS 5004) according to requirement. For further information, request Bulletin STI-504.

Portable Frequency Meter

Development of a new portable frequency meter, featuring a 25% reduction in size and weight, transistorized operation and a new film calibrated readout, was announced by Stephen D. Lavoie, president of **Lavoie Laboratories, Inc.,** Morganville, N. J.

Designated the Lavoie Portable Frequency Meter, LA-32, the new self-contained instrument will measure frequencies in the range of 125 kc to 1000 mc to a specified accuracy of 0.01% from −4 to −125°F, and can also be used as a signal generator for testing and calibrating radio equipment.



The LA-32 includes a fully automatic factory calibrated direct frequency readout in a film presentation which eliminates the difficulty of reading a closely calibrated vernier dial. Providing an illuminated and magnified readout of the measured frequency, the new system eliminates the possibility of initial calibration or readout error by presenting only the frequency band in use as visible to the operator, while also improving the usable accuracy of the instrument.

Meeting all of the requirements of MIL-F-13569A, the newly developed instruments weigh 28 lbs. with batteries and power supply, compared to previous units, the over-all package of which weighed 67 lbs.

The unit can be powered by a standard flashlight battery or from an AC source of 100–250 V, 50–400 cps. The switch over is automatic. Because of low power consumption, the unit can be operated 80 hours on one set of batteries.

(Continued on page 122A)

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companies pay interview, relocation and agency expenses

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(a division of the
Permanent Employment Agency)
825 San Antonio Rd.
Palo Alto, Calif.

ADVANCED SYSTEMS DEVELOPMENT

Emerging now from three years' intensive effort at International Electric is a computer-based communication system that equals the state-of-the-art. This system was made possible by our systems management capability, which includes the design, development, production monitoring and installation of advanced electronic systems. Our continuing progress will be determined by our ability to advance the state-of-the-art in our design and development of future systems.

This requires the application of creativity and technical experience by engineers, mathematicians and physicists. To our recently organized Development Division we seek to add *Development Specialists* with at least six years' association with large projects. They will evolve requirements for systems users, working in such areas as air traffic control, ASW, satellite control and command and control systems, and in such projects as information retrieval, man/machine communications and advanced computer utilization.

Digital Equipment Engineers are needed to translate systems requirements into system hardware. We seek E.E.'s with at least five years' experience to work in such areas as logical design; core, drum and tape memories; peripheral equipment; and large-scale switching center techniques.

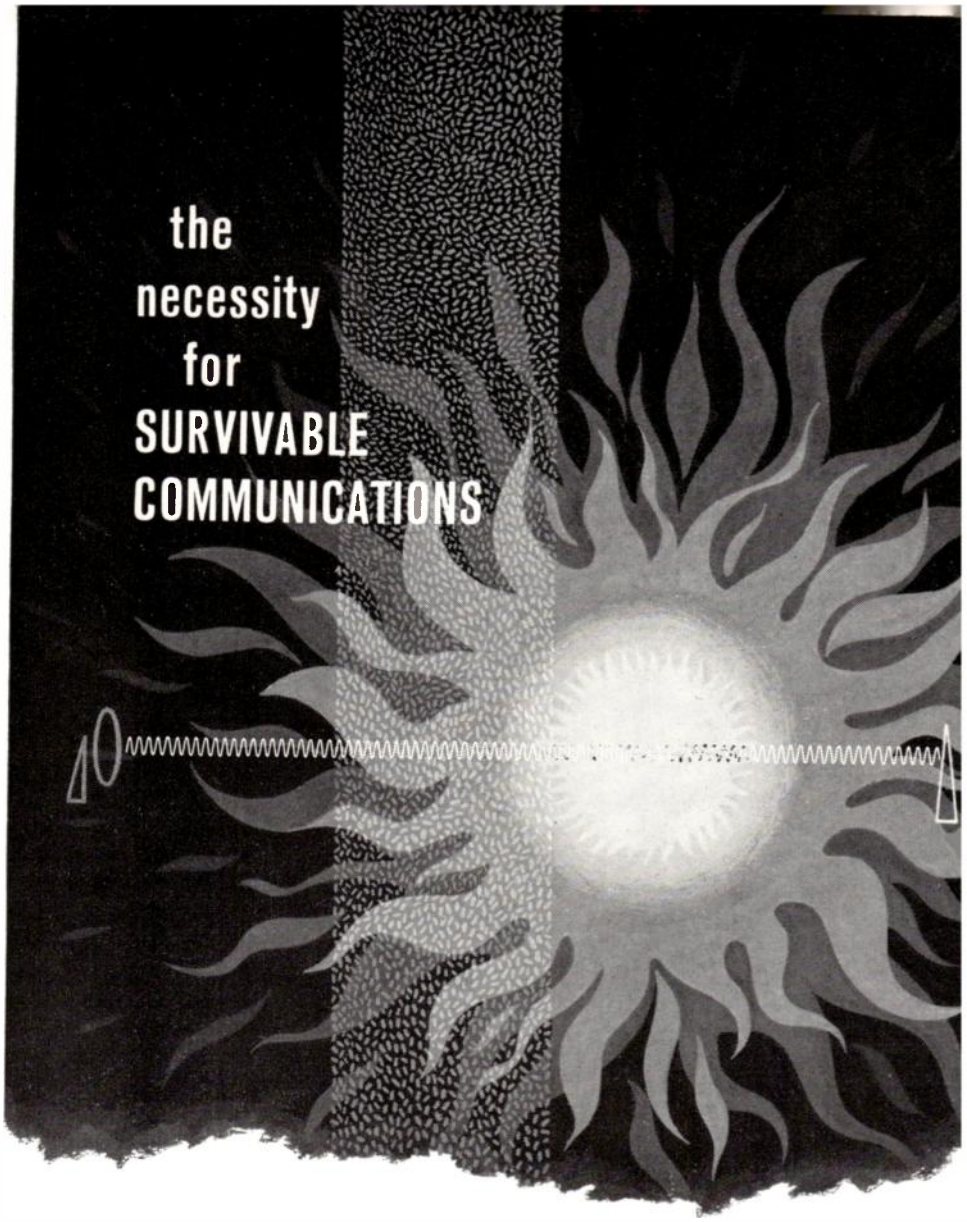
If you have the experience and interest in either of these programs, and wish to associate yourself with the professional challenges offered by them, please send resume to Manager of Technical Staffing, Dept. PO.

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A key factor in the survivability of our nation is the systems reliability of our weapons command and control.
It is imperative that secure communications be maintained in severely disturbed environments. The scientists and engineers at the Amherst Laboratories are engaged in a continuing program to provide these capabilities for our nation's defense.

STAFF AND MANAGEMENT CAREER OPPORTUNITIES
exist for scientists and engineers with leadership qualities and 5-10 years of experience in command and control systems including low frequency, high frequency and scatter systems. Advanced knowledge of communications techniques, including optimum detection, anti-jam and wide band techniques is desired. Confidential inquiries are invited. Please forward resume to Mr. H. L. Ackerman, Professional Employment,

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SYLVANIA **ELECTRONIC SYSTEMS**
Government Systems Management
for **GENERAL TELEPHONE & ELECTRONICS**

GENERAL SYSTEM

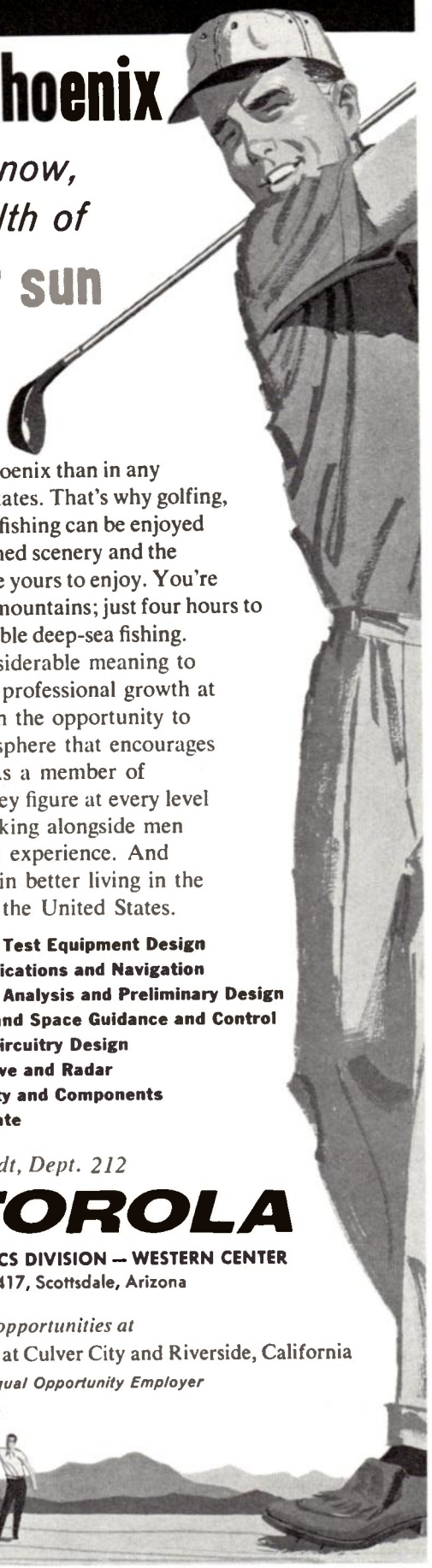
"Shrinking the Universe . . . through Communications"

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are awaiting you now, along with a wealth of

warm winter sun



The sun spends more time in Phoenix than in any other major city in the United States. That's why golfing, gardening, picnics, boating and fishing can be enjoyed around the calendar. World-famed scenery and the freedom of wide open spaces are yours to enjoy. You're 90 minutes from pine-forested mountains; just four hours to Mexico; six hours to incomparable deep-sea fishing.

These advantages add considerable meaning to opportunities for personal and professional growth at Motorola. Here, you'll be given the opportunity to express your ideas in an atmosphere that encourages initiative and independence. As a member of a project team, you become a key figure at every level of creative engineering — working alongside men with proven ability and broad experience. And in addition you'll get a bonus in better living in the sunniest, healthiest climate in the United States.

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- **Systems Test Equipment Design**
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- **Digital Circuitry Design**
- **Microwave and Radar**
- **Reliability and Components**
- **Solid State**

Write Phil Nienstedt, Dept. 212

MOTOROLA

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P.O. Box 1417, Scottsdale, Arizona

Motorola also offers opportunities at
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NEWS New Products

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(Continued from page 120A)

Additional characteristics include: greater electrical and mechanical stability through design, minimum drift (not over 0.0025% in 5 minutes) and resetability, 0.005%. The units measure 11¾" deep, nearly 13" wide and 11¼" high.

Wide-Range Pulse Generator

A compact, versatile pulse generator, Type 1217-B Pulse Generator, offering a wide range of pulse durations and repetition rates, has been announced by the General Radio Co., West Concord, Mass.



Pulses are either generated internally at pulse repetition rates from 2.5 to 500,000 cps—a frequency range comparable to that of most general-purpose oscilloscopes

(Continued on page 124A)

TROPOSPHERIC SCATTER DESIGN ENGINEERS

Collins Radio Company, pioneer in the development of tropospheric scatter communications and a leading supplier of scatter equipment and systems, has several attractive openings for engineers experienced in communication equipment design.

These positions are in the design of tropospheric scatter equipment, parametric amplifiers and FM detection devices. Those selected will work in Collins' new research and development facility, located near Dallas.

Minimum requirements include a degree in electrical engineering and at least two years' design experience in scatter, microwave or UHF equipment. These are minimum requirements; more responsible positions are available for those qualified.

Send resume, in confidence, to Ben Jeffries, Collins Radio Company, Dallas, Texas

COLLINS

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USAF/AWCS-412-L Consisting of a closely coordinated network of data acquisition stations, data processing and display centers and weapon bases, **Air Weapons Control System 412-L** provides the tools for effective and flexible air space management, continent-wide or in single point defense. Vital detection and tracking information is supplied automatically to human decision-makers within seconds. Effective direction of both manned and unmanned weapons, including return of manned aircraft to base, is a system function.



412-L STAFF EXPANDING

as program moves ahead in advanced development phase

If you are one of the fast-growing group of scientists and engineers who recognizes the unequaled career advantages and satisfactions to be gained through work on very large-scale, complex systems, we urge you to scan this abbreviated list of current openings on 412-L. If your background, experience or interests are represented, write us today. We promise you a careful review of your qualifications and a prompt reply.

COMMUNICATION SYSTEMS

Training in information theory and switching theory; broad experience in both RF and wire communication systems design, especially automatic subscriber and dial systems; to determine and specify characteristics for intra- and inter-site command and control complexes.

OPERATIONS ANALYSIS

Involves synthesis/analysis of computer oriented weapons control systems (ground/air environments) from an operational standpoint. Estimates system (personnel & equipment) capabilities; derives and evaluates system procedures using analyses and computer simulation as tools. BS/MS – minimum 3 years experience.

EQUIPMENT EVALUATION

Solving man-machine problems, evaluating alternative components, displays, or techniques, devise simulators. Advanced degree in experimental psychology required.

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SYSTEM EQUIPMENT ANALYSIS

Performs analysis on system-equipments and their functions; derives criteria and parameters to meet operational requirements. Work in areas of signal processing, detection, tracking and digital communications. Requires BS/MS/EE/Physics/Math—knowledge of simulation techniques, feedback theory, probability and information coding desirable. 3 years experience.

WEAPONS INTEGRATION

Determines and specifies optimum weapon utilization. Specifies weapon interfaces to equipment designers to insure system compatibility. Support / provide analytical inputs relative to weapons capabilities; analytical studies to assure optimum weapon employment within specific computer capabilities. BS/MS. 2 years manned/unmanned weapons system experience.

INFORMATION PROCESSING & DISPLAY SYSTEMS

Training in information and statistical theory; broad experience in design and integration of tactical and strategic data processing systems; to define system parameters, identify interface problems, perform math analyses of control loops as applied to real-time processing and display systems.

TECHNICAL WRITERS

Creation of systems manuals, engineering reports, proposals; required is at least 2 years of technical education with minimum of 5 years in engineering writing. Knowledge of graphic arts techniques would be helpful.

PROGRAMMING ANALYSIS

Degree required. Math background; 2 years experience on medium/large-scale digital computers solving real-time military operational problems. Activity ranges from problem analysis to program development, including flow diagramming, coding and debugging in the utility control, simulation and operation areas.

ANTENNA & MICROWAVE COMPONENTS

BSEE, advanced degree preferred; experience with high power phenomenon, antenna development, RF plumbing, modulators, transmitters and receivers.

APPLIED MATHEMATICS

Performs mathematical analyses on all phases of ground electronic control systems. Utilizes probability theory, numerical analysis, complex variables; knowledge and utilization of analog/digital computers desirable. BS/MS Math required; 2 years experience on applied math problems in engineering field.

Write in confidence to Mr. P. W. Christos, Div. 53-MB
DEFENSE SYSTEMS DEPARTMENT (A Department of the Defense Electronics Division),
General Electric, Northern Lights Office Building—Syracuse, New York

GENERAL ELECTRIC



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(Continued from page 122A)

—or triggered by external signals from dc to 1 mc. Main output pulses are simultaneous positive and negative 40-ma pulses, dc-coupled. Amplitude is adjustable, to a maximum of 40 v into 1000 ohms.

Duration of the main pulse is adjustable from 0.1 μsec to 1 sec, in seven decade ranges. Each decade range is covered by the same 5½ inches of a precisely calibrated dial; on the lowest range, this 5½-inch span covers but 1 μsec.

The rise time of the main pulse depends on the termination used. It is less than 20 nsec for all transitions into 50- and 100-ohm terminated systems, and depends only on load capacitance at higher impedances; e.g., 70–80 nsec into a 15-pf probe at the 40-v, 1000-ohm output level. No overshoot is present when the pulse is capacitively terminated at high impedance.

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Additionally, the pulser can be used as a simple time-delay generator over its range of available pulse durations. The delayed pulse—coincident with the end of the main pulse—is a 5-v, 100-nsec negative-going transition, followed by a 5-v positive-going transition to reset the input circuits of another pulse generator. Thus, two or more Pulse Generators can be tied together as a multiple delay generator.

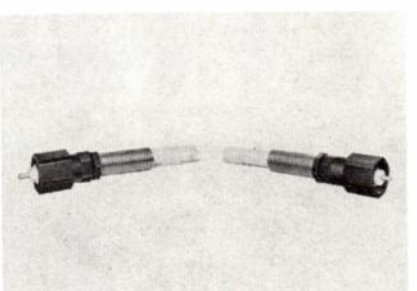
The circuit combines the best features of tubes and transistors, using each where it performs best. The tubes are operated at no more than 50 percent of maximum plate dissipation. This safety factor, plus a well-ventilated layout and thermal separation of the tubes and transistors, ensures trouble-free operation.

Power input for the generator is 300 v at 55 ma, and 6.3 v at 3 amperes. Any G-R Unit Power Supply can be used.

The over-all dimensions are 9½″ w × 5¾″ h × 6½″ d.

Price of the Type 1217-B Pulse Generator is $250.00, f.o.b., West Concord, Mass.

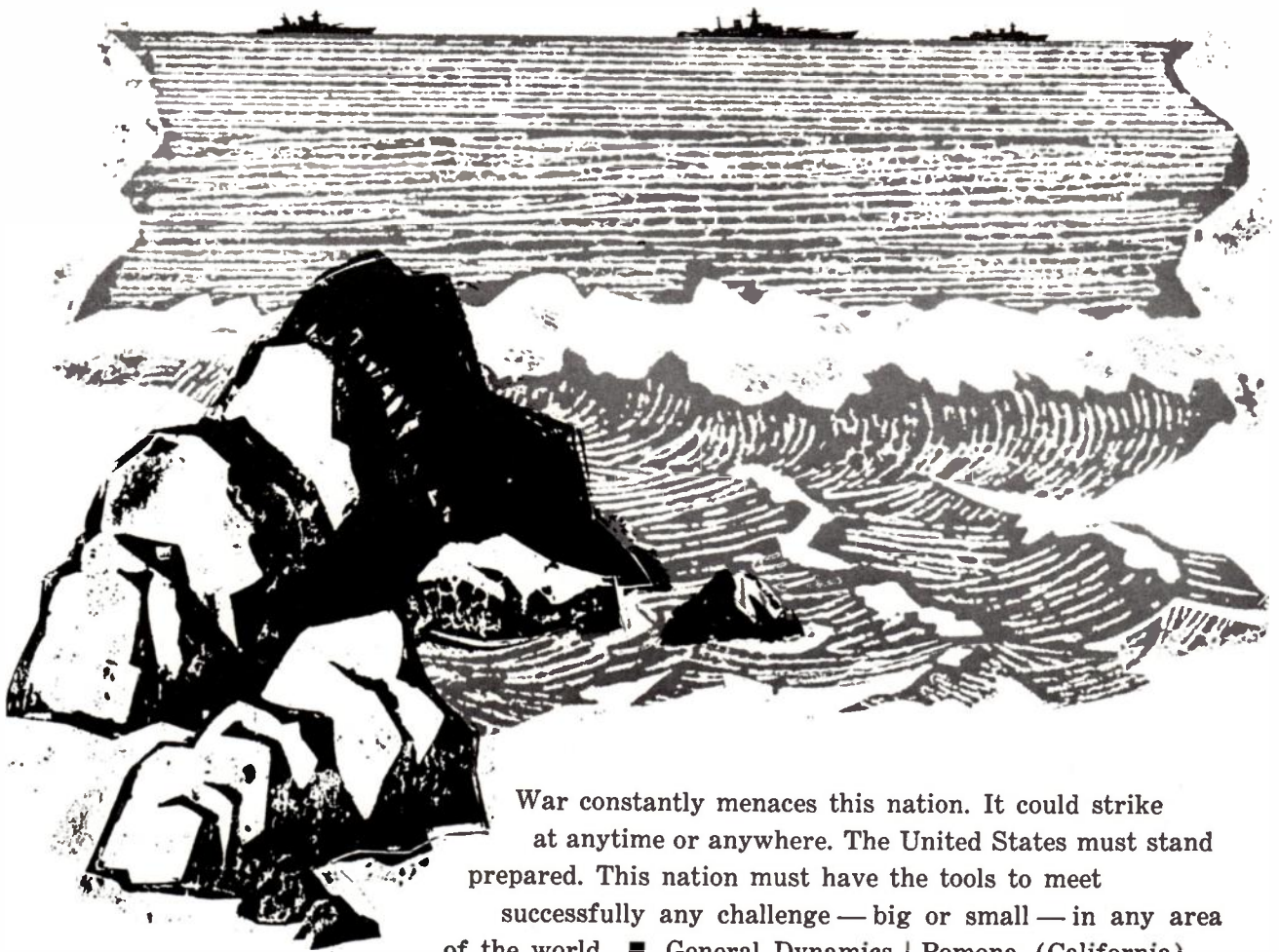
Coaxial Plug



The ALLT, a lightweight aluminum coaxial plug developed by Cannon Electric Co., 3208 Humboldt St., Los Angeles 31,

(Continued on page 126A)

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(Continued from page 124A)

Calif., for use with RG-209/U cables, is said to embody the highest level of performance yet achieved in RF coaxial plugs. The plug is capable of continuous operation at 1000 watts CW over a range of 0.03–5.0 kmc sea level or up to an altitude of 70,000 feet. It remains sealed even while unmated, and the center contact is capable of withstanding an axial pull of 150 pounds. Also, the ALLT will resist 170 hours of salt spray; 10 cycles of 240 hours of humidity and elevated temperatures; 12 hours of sand and dust; and 6 cycles of thermocycling in a temperature range of −55°C to +95°C. The plug is capable of withstanding 8 hours of vibration-altitude with RF power applied at temperatures from −65°C to +95°C—again at sea level or up to 70,000 feet. The cable assemblies consisting of two straight plugs on 50 inches of cable have no greater VSWR than 1.2:1, and similar assemblies consisting of two right angle plugs and cable have no greater VSWR than 1.25:1. The individual plug and its cable junction exhibit extremely low reflections as the product of all the possible discontinuities in measuring the VSWR does not exceed 1.2:1.

Thermionic Converter



Thermo Electron Engineering Corp., 85 First Ave., Waltham 54, Mass., announces a significant advancement in the development of cesium thermionic converters. For the first time a cesium thermionic converter has been life-tested in excess of 900 hours. This duration is nearly 600 hours longer than any previously announced. Even more significant is the fact that the converter is a hardware type converter which was developed in connection with the National Aeronautics and Space Administration's Jet Propulsion Laboratory's Set Solar Thermionic Power System Program. Thermo Electron is subcontractor to Electro-Optical Systems, Inc., of Pasadena, Calif., for the thermionic generator used in the Set Systems. The tiny converter weighs less than one eighth ounce.

(Continued on page 128A)

MICROWAVE COMMUNICATIONS ENGINEERS

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ELECTROMAGNETIC ENGINEERS

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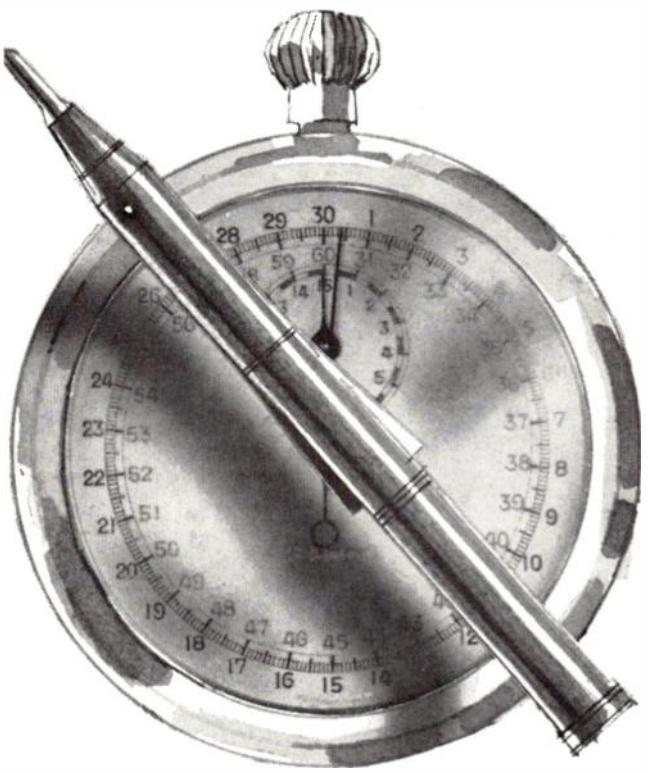
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NEWS New Products

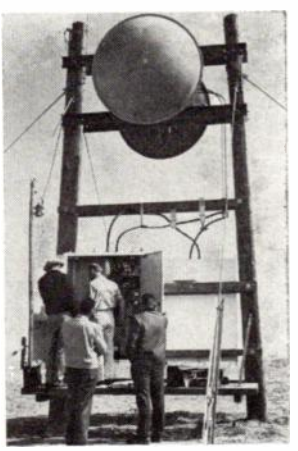
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 126A)

Microwave System

The new TL Microwave system, developed by **Bell Telephone Laboratories**, and manufactured by **Western Electric Co.**, Merrimac Valley Works, North Andover, Mass., is one of the first transistorized microwave systems to go into production in this country. Because of the low power requirements of the solid-state circuits, and efficient new klystron power tubes, the system requires only 150 watts of power per terminal and can operate on storage batteries should a power failure occur. The use of batteries for emergency power and the compactness of the solid-state circuitry permit a complete repeater station to be mounted directly on the poles or tower supporting the antenna. Previous systems have required separate buildings to house electronic equipment and engine generators for standby power.

Designed for relatively light traffic and distances up to 200 miles, the TL microwave system provides up to six two-way broadband channels, each of which can carry 240 telephone conversations. Its primary uses will be to replace or augment existing open wires, provide spurs on heavier-traffic microwave routes, and provide short-term or emergency restoration service. In most parts of the United States, stations will average about 15 miles apart.



This relay station near Billings, Montana is part of the first installation of the new TL microwave system.

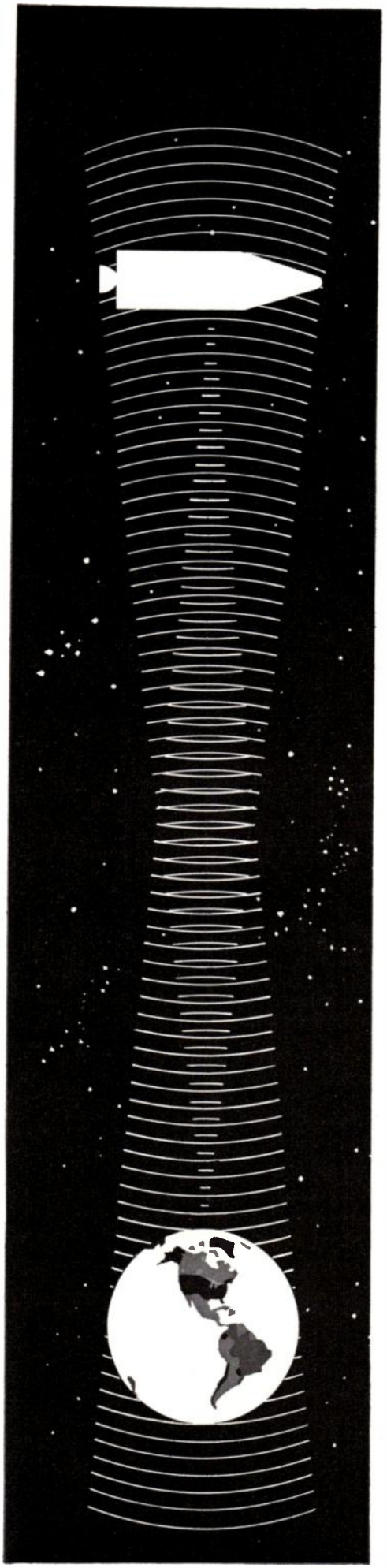
The new system operates in the 11,000 mc frequency range —far above the 4000 and 6000 mc ranges which carry the major share of telephone and television traffic. The higher frequency was chosen to prevent interference with present microwave channels, which are becoming increasingly congested.

At the heart of the TL system are two newly-developed klystron tubes, which serve as the high frequency sources for the transmitter and the receiver oscillator. These tubes are designed for an operating life of 40,000 hours, and are the only electron tubes in the radio system. All other active circuit elements are solid-state devices. The klystrons are maintained at an almost constant temperature over a wide range of conditions by an ingenious vapor cooling system. A small boiler, partially filled with a fluorochemical that boils at 214°F., is in close contact with the klystrons. When their temperature exceeds 214°F., the chemical begins to boil. The vapor flows out into a sealed cooling system where it condenses and flows back into the boiler. The vapor cooling system is so effective over an outdoor temperature range of −40 to 120°F. that no automatic frequency control is necessary in the transmitter.

A total of 27 transistors and 23 wideband ferrite transformers are used in the receiver part of the TL system to provide an overall IF gain of 110 db. Automatic gain control is provided by a circuit employing special diodes which attenuate the input signal by an amount depending on an applied dc current. The dc current is controlled by the strength of the signal delivered by the IF amplifier. Two of these circuits, each containing three diodes, are interspersed between three sections of the IF amplifier.

Many components of the system were designed as plug-in units for simplified maintenance. The system can be maintained with one portable transistorized test set.

(Continued on page 130A)



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Unusual analysis and design problems in a variety of Missile and Space programs at Lockheed Missiles & Space Company are challenging the state-of-the-art of equipment and system design.

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- High density magnetic and solid state circuit packaging for severe environmental conditions.
- Development of vehicle-borne signal handling equipment including multiplexing, digitizing, and processing devices.
- Development of new and flexible ground-based detection, demultiplexing and decoding devices.
- Analysis of capabilities and limitations of new techniques in communication systems for satellites and deep space probes.
- Development of new coding and modulation techniques for improved communication efficiency and link security.

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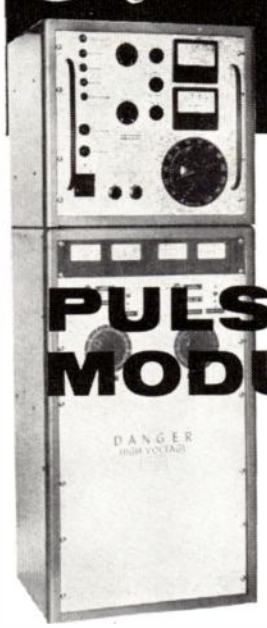
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NEWS New Products

(Continued from page 128A)



Assembly of equipment is carried out at Western Electric's Merrimack Valley Works. Here Maurice Lynch inspects TL equipment mounted in relay rack while (left to right, foreground) tester Paul Marquis and inspection supervisor Bud Collins check wiring diagram.

At each TL station, the incoming signal is demodulated to baseband (the range of "carrier" frequencies used to carry telephone conversations over wires). This provides access to all the telephone channels at each station. The baseband signal is then amplified, remodulated onto the microwave signal and beamed to the next station.

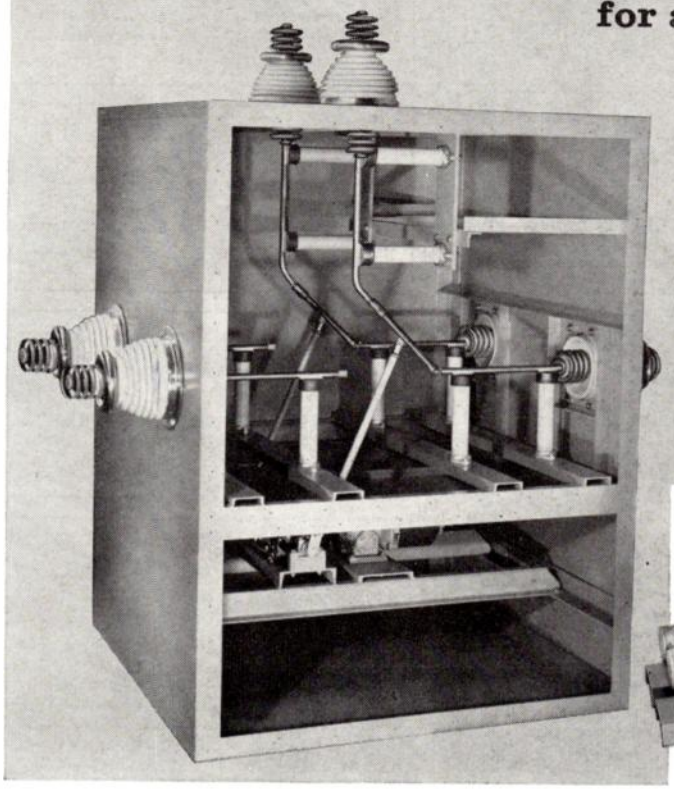
Normally one receiver and transmitter are provided at each repeater station for each direction of transmission but where the ultimate in reliability is required, duplicate or diversity facilities can be installed. In these cases, a diversity switch is provided which automatically chooses the receiver having the stronger signal at any instant. The system is monitored by a continuous 2600 cps tone, which is received at a control terminal so long as all the units are functioning normally. Any interruption of this tone automatically sounds an alarm. An attendant then can interrogate each station with a tone unique to that station to find out where the failure is.

A number of different antenna and tower arrangements can be used with the TL system. Although a five- or ten-foot para-

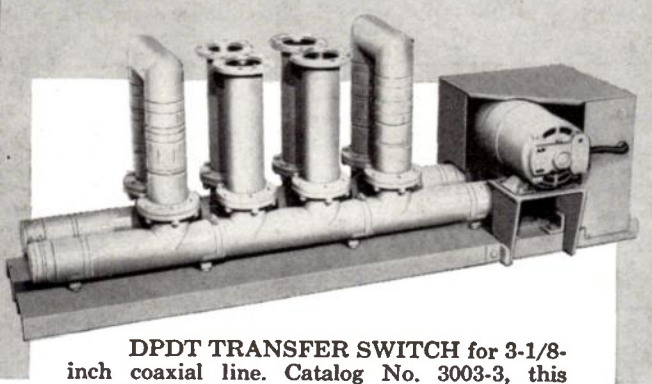
(Continued on page 132A)

High Power RF Switches...

for application to open-wire and coaxial transmission systems



Typical Open-Wire DPDT Switch. Designed and manufactured for Page Communications Engineers, Inc., a number of these switches are installed at the RADC Ava Transmitter Test Site. Impedance is 425 ohms. Under 100% sine-wave AM conditions, switch is rated at 600kw peak (225kw average, 150kw carrier) after allowing for 2:1 VSWR and a safety factor of 2 on peak voltage.



DPDT TRANSFER SWITCH for 3-1/8-inch coaxial line. Catalog No. 3003-3, this switch may be used to 700 mc/s. Power rating is that of transmission line under the same operating conditions.

To control and direct high-level r-f power, *Dielectric* offers both coaxial and open-wire switches with VSWR's typically below 1.05 at normal operating frequencies and less than 1.1 at the highest frequencies for which the transmission line is usable.

In coaxial switches, both cold- and hot-break models are available, pressurized and non-pressurized, in line sizes from 1-5/8 to 9-3/16-inch and larger diameters. Power ratings are equal to line ratings for both cold-break and hot-break models. Upper frequency limits vary from 1400 mc/s for 1-5/8-inch line to 240 mc/s for 9-3/16-inch line.

Cold-break, weatherproof, open-wire switches for 1-30 mc/s, in DPDT or matrix designs, can handle up to 500 kw carrier power (2 megawatts peak, 750 average) under 100% sine-wave AM conditions and after allowing for 2:1 load VSWR. No stubs are employed, isolation between circuits is typically >70 db, and impedance may range from 75 to 600 ohms. Switching time is 2 to 5 seconds depending on size.

All *Dielectric* switches, coaxial or open-wire, may employ any drive: electric motor, pneumatic or hydraulic cylinder. Remote control and interlocking are standard. Coaxial switches are presently available in SPDT, DPDT, and transfer contact forms, and may be stacked for more complex switching. Soon to be announced is a "building-block" model for matrixing. Open-wire switches are available in DPDT and matrix designs.

For further information, request Coaxial-Switch Bulletin No. 61-9 and/or Open-Wire Switch Bulletin No. 61-10.

To meet future needs, when mechanical switches will be inadequate, *Dielectric* has investigated the feasibility of high-power, hot-break switching with operating time measured in nanoseconds. Work to date shows that ultra-fast switching can be accomplished and that practical hardware can be constructed.

Dielectric does not offer these switches as catalog items but is interested in discussing potential applications with interested groups.

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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 130A)

bolic antenna can be mounted as a direct radiator when a low tower is used, the antenna is more commonly mounted at the base of the tower. The antenna then radiates directly upward, and the signal is reflected horizontally toward the next station by a flat metallic reflector at the top of the tower. This "periscope" arrangement produces less loss than transmitting the signal through long waveguides to a direct-radiating antenna at the top of a tall tower. The reflector size ranges from 6×8 ft. to 10×15 ft., depending on tower height. One antenna is used for both sending and receiving, the incoming and outgoing signals being separated by their difference in frequency and polarization.

Design of the TL system was carried out at the Merrimack Valley location of Bell Laboratories in North Andover, Mass. A research system which demonstrates the principles of the TL system was designed at Bell Laboratories, Holmdel, N. J., location and has been operating for 42 months. The rapid development of the present TL system, in 19 months, is the result of close cooperation among research, development and production engineers.

Semitronics Appoints Bird and McMahon

Semitronics, Incorporated, Winchester, Mass., announces the appointment of Gordon S. Bird, Jr., and David A. McMahon as Senior Staff Scientists.

Bird received the degree of Bachelor of Science in Geology from Bates College in 1958. Boston University awarded him the degree of Master of Arts in Mineralogy in 1960. Bird is presently a candidate for the Ph.D. degree and has completed all requirements except for the submission of his dissertation. His thesis research on Mineral Paragenesis has included X-ray crystallography and geometrical crystallography as well as studies in natural crystal and analysis.

During his graduate studies, Bird was awarded a graduate teaching assistantship and a teaching fellowship. In the period 1959–1961 Bird was employed by the U. S. Geological Survey as a geologist. His duties included mineral and structure an-

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alysis. In 1961 Bird joined Semitronics, Incorporated. His responsibilities include structure analysis, defect determination, test techniques and electrical properties measurement.

McMahon was awarded a Bachelor of Arts degree at Boston University in 1960. His areas of study were mineral chemistry, crystallography, and physics. McMahon has undertaken a program of graduate studies in the Boston University Graduate School and is expected to complete requirements in 1962.

During the period of 1959 through 1960, McMahon was employed at Transitron, first as production machine operator and then as production supervisor. He was transferred to the Research Department of Trancoa to assist in the development of improved high vacuum silicon single crystal techniques. He also participated in research in methods of purification of manganese, titanium, titanium tetraiodide, and lead telluride. In 1960 McMahon joined Solid State Materials Corporation as a project engineer. In this position he was assigned to single crystal growth studies using Czockralski, Bridgmann, hydrothermal and zoning techniques. The materials whose crystallization he studies included lead telluride, the III–V, II–VI compounds and ferromagnetic materials.

In 1961 McMahon joined Semitronics as Staff Scientist. His duties included crystal growing, junction development methods, diffusion rate studies, and surface defect density studies.

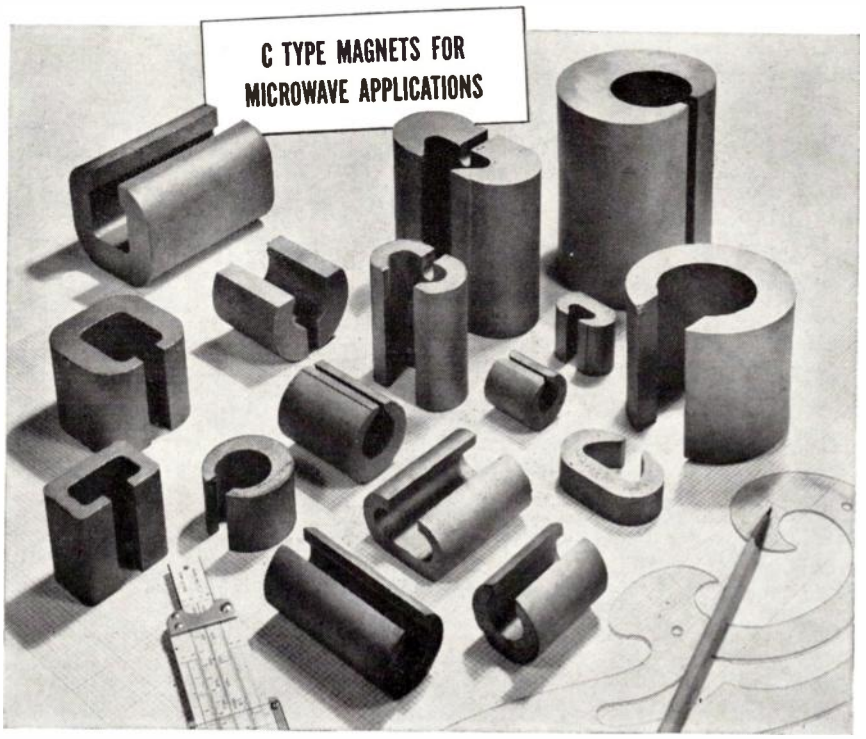
Research-Cottrell Announces New Division

Research-Cottrell, Inc., Bound Brook, N. J., has formed an **Electronics Div.** to market customs designed high-voltage equipment to the electronics industry. The division will offer power supplies, transformers, control systems, and other special electronic equipment, initially in the range of 5,000 to 150,000 volts with power outputs to 500 kw, for specialized applications. The new division was formed to manufacture high-voltage equipment for manufacturers and suppliers of microwave systems, radar, and nuclear power equipment, and by research-development programs requiring custom design high-voltage components and power systems.

Activities of the division have already commenced and orders are in progress for transformers for the Plasma Physics Laboratory at Princeton University's James Forrestal Research Center, and power supply equipment for the Plasma Propulsion Laboratory of Republic Aviation Corp. for use in research and development.

Research-Cottrell is credited with developing and introducing many electrical equipment innovations in the field of electrostatics for industrial gas cleaning. They were the first to adapt automatic rectifier control systems, silicon rectifiers, and continuous electrode rapping systems for electrostatic processes to attain higher cleaning efficiencies.

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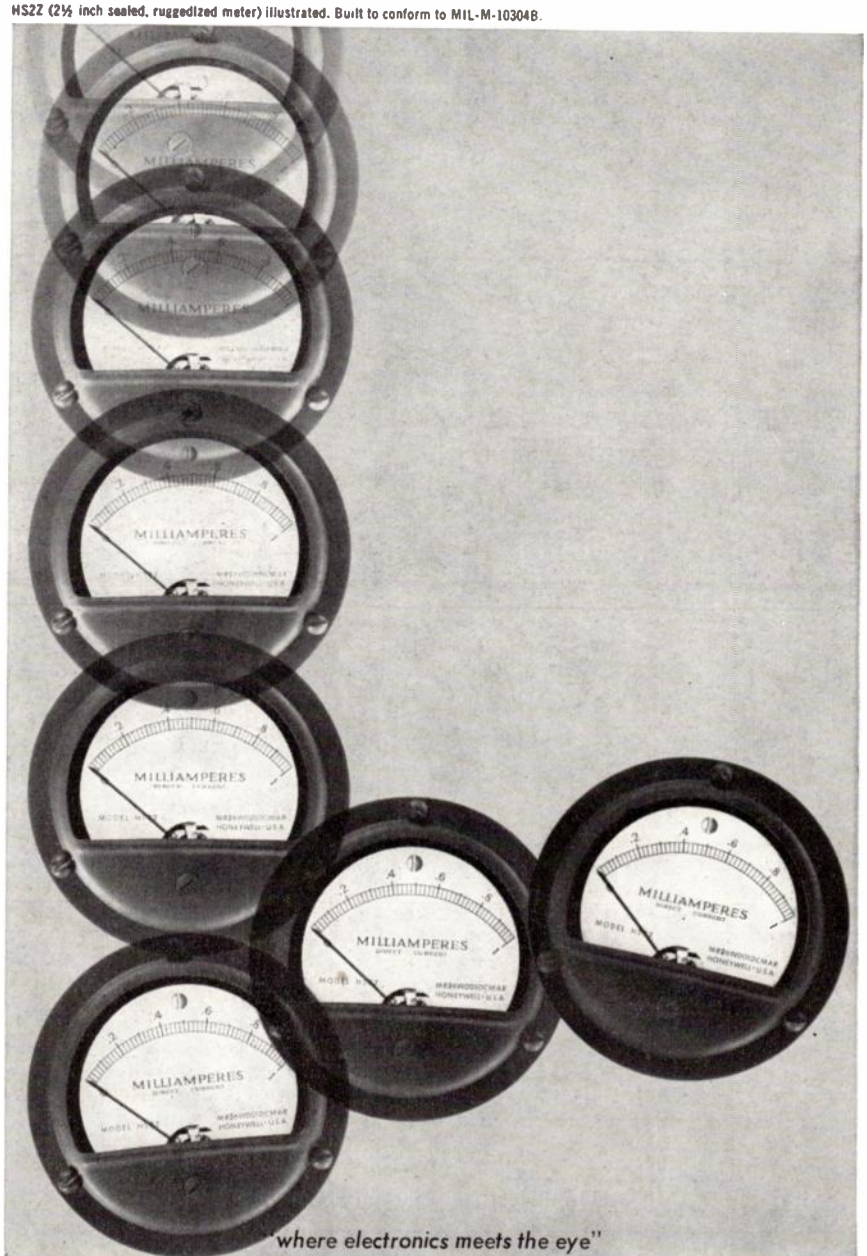
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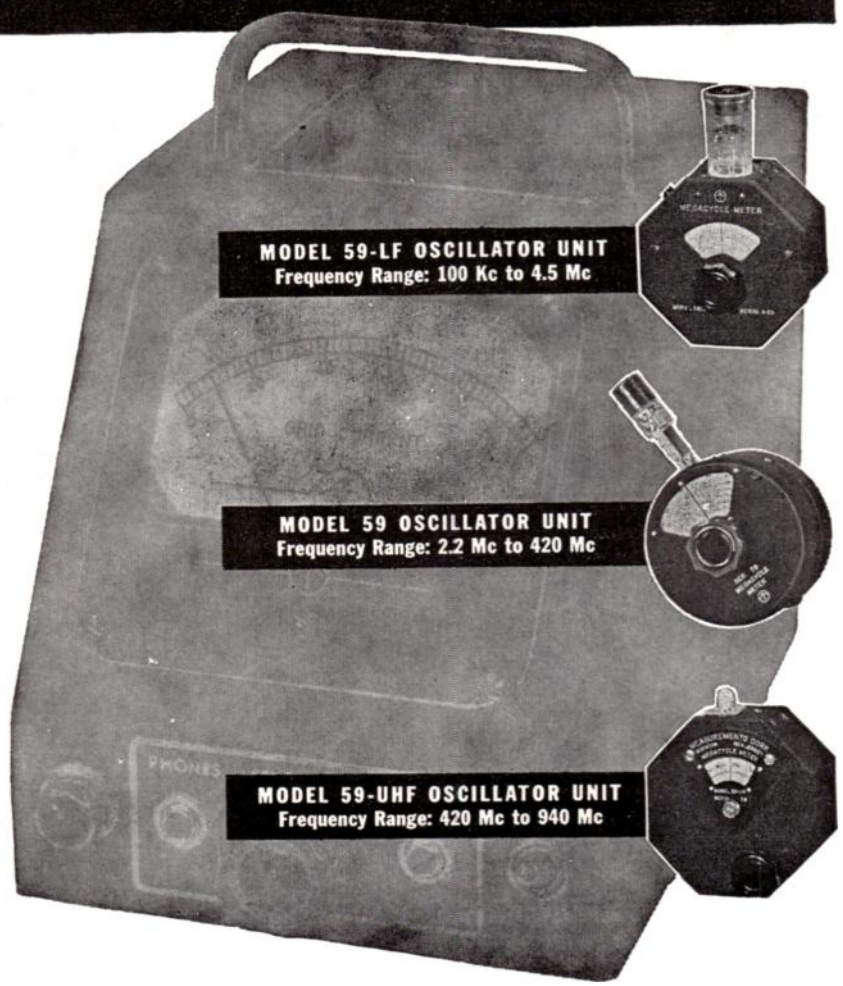


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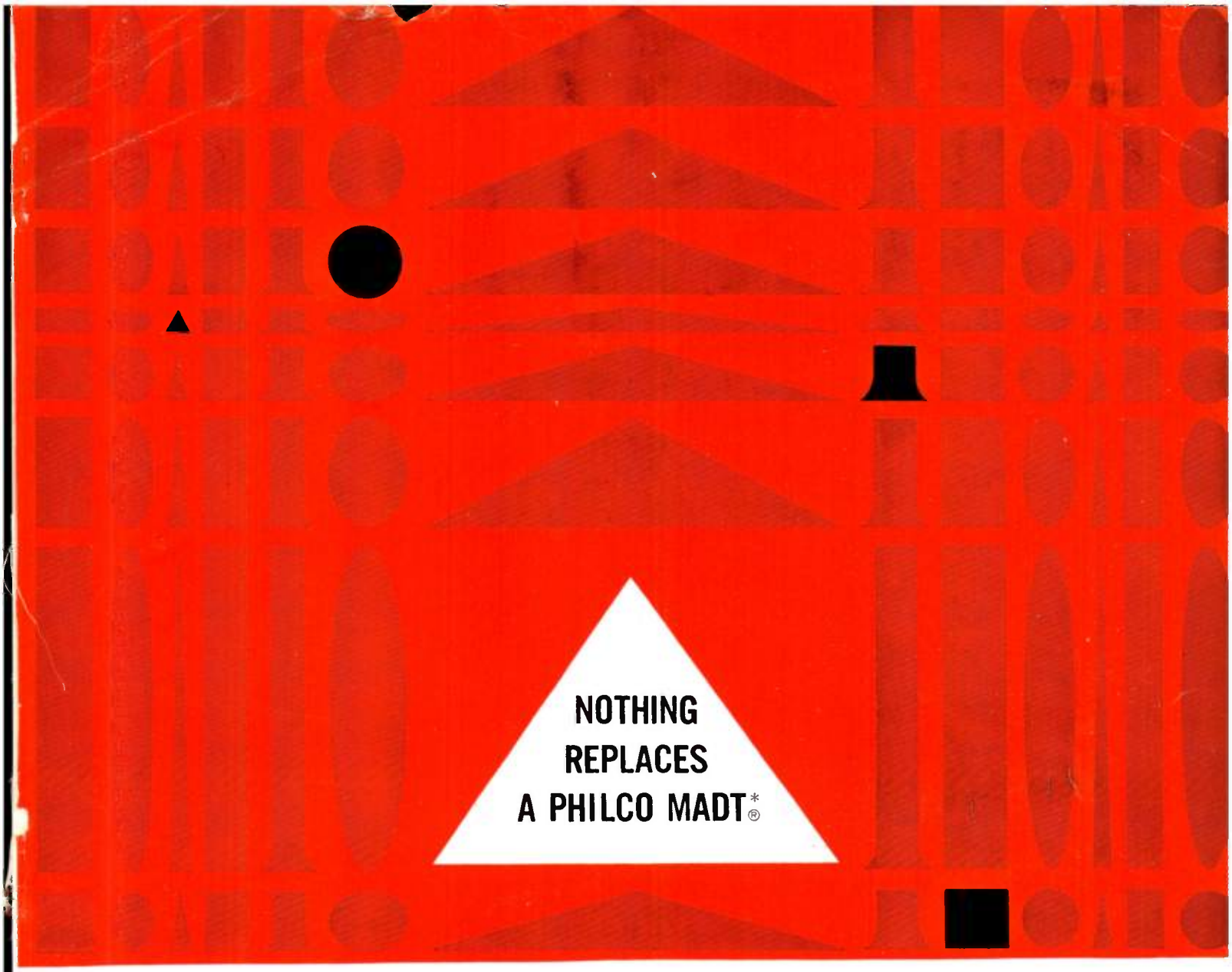
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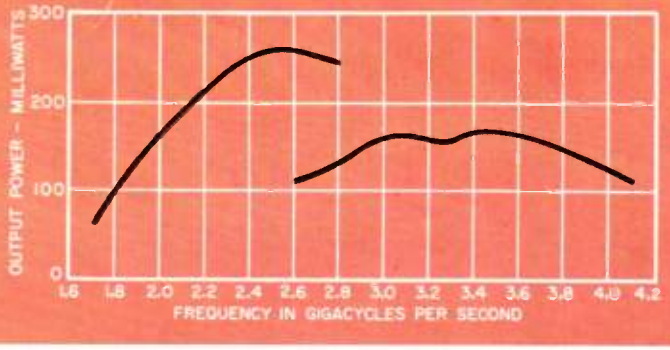
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