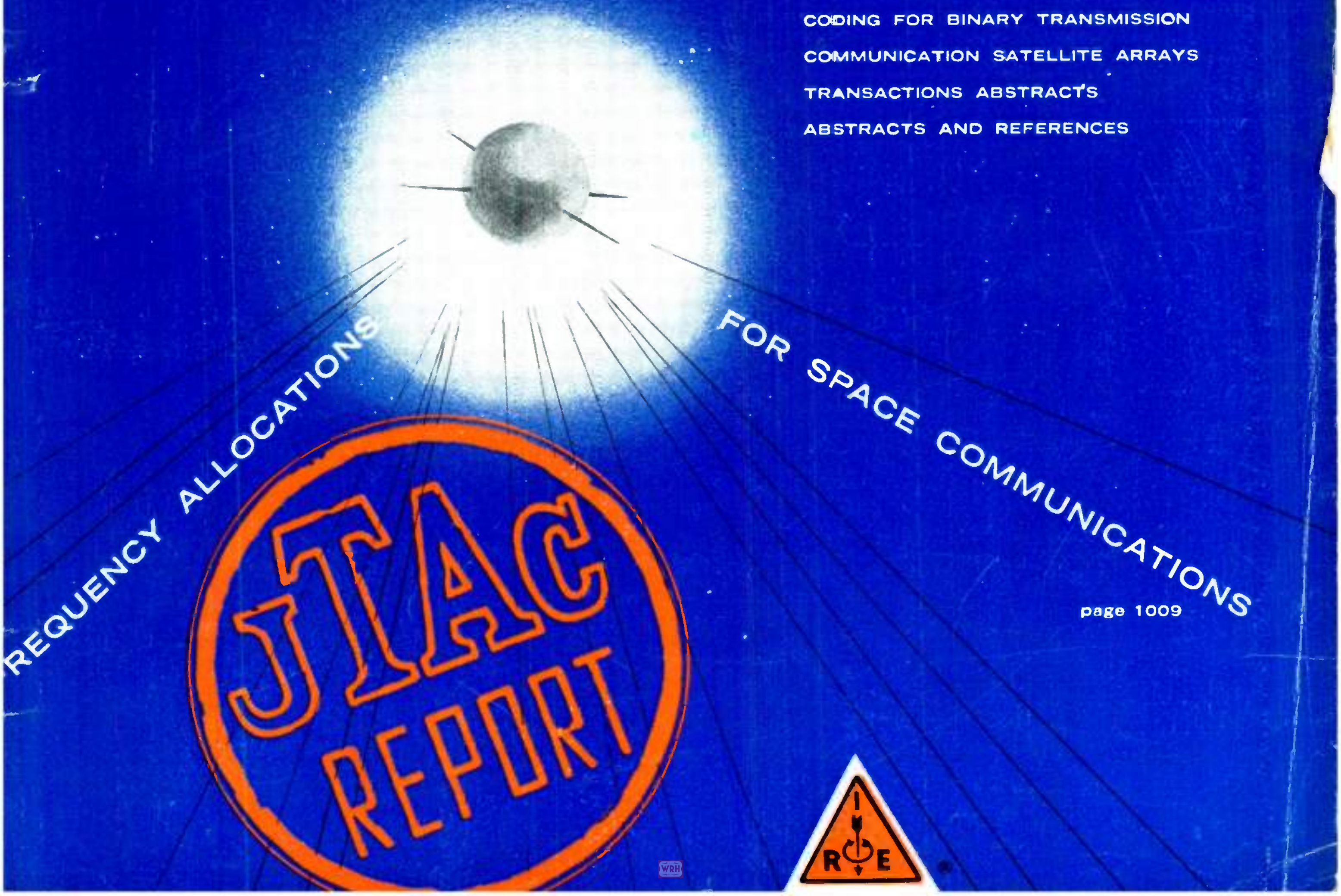
june 1961

the institute of radio engineers

Proceedings of the IRE

in this issue



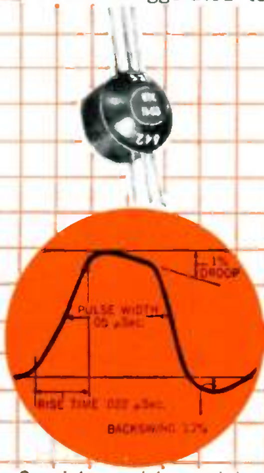
page 1009

SPECIAL PULSE TRANSFORMERS *TO YOUR REQUIREMENTS*

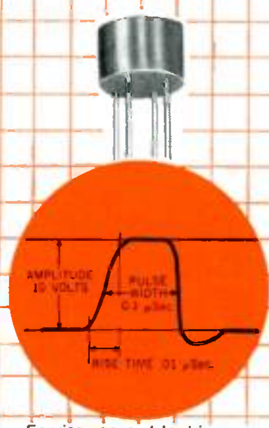
The pulse units illustrated below show a few of the thousands of special types designed and produced by UTC, to customers' requirements. Range covered on special pulse units is from a few microwatts to 10 megawatts. Rectangular pulse shapes are deliberately shown exaggerated to clarify parameters.

Almost thirty years of pioneering in the design and production of transformers plus exhaustive life testing programs and rigid quality control measures guarantee components of the highest reliability in the industry. . . . You can stake YOUR reputation on UTC products.

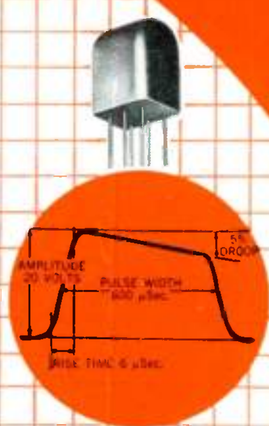
25 KC pulse transformer, DO-T or DI-T configuration. Pulse width 60 μSec. Rise time less than 0.5 μSec. Secondary C.T. balance each side to within 1% to ground. MIL-T-27A, GR 4. Size: DO-T, 5/16 dia. x 13/32", wt. 1/10 oz.; DI-T, 5/16 dia. x 1/4", wt. 1/20 oz.



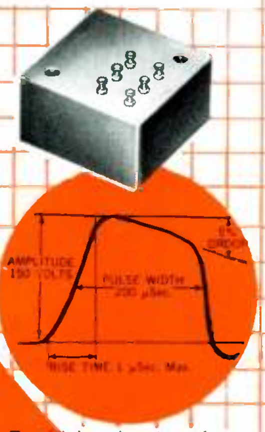
Special precision miniature pulse transformers. Designed in our standard stock mold to your specs. Checked and precisely adjusted in your tube or transistor blocking oscillator circuit. Sizes: 3/8 dia. x 3/8", 1 gram; 9/16 dia. x 9/16", 4 grams; 5/8 dia. x 5/8", 6 grams.



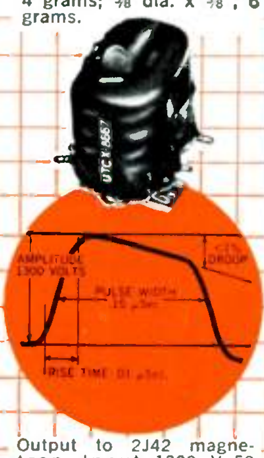
Ferrite core blocking oscillator transformer. 0.1 μSec. ±10% @ 200 KC PPS. 2 windings, rise time .01 μSec. Epoxy case. MIL-T-27A; 7/16 dia. x 1/4", .07 oz.



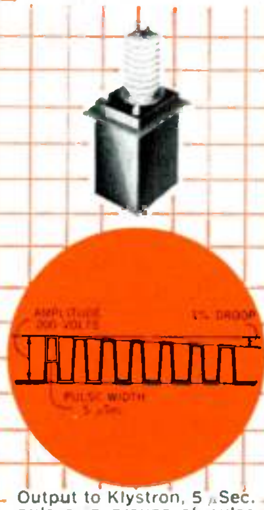
600 μSec. coupling transformer for printed circuit application. 3 windings, mu metal case for extreme shielding. Z=20 K Ω, 10 V. MIL-T-27A; standard UTC ML case; 7/16 x 31/64 x 5/16" h., .2 oz.



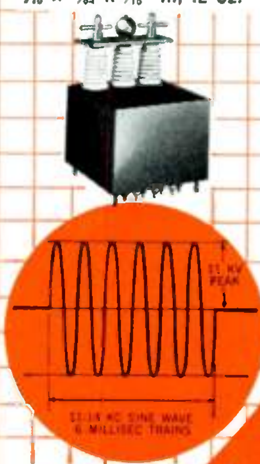
Toroidal pulse transformer, 150 V. 200 μSec. @ 400 PPS. Molded in epoxy. 3 windings, low leakage, less than 1 μSec. rise time. —40°C. to +85°C., output voltage within 1%. MIL-T-27A; 1⅜ sq. x ¾", 1.5 oz.



Output to 2J42 magnetron. Input 1300 V.-50 ohms. Output 6.5 KV to 1200 ohms and .6A. bifilar filament winding. .15 μSec., 1000 PPS. Trigger winding. MIL-T-27A GR 5; 1½ x 2½ x 2¾", 10 oz.



Output to Klystron, 5 μSec. pulses in groups of pulse trains at high rep rate. Droop 1% over pulse trains. 30 KV W.V., 43 KV hipot; —53°C. to +85°C.; MIL-T-27A; 4¼ x 5 x 6¾"; 11½ lbs.



Sonar sine wave pulse output transformer. PP 4-65A's, 11-14 KC flat. Pri. 11 KV; 28 KV hipot. Spark gap protected. Sec. 1500 V. @ 800Ω, 60 millisec on, 6% duty cycle. MIL-T-27A; —55°C. to +85°C.; 6 x 6 x 8½", 13 lbs.

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VOLUME 49, NUMBER 6

June, 1961

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

contents

COVER

The Joint Technical Advisory Committee has undertaken a special study of the technical factors which are involved in the allocation of frequencies for communication satellites. The preliminary findings of the JTAC appear on page 1009.

JUNE

JUNE

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

continued

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TYPE 2N768

- Micro-energy switch—designed for low current, low voltage, high speed applications
- 10 mc pulse rates, collector currents as low as 1 ma, collector supply voltages as low as 1 volt
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TYPE 2N769

- World's fastest switch—will operate reliably at speeds in excess of 100 mc
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- Manufactured with tighter parameter control than any other transistor in the industry
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2N501A	Ultra High Speed Switch (Storage Temperature, 100 C)
2N504	High Gain IF Amplifier
2N588	Oscillator, Amplifier, to 50 mc

For complete engineering information on the types in which you are interested, write Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

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MAGNETIC COMPONENTS	PULSE-FORMING NETWORKS	FUNCTIONAL DIGITAL CIRCUITS

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AIL is engaged in a cooperative program for the NASA with the Central Radio Propagation Laboratory designed to orbit an ionosounder well above the F_2 region of the ionosphere. We call this the Topside Sounder program. In this first article of a short series, Warren Offutt describes some of the history leading up to this fascinating space program of ionospheric research.

Sounding the Ionosphere — From Topside

PART I

On 12 December 1901, an experiment was concluded with results that startled the scientific community. Marconi successfully received wireless signals at St. John's in Newfoundland that had been transmitted at Poldhu, Cornwall. Although classical diffraction theory clearly predicted that the attenuation over the earth's curvature doomed the experiment to failure, the signals were satisfactorily received. A plausible explanation was not long in coming: in 1902 Heaviside in England and Kennelly in America independently postulated the existence of a conducting layer in the earth's upper atmosphere.

Proof of the presence of the Kennelly-Heaviside layer through measurement of its altitude was obtained in 1924. In that year Appleton and Barnett of the Cavendish Laboratory pointed out that the radially symmetric variation in field strength they observed as a function of distance from a powerful radio transmitter could best be interpreted as interference fringes resulting from a conducting sheet some 60 miles overhead. Not long thereafter, Breit and Tuve in America strengthened the proof by operating a "radar" (ionosounder) in the HF band. During the next 15 or 20 years a few dozen ionospheric observatories around the world were equipped wtih ionosounders, and a program of systematic observation and study of the bottom side of the ionosphere emerged (Figure 1). A sufficiently good understanding of the behavior of the free electron density of the bottom side has been acquired to permit excellent predictions of its diurnal, seasonal, and solar cycle variation.

These studies of the ionosphere have extended from ground level up to about 300 km. Unfortunately, when the relatively simple sounding techniques are used from the ground, they cannot provide data for altitudes greater than that at which the peak density occurs because of shielding by the high-density region.

The free electron density (Ne) at greater altitudes than F_2 max is very

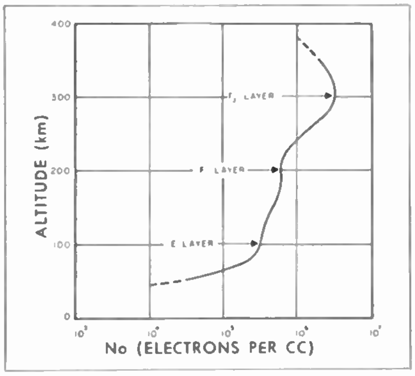

FIGURE 1. TYPICAL UPPER ATMOSPHERE IONIZATION AT MIDDAY DURING SUNSPOT MAXIMUM.

much of interest, but it is only within the recent past that any data have been acquired. It is interesting to note that many of the theoretical physicists during the 1930's predicted very low values for Ne in interplanetary space, and indeed even for altitudes of only a few thousand kilometers. During the early 1950's, L. R. O. Storey inferred from theoretical studies and measurements of the whistler mode of signal propagation values of Ne on the order of a few hundred electrons/cc at altitudes of 10,000 to 20,000 km. Although his conclusions were not in good agreement with those based upon the then accepted model atmosphere, measurements during the past few years of Ne versus altitude now tend to support Storey's conclusions.

We have no detailed data on the topside at all. Indeed, a few rocket probes and density measurements have resulted in a change in belief of fairly major proportions. We don't know whether or not the topside is smooth either in Ne versus altitude or geographically, or both. We don't know whether "spread-F" conditions exist above as well as below the maximum density region. We have scarcely any information on how the topside variations correlate with time, season, or solar disturbances. One might say we barely know that the topside of the ionosphere exists at all.

The desire to acquire knowledge about the topside stems from two types of interests. First, there is the immediately practical use to which such knowledge could be put—for example, to enhance the predictability of HF communication circuit quality through better understanding of the ionosphere. Second, the topside of the ionosphere is a basic part of the near-space environment, about which we must learn much more before man can build a home in the sky.

The techniques that can be applied to acquire some of this knowledge include: (1) rocket probes for occasional spot sampling, (2) high-power VHF vertical-incidence soundings of electron-scattered signals for continuous sampling at one location, (3) an orbiting ionosounder, and (4) further studies of whistler mode propagation using man-made signals under controlled conditions. AIL is at present actively engaged in (1), (3), and (4). Next month on this page we will tell you of our work with NASA and others in this program. We think you will find the photographs and diagrams of the rocket and satellite equipment of general interest, and the photographs of A-scope ionosphere echoes of special interest.

A complete bound set of our fifth series of articles is available on request. Write to Harold Hechtman at AIL, Comac Road, Deer Park, L.I., New York for your set.

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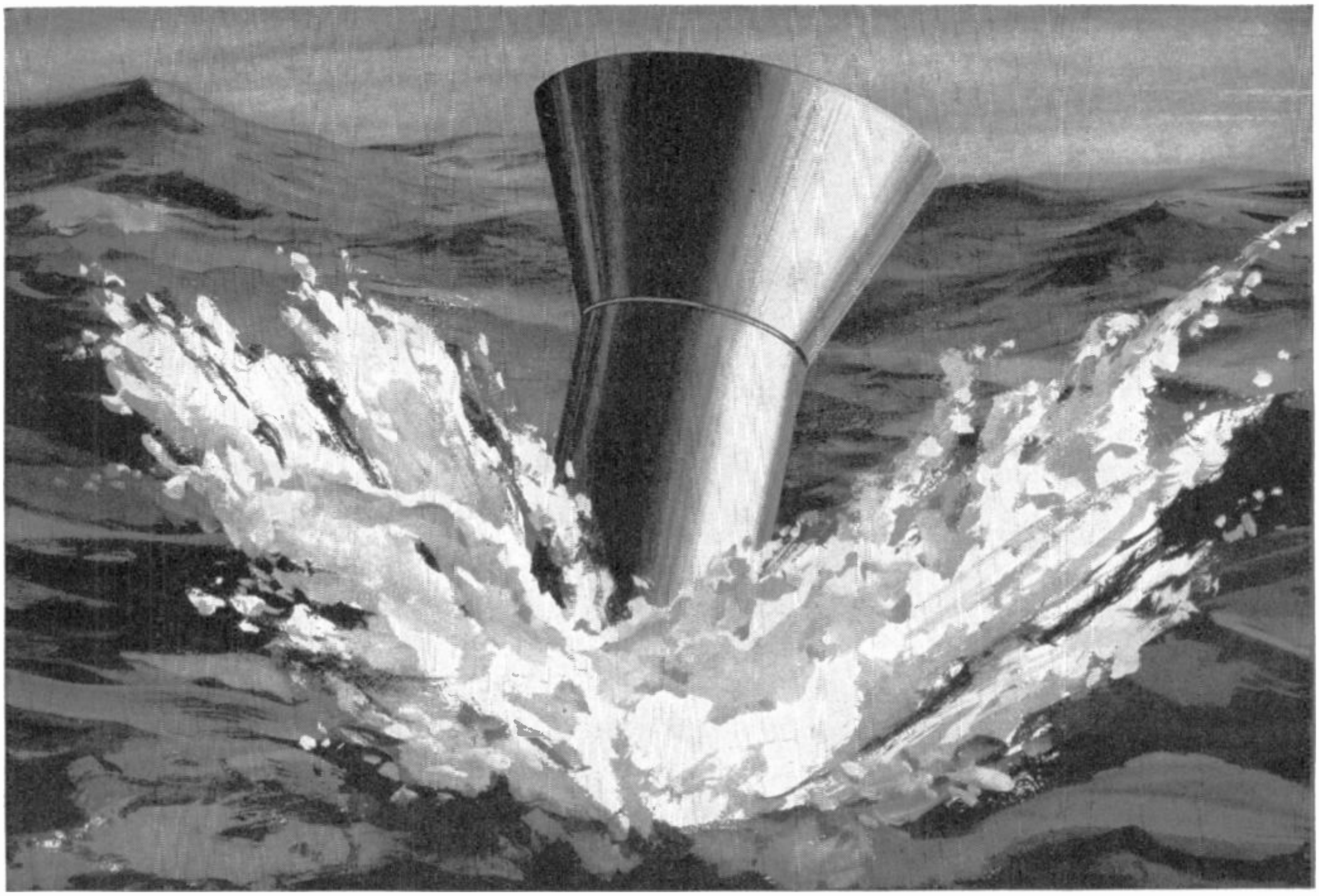

HOW THE OCEAN GREW "EARS" TO PINPOINT MISSILE SHOTS

A quarter of the world away from its launching pad an experimental missile nose cone enters its ocean target area.

How close has it come to the desired impact point?

Where actually did the nose cone fall?

To answer these questions quickly and accurately, Bell Laboratories developed a special system of deep-sea hydrophones—the Missile Impact Locating System (MILS) manufactured by Western Electric and installed by the U. S. Navy with technical assistance from Western Electric in both the Atlantic and Pacific Missile ranges. MILS involves two types of networks.

- One is a long-distance network which utilizes the ocean's deep sound channel. It monitors millions of square miles of ocean. The impacting nose cone releases a small bomb which sinks and explodes at an optimum depth for the transmission of underwater sounds. Vibrations from the explosion are picked up by hydrophones stationed at the optimum depth and carried by cables to shore stations. Time differences in arrivals between these vibrations at different hydrophones are measured and used to compute location of the impact.
- The other is a "bull's-eye" network that monitors a restricted target area with extraordinary precision. This network is so sensitive it does not require the energetic explosion of a bomb but can detect the mere splash of a nose cone striking the ocean's surface—and precisely fix its location.

The universe of sound—above the earth, below the ocean—is one of the worlds of science constantly being explored by Bell Laboratories. The Missile Impact Locating System reflects the same kind of informed ingenuity which constantly reveals new ways to improve the range of Bell System services.

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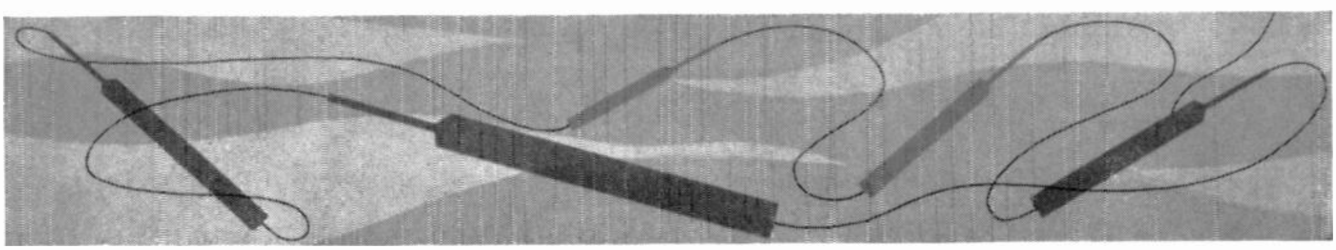

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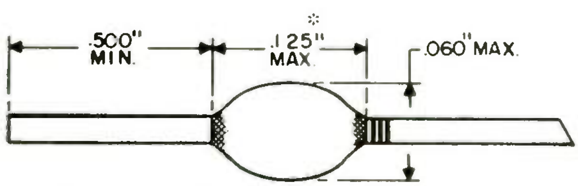

*Available also in .090" with slightly less power dissipation.

ALL micro glass DIODES MEET ENVIRONMENTAL REQUIREMENTS OF MIL-S 19500B

TYPE	FORWARD CURRENT IF @ 1 v DC (mA)	BREAKDOWN VOLTAGE @100 µA (volts)	CAPACITANCE (max.) (µµf)	REVERSE CURRENT (µA) 25°C	REVERSE CURRENT (µA) 150°C	REVERSE RECOVERY max. (nanosec.)	POWER DISSIPATION (mW)
MA-4303	10.0	50.0	1.5 @ — 6v	.025 @ — 40v	25.0 @ — 40v	4.0	125
MA-4304	10.0	50.0	1.5 @ — 6v	.025 @ — 40v	25.0 @ — 40v	4.0	200
MA-4305	10.0	50.0	1.5 @ — 6v	.025 @ — 20v	25.0 @ — 20v	4.0	125
MA-4306	10.0	50.0	1.5 @ — 6v	.025 @ — 20v	25.0 @ — 20v	4.0	200
MA-4307	30.0	100.0	2.0 @ 0v	.050 @ — 75v	50.0 @ — 75v	4.0	125
MA-4308	30.0	100.0	2.0 @ 0v	.050 @ — 75v	50.0 @ — 75v	4.0	200

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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.

June 6-8, 1961

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

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

June 13-14, 1961

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

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

June 19-20, 1961

Second National Conference on Broadcast and Television Receivers, O'Hare's Inn, Des Plaines, Ill.

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

June 26-28, 1961

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

Exhibits: Mr. L. David Whitelock, 6514 Greentree Road, Bethesda 14, Md.

July 16-21, 1961

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

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

August 22-25, 1961

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

Exhibits: Mr. Don Larson, WESCON, 701 Welch Road, Palo Alto, Calif.

September 6-8, 1961

National Symposium on Space Electronics & Telemetry, Albuquerque, N.M.

Exhibits: Mr. V. V. Myers, 2912 Texas N.E., Albuquerque, N.M.

October 2-4, 1961

Seventh 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 2-4, 1961

IRE Canadian Convention, Automotive Building, Exhibition Park, Toronto, Canada.

Exhibits: Business Manager, IRE Canadian Convention, 819 Yonge St., Toronto 7, Ontario, Canada.

(*Continued on page 10A*)

Meetings with Exhibits

(Continued from page 8A)

October 9-11, 1961

National Electronics Conference, Hotel Sherman, Chicago, Ill.

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

October 19-20, 1961

Sixth Annual North Carolina Section Symposium on Electronics, Engineering and Education, Greensboro Coliseum, Greensboro, N.C.

Exhibits: Mr. H. G. Eidson, Jr., Dept. 8760, Charham Road Plant, Western Electric Co., Inc., Winston-Salem, N.C.

October 23-25, 1961

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

Exhibits: Mr. Robert J. Henderson, Martin Company, Ground Support Equipment Dept., Baltimore, Md.

October 23-26, 1961

Eighth Annual Meeting, Professional Group on Nuclear Science (Aero-Space Nuclear Propulsion), Hotel Riviera, Las Vegas, Nevada

Exhibits: Mr. D. J. Knowles, Union Carbide Company, Oak Ridge National Lab., Oak Ridge, Tenn.

October 26-27, 1961

Electronic Techniques in Medicine & Biology Conference, University of Nebraska, Omaha, Neb.

Exhibits: Mr. Harold G. Beenken, University of Nebraska, College of Medicine, 428 Dewey Avenue, Omaha, Neb.

November 14-16, 1961

Mid-America Electronics Conference (MAECON), Hotel Muehlebach, Kansas City, Mo.

Exhibits: Mr. Felix A. Spies, Bendix Corporation, P.O. Box 1159, Kansas City 41, Mo.

November 14-16, 1961

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

Exhibits: Miss Shirley Whitcher, IRE Boston Office, 313 Washington St., Newton, Mass.

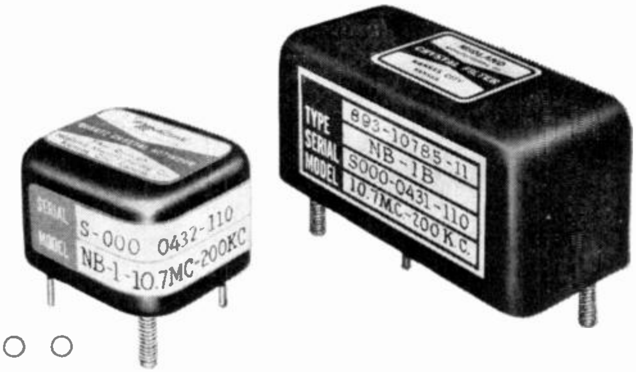
November 30-December 1, 1961

PGVC Annual Meeting, Hotel Raddison, Minneapolis, Minn.

Exhibits: Mr. Harold T'Kach, Mobile Engineering, Inc., 620 North Sixth St., Minneapolis, Minn.

Δ

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



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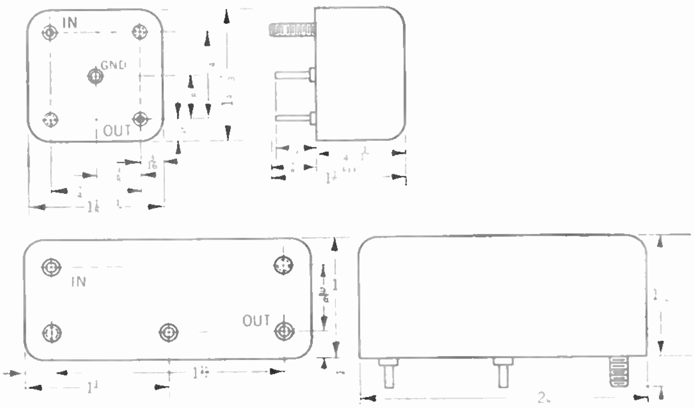

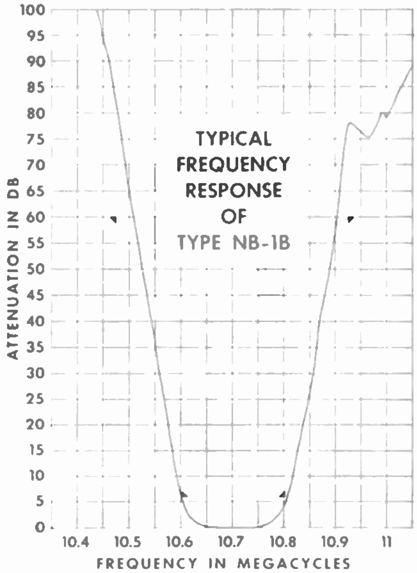

CRYSTAL FILTER SPECIFICATIONS

	TYPE NB-1	TYPE NB-1B
Center Frequency	10.7MC ± 3KC	10.7MC ± 3KC
Bandwidth @ 6db	200KC + 10KC, — 0 KC	200KC + 10KC, — 0 KC
Bandwidth @ 60db	450KC Max.	450KC Max.
Bandwidth Ratio	2.25:1 Max.	2.25:1 Max.
Ultimate Rejection	90db Min.	100db Min.
Insertion Loss	12db Max.	*8db Max.
Required Source/Load Resistance	50 ohms ± 5%	50 ohms ± 5%
Inband Ripple	1db Max.	*1.5db Max.
Inband Ripple at Temperature Extremes	1.5db Max.	*2.0db Max.
Operating Temperature	—55° C to +90° C	—55° C to +90° C
Shock	100 g	100 g
Vibration	15 g to 2KC	15 g to 2KC

*The Type NB-1B can also be provided with an insertion loss of 12db max.; inband ripple of .5db max.; and inband ripple at temperature extremes of 1db max. When ordering, specify required insertion loss and ripple.

MID 2-61

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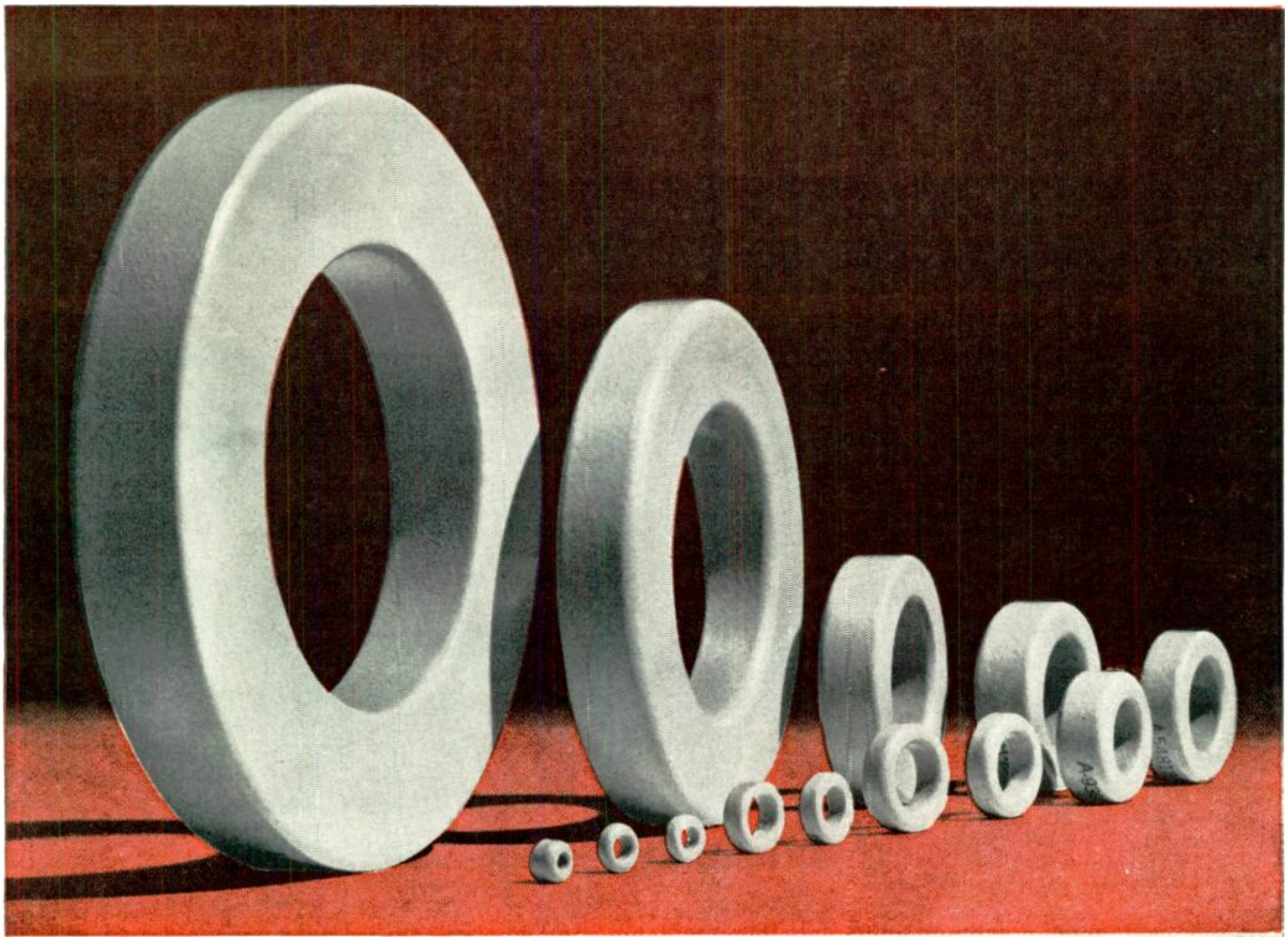

IRE News and Radio Notes

Current IRE Statistics

(As of April 30, 1961)

Membership—86,637
Sections*—110
Subsections*—30
Professional Groups*—28
Professional Group Chapters—286
Student Branches†—207

* See May, 1961, for a list.
† See this issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1961

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

5th Nat'l. Conf. on Product Engrg. and Production, Hotel Sheraton, Philadelphia, Pa., June 14–15.

Chicago Spring Conf. on Broadcast and Television Receivers, O'Hare Inn, Des Plaines, Ill., June 19–20.

MIL-E-CON 1961, Shoreham Hotel, Washington, D. C., June 26–28.

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

4th Int'l. Conf. On Medical Electronics & 14th Conf. on Elec. Techniques in Medicine & Biology, Waldorf-Astoria Hotel, New York, N. Y. July 16-21.

WESCON, Cow Palace, San Francisco, Calif., Aug. 22–25.

3rd Int'l. Conf. on Analog Computation, Belgrade, Sept. 4–9.

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

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

Int'l. Conf. on Electrical Engrg. Education, Syracuse Univ., Adirondacks, N. Y., Sept. 6–13.

IRE Conf. on Technical-Scientific Communications, Bellevue Stratford Hotel, Philadelphia, Pa., Sept. 13–15.

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

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

CISPR, Univ. of Pennsylvania, Philadelphia, Oct. 1–6.

7th Nat'l. Communications Symp., Utica, N. Y., Oct. 2–4. (DL*: June 1, R. K. Walker, 34 Bolton Rd., New Hartford, N. Y.)

IRE Canadian Electronics Conf., Automotive Bldg., Exhibition Park, Toronto, Canada, Oct. 2–4.

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Industrial Electronics Symposium Will Emphasize Supplier-User Cooperation

Industrial progress through cooperation and understanding between electronic equipment suppliers and industrial users of such equipment will be stressed at the 1961 Industrial Electronics Symposium. Sponsored jointly by the IRE (Professiona Group on Industrial Electronics), the AIEE, and the ISA, the Symposium will be held on September 20–21, 1961, in the Bradford Hotel, Boston, Mass.

The Symposium is honored to have C. Metcalfe, Chairman of the Electronic Forum for Industry, London, England, as a luncheon speaker. The Electronic Forum for Industry is a British organization formed in Feburary, 1959, to foster a much closer liaison between the manufacturers and users of electronic equipment.

A highlight of the meeting will be a session consisting of a panel of electronic experts in various phases of industrial electronics who will answer industrial users' questions. Moderator for the panel discussion will be William Vannah, former Chief Editor of *Control Engineering*, who is now an associate for advanced engineering of the Foxboro Company, Foxboro, Mass.

Three additional technical sessions are planned. The first one, on Measuring Techniques for Industry, will have Dr. C. W. Clapp, Consulting Physicist, Instrument Department, General Electric Company, as Session Chairman. Papers to be presented will cover topics such as noncontact measuring techniques, nondestructive fault testing, moisture measurement, and infrared and chromatography techniques.

The second session will deal with Digital and Analog Techniques in Industry. F. W. Atkinson of Owens-Corning Fiberglas will be Chairman. This session will cover both analog and digital blending techniques stressing actual case histories, three-mode controllers and computers.

The final session will cover New Power Conversion Techniques. C. H. Chandler of Gillette Safety Razor Company will be Session Chairman. Papers in this session will cover topics such as a novel system for coupling a turbine to a variable-speed drive with silicon-controlled rectifiers, ultrasonic machining techniques, industrial applications of thermoelectricity and electronic sources of industrial heat.

Air Force MARS Announces Schedule

The following is the June, 1961, schedule of broadcasts of the Air Force MARS Eastern Technical Net, operating Sundays from 2 to 4 P.M., EDST, on 3295, 7540, and 15,715 kc.

June 4—"Transistor Reliability," C. H. Zierdt, Product Manager, General Electric Company.

June 11—"Advancements in Broad Band Communications," A. Peterson, Communications Department, General Electric Company.

June 18—"Space Tracking," R. Anderson, General Electric Laboratory.

The Eastern Technical Net will recess until September 17, 1961, when it will begin its fifth year of weekly educational broadcasts. Listeners are invited to assist in programming the 1961–1962 season by suggesting the subjects in which they have particular interest. Every effort will be made to program for maximum technical interest. Suggestions and inquiries should be directed to: J. H. McCoy, Director, USAF MARS, Eastern Technical Net, Military Affiliate Radio System, Dept. of the Air Force, 109 Willow Ave., Huntington, N. Y.

The above photograph was taken in the Winston Churchill Auditorium, Technion City, Haifa, at the meeting of the Forum of the Fourth Israel; National Convention of Electronic Engineers and Annual Meeting of the IRE Israel Section. The Convention was held on February 20–21, 1961.

MIT Announces Special Summer Session

A summer program, "Engineering Magnetohydrodynamics," will be given at the Massachusetts Institute of Technology, Cambridge, on June 19–30, 1961, by faculty members in the departments of Aeronautics and Astronautics, Electrical Engineering, Mechanical Engineering, and Nuclear Engineering. The purpose of the program is to discuss the application of magnetohydrodynamic principles to engineering problems such as power generation, propulsion, flight control, fusion reactors, and liquid metal pumping. Lectures developing magnetohydrodynamics from both a continuum and particle point of view will be accompanied by treatments of applied topics such as magnetic confinement, instability, dc arcs, gas conductivity, MHD channel flows and boundary layers, traveling-wave accelerators, fusion devices, MHD shock tube experiments, superconducting magnets, etc. There will be separate panel discussions of power generation, propulsion, and fusion devices, in which invited guests will participate.

Inquiries concerning attendance at this special summer program should be directed to: The Office of the Summer Session, Massachusetts Institute of Technology, Cambridge 39, Mass.

Raymond F. Guy Receives NAB Award

Raymond F. Guy (A'25–M'39–F'39), American aid advisor now in Viet-Nam to assist in the establishment of a national radio network, has been awarded a distinguished service award by the National Association of Broadcasters.

He has received NAB's highest technical tribute, the 1961 Engineering Achievement Award for his service to his profession, the industry, and the nation.

The presentation was made at the NAB Convention on May 7–10, 1961, in Washington, D. C.

Past recipients of the NAB honors have included David Sarnoff, head of RCA, ex-President Herbert Hoover, and W. S. Paley of the Columbia Broadcasting System.

Mr. Guy is credited with the longest continuous service as a full-time broadcast engineer in the world.

Before going to Viet-Nam, he was a consultant in Haworth, N. J., and served RCA and NBC for nearly 40 years, entering boradcasting in New York in 1921 as engineering-announcer for WJZ. He has participated in most of the major international electronic conferences as well as dozens of important industry committees. He directed NBC's FM field tests in 1939–1940 and UHF TV tests at Bridgeport, Conn., in 1951–1953. He has been active in allocations of frequencies.

Other activities include both Treasurer and President of the IRE. He is Secretary and Past President of Broadcast Pioneers, and President of the Veterans Wireless Operators Association. He is a Fellow of the American Institute of Electrical Engineers and the Radio Club of America. He holds citations for service to broadcasting from the Radio and Television Executives Society and Broadcast Pioneers and holds the Marconi Gold Medal of the Wireless Operators Association.

Conference on Magnetism and Magnetic Materials Will Meet in Phoenix

The Seventh Annual Conference on Magnetism and Magnetic Materials will be held in Phoenix, Ariz., on November 13–16, 1961, at the Hotel Westward Ho. This Conference is sponsored jointly by the American Institute of Electrical Engineers and the American Institute of Physics, in cooperation with the Office of Naval Research, the Institute of Radio Engineers, and the Metallurgical Society of the AIME.

The deadline for abstracts is August 18, 1961. Further details can be obtained from the local chairman: P. B. Meyers, Motorola Semiconductor Products Division, 5005 E. McDowell Rd., Phoenix, Ariz.

The above photograph was taken at a meeting of the Executive Committee of the Seventh National Communications Symposium, which will be held in Utica, N. Y., on October 2–4, 1961. (Seated, *left* to *right*): R. K. Walker, Technical Program Chairman; Mrs. R. K. Walker, Women's Program Chairman; R. E. Bischoff, General Chairman; and R. E. Gaffney, Exhibits Chairman. (Standing, *left* to *right*): R. L. Libby, Future Scientists Chairman; R. L. Marks, Board of Governors; A. A. Kunze, Co-Executive Vice Chairman; W. P. Bethke, Scholarship Vice Chairman; B. H. Baldridge, Co-Executive Vice Chairman; C. J. Civin, Public Relations Chairman; C. W. Gordon, Chairman Board of Governors; and Lt. Col. T. G. Williams, Classified Sessions Chairman.

Calendar of Coming Events and Authors' Deadlines*

(*Continued from page 14A*)

11th Ann. Broadcast Symp., Willard Hotel, Washington, D. C., Oct. 6–7.

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

5th Nat'l. Symp. on Engrg. Writing and Speech, Kellogg Ctr. for Continuing Education, Michigan State Univ., East Lansing, Oct. 16–17. (DL*: July 15, J. Chapline, Philco Corp., Computer Div., 3900 Welsh Rd., Willow Grove, Pa.)

Int'l. Conf. on Ionization of the Air, Franklin Inst., Philadelphia, Pa., Oct. 16–17.

6th Ann. North Carolina Section Symp., Greensboro Coliseum, Greensboro, N. C., Oct. 19–20.

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

URSI-IRE Fall Mtg., Univ. of Texas, Austin, Oct. 23–25.

PGNS 8th Ann. Mtg., Hotel Riviere, Las Vegas, Nev., Oct. 23–26. (DL*: July 1, P. M. Uthe, Univ. of Calif., Lawrence Radiation Lab., Box 808, Livermore, Calif.)

Symp. on Instrumentation Facilities for Biomedical Res., Sheraton Fontenelle Hotel, Omaha, Neb., Oct. 26–27.

1961 Electron Devices Mtg., Sheraton-Park Hotel, Washington, D. C., Oct. 26–28.

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

6th Ann. Special Technical Conf. on Nonlinear Magnetics, Statler Hilton Hotel, Los Angeles, Calif., Nov. 6–8. (DL*: June 1, T. Bernstein, Space Technology Labs., Inc., P. O. Box 95001, Los Angeles, Calif.)

Conf. on Electrically-Exploded Wires, Kenmore Hotel, Boston, Mass., Nov. 13–14.

7th Ann. Conf. on Magnetism and Magnetic Materials, Hotel Westward Ho, Phoenix, Ariz., Nov. 13–16.

NEREM, Boston, Mass., Nov. 14–16.

Electronic Systems Reliability Symp., Linda Hall Library Auditorium, Kansas City, Mo., Nov. 14. (DL*: Sept. 1, Dr. A. Goldsmith, Wilcox Electric, Kansas City, Mo.)

PGVC Conf., Hotel Raddison, Minneapolis, Minn., Nov. 30–Dec. 1.

Eastern, Joint Computer Conf., Sheraton-Park Hotel, Washington, D. C., Dec. 12–14.

1962

8th Nat'l. Symp. on Reliability and Quality Control, Statler Hilton Hotel, Washington, D. C., Jan 9–11. (DL*: May 15, 1961, E. F. Jahr, IBM Corp., Owego, N. Y.)

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

8th Scintillation and Semiconductor Counter Symp., Shoreham Hotel, Washington, D. C., Mar. 1–2.

* DL = Deadline for submitting abstracts.

JACC Announces Additions to Program

Related meetings are being scheduled to precede the main forum of the Joint Automatic Control Conference, which will be held on June 28–30, 1961, at the University of Colorado, Boulder.

The Conference was conceived to avoid duplicate, redundant, and overlapping control meetings. Because this wide assemblage of control personnel occurs only at the JACC, it has been decided to schedule ancillary meetings, two of which are definitely arranged:

1) The Simulation Techniques Subcommittee of the AIEE Feedback Control Systems Committee will conduct a panel discussion on simulation techniques on Tuesday, June 27. Topics will range from specific techniques, such as simulating samplers and quantizers, to simulating broad industrial and economic systems. For details write to Prof. W. W. Seifert, MIT, Cambridge 39, Mass.

2) The Nonlinear Control Theory Committee of the AIEE will organize a workshop on dynamic programming on Tuesday, June 27. Topics will include adaptive control, sampled data, and industrial process control. For details write to Dr. R. E. Kuba, P.O. Box 3556, Beechwold Station, Columbus, Ohio.

On Wednesday morning, June 28, H. M. Paynter, Chairman of the Program Committee, R. K. Adams, General Conference Chairman, and D. A. Rodgers will participate in the opening introductions and announcements. At this time the American Automatic Control Council Award will be presented. Dr. M. L. Minsky, of the Massachusetts Institute of Technology, will deliver a talk on "Artificial Intelligence and Automatic Control." Dr. Kan Chen, of the Westinghouse Research Laboratory, will speak on "Modern Research in Automatic Control."

The complete program of the Joint Automatic Control Conference was given in the IRE News and Radio Notes section of the May, 1961, issue of PROCEEDINGS.

Electrically Exploded Wires Topic of Fall Conference

A conference on Electrically Exploded Wires, sponsored by the Thermal Radiation Laboratory of the Geophysics Research Directorate of the Air Force Cambridge Research Laboratories, will be held at the Kenmore Hotel, Boston, Mass., on November 13–14, 1961. Invited and contributed papers on both theory and uses of exploding wires will be presented.

For further information, contact: W. G. Chace, Thermal Radiation Lab., CRZCM, Geophysics Research Directorate, Air Force Cambridge Research Labs., Bedford, Mass.

IRE Establishes New Subsection

At its meeting on March 24, 1961, the IRE Executive Committee approved establishment of a new IRE Subsection, to be known as the Victoria Subsection of the Vancouver Section, and to encompass all of Vancouver Island in the Province of British Columbia, Canada.

Kansas City Section Schedules Fall Symposium

The Kansas City Section of the IRE will sponsor a Symposium on Electronic Systems Reliability on November 14, 1961.

The Technical Program Chairman, Dr. Arthur Goldsmith, Director of Engineering for Wilcox Electric, is planning the Symposium to stress product or "hardware" reliability; papers on MTF prediction, reliability evaluation, and process and fabrication techniques to improve reliability are planned.

To provide a stimulating environment for discussion, the Symposium *Proceedings* will be published and mailed to those who have pre-registered, at least two weeks before the Symposium. Speakers will then present only an abstract of their papers: about half of each session will be for discussion and questions and answers. An addendum to the *Proceedings* on the discussion is also being planned.

The Mid-America Electronic Conference (MAECON), usually held in November, is now scheduled only for even-numbered years; symposiums are planned for 1961 and 1963.

EIA Medal of Honor Awarded to J. B. Wiesner

Dr. Jerome B. Wiesner (S'36–A'40–SM'48–F'52), Special Assistant to the President for Science and Technology, who was chosen by the EIA Board of Directors for award of the 1961 EIA Medal of Honor for "distinguished service contributing to the advancement of the electronics industry," is the first government official to be selected for the award since its establishment in 1952. Presentation of the medal was made at the Award Dinner on May 23, 1961, during EIA's annual convention in Chicago, Ill.

Dr. Wiesner, long known in scientific and industry circles, is now on leave from his position as Director of the Research Laboratory of Electronics at the Massachusetts Institute of Technology, Cambridge. He was, until his appointment to the White House post, a Director of the Sprague Electric Company, and also has served as industry consultant to other electronic companies.

Long a leader in the rapid development of communication sciences, he was Chairman of the steering committee of the Center for Communication Sciences established at MIT in 1958 to study both man-made and natural communication systems. He also has been active in advancement of technology related to the nation's international problems both as a member of the President's Science Advisory Committee and of the committee which prepared the Gaither report. He also was Staff Director of the United States delegation to the 1958 Geneva Conference on the prevention of surprise attack.

He was born in Detroit, Mich., in 1915, and received the degrees of Bachelor of Science, Master of Science, and Doctor of Philosophy from the University of Michigan, Ann Arbor, in 1937, 1938, and 1950, respectively.

In 1940 he was appointed Chief Engineer of the Acoustical and Record Laboratory in the Library of Congress, where, under a Carnegie Corporation grant, he assisted in developing sound recording facilities and associated equipment.

Shortly after the beginning of World War II, he joined the staff of MIT's Radiation Laboratory as associate leader of the radio frequency development group. Later he became project engineer of a key radar development program. In 1945 he served on the staff of the Los Alamos Laboratory.

Eta Kappa Nu Association, the electrical engineering honor society, voted Dr. Wiesner an honorable mention as an outstanding young electrical engineer for the year 1947. In recognition of "outstanding services to his country," he was awarded the President's Certificate of Merit, the second highest civilian award, in 1948. In 1952, he was elected a Fellow of the IRE for his "outstanding contributions to radio or allied fields." He is a Fellow of the American Academy of Arts and Sciences and a member of the ASA.

Communications Conference To Meet in Philadelphia

The IRE Conference on Technical-Scientific Communications, sponsored jointly by the National Professional Group on Engineering Writing and Speech (PGEWS) and the Philadelphia Section of the IRE, will be held on September 13–15, 1961, at the Bellevue-Stratford Hotel in Philadelphia, Pa.

Seeking improved solutions to problems of communicating and utilizing technical and scientific information, the Conference has selected as its theme, "The Crisis in Technical-Scientific Communications, and What We Can Do About It."

Papers will be presented by authorities in science, industry, the military, education, publications and documentation. Conference sessions will feature subjects such as: "The High Cost of Research and Development Communication Failures," "Increasing Technical-Scientific Manpower Five to Twenty Per Cent by Improving Communications," "Literature Search and This 'Technical-Scientific Tower of Babel'," and "A Field Theory of Communication."

For further information, contact: George Boros, Program Chairman, Conference on Technical-Scientific Communications, Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia 4, Pa.

Professional Group News

At its meeting on March 24, 1961, the IRE Executive Committee approved the following new Chapters: PG on **Bio-Medical Electronics**—Rochester Chapter; PG on **Electronic Computers**—Orange County Chapter; PG on **Electronic Computers**—Syracuse Chapter; PG on **Engineering Management**—Seattle Chapter; PG on **Information Theory**—San Francisco Chapter.

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BOMAC K_u BAND MAGNETRONS

Designers of radar equipment will find Bomac Laboratories' new BLM-071 K_u-band pulse magnetron meets exacting requirements for airborne systems: lightweight, rugged, powerful. This newest contribution from Bomac is a fixed-frequency tube (15.9-16.1 kMc) rated at 100 kW peak, at 0.001 duty cycle.

Cathode structure is greatly improved over similar magnetrons. Operable at high ambient temperatures, with input/output terminals permitting pressurization to 30 psia. Special construction minimizes leakage current. High power output and low operating voltage are combined in a compact, ruggedized unit. Long life. Weight: less than 8½ lbs.

The many advantages to Bomac's BLM-071 magnetron make it readily adaptable to navigation, high-altitude mapping, airport surveillance, and similar applications. Write for full technical details.

FEATURES: Frequency 15.9-16.1 kMc.
Peak Power 100 kW.
Normal efficiency 30%.
Duty cycle 0.001 Max.
Pulse width 0.06 to 1.2 usec.

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R. P. Gifford Appointed Member of JTAC

Richard P. Gifford, Manager of the Communication Products Department of the General Electric Company, Lynchburg, Va., has recently been appointed a member of the Joint Technical Advisory Committee (JTAC) by the Committee's sponsors, the Electronic Industries Association and The Institute of Radio Engineers. He fills the vacancy created by the death of Dr. John V. L. Hogan, an active member of JTAC since its inception in July, 1948.

General Electric appointed him as Chairman of its team of experts in preparing the Company's reply to the FCC in connection with Docket No. 13522.

Mr. Gifford has been a member of the EIA (formerly RETMA) for the past 13 years and has been Chairman of the Industrial Electronics (TR) Panel for several years. He is also the Engineering Liaison Chairman of the EIA Industrial Electronics Division and a member of the EIA General Standards Committee.

He has been a member of the IRE Professional Group on Vehicular Communications since 1955 and is its present Chairman. His membership in the Professional Group on Radio Frequency Interference dates back to 1958.

More than 500 delegates— nearly a quarter of the membership—attended a convention in Sydney organized by the Australian Institution of Radio Engineers.

Dr. R. Kompfer, Director of Radio and Electronic Research at Bell Telephone Laboratories, N. J. (*right*), shows the Australian Postmaster General, C. W. Davidson (*center*), and the President of the Australian Institution of Radio Engineers, F. W. J. Orr, the traveling-wave tube which he invented.

IFIPS Announces 1962 Congress

The International Federation of Information Processing Societies (IFIPS) will hold a Congress in Munich, Germany, from August 27–September 1, 1962.

The Congress will cover all aspects of Information Processing and Digital Computers, including the following:

1) Business Information Processing, *e.g.*, data processing in commerce, industry, and administration.

2) Scientific Information Processing, *e.g.*, numerical analysis; calculations in applied mathematics, statistics, and engineering; data reduction; problems in operations research.

3) Real Time Information Processing, *e.g.*, reservation systems, computer control, traffic control, analog-digital conversion.

4) Storage and Retrieval of Information, *e.g.*, memory devices, library catalogs.

5) Language Translation and Linguistic Analysis.

6) Digital Communication, *e.g.*, encoding, decoding, error-detecting and error-correcting codes for digital data transmission.

7) Artificial Perception and Intelligence; *e.g.*, pattern recognition, biological models, machine learning, automata theory.

8) Advanced Computer Techniques, *e.g.*, logical design, logical elements, storage devices, ultra high-speed computers, program techniques, ALGOL.

9) Education, *e.g.*, selection and training of computer specialists in the use of computers, information processing as a university subject.

10) Miscellaneous subjects, *e. g.*, growth of the information processing field.

In each category it is planned to cover, where appropriate, the applications of digital computers, programming, systems design, logical design, equipment, and components.

Those wishing to submit papers are invited to send abstracts of 500-1000 words to: Dr. E. L. Harder, Westinghouse Electric Corp., East Pittsburgh, Pa., by September 15, 1961. These abstracts will be considered by the international program committee of IFIPS, and authors of selected papers will be invited to submit their complete papers (in French or English) for consideration by the program committee in March, 1962.

In adition to accepted papers, there will be invited papers, symposia, and panel discussions.

The IRE, ACM, and AIEE are represented in the International Federation of Information Processing Societies by the National Joint Computer Committee of these three societies, which also holds the Eastern and Western Joint Computer Conferences.

WESCON Adopts System To Ease Distribution Of Exhibitor Literature

WESCON'S anticipated 35,000 visitors will use special "credit cards" to request product literature at the August show and convention, it has been announced by Manager Don Larson.

Thousands of embossed plastic "inquiry cards" will be issued to WESCON registrants at three stations within the Cow Palace, San Francisco, Calif., on August 22–25, 1961. The cards will be embossed with the name, title, company affiliation, and address of the visitor

WESCON engineers and executives can simply present the inquiry card to the booth representative of any company or product line in which they are interested. The exhibitor-companies, all of whom will be supplied with imprinting machines by WESCON, will quickly record all the information on index cards also provided by the management of the show. Companies will then use the index cards for prompt mail service of product and company literature directly to the inquirer.

The service presents several major advantages to exhibitors and visitors alike. Briefcase "tonnage" will be reduced for registrants interested in getting the latest line of developments of 1180 exhibiting companies.

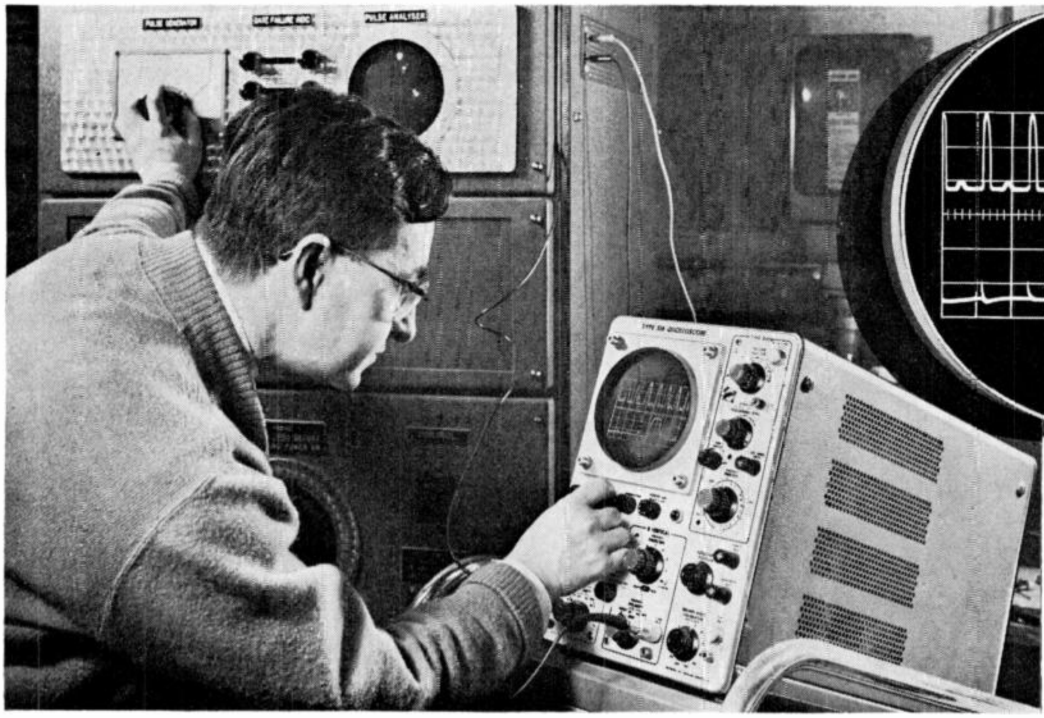
Chances of loss or misplacement of important materials during the busy four-day show are also eliminated, and exhibitor companies are free of the pressures of supplying their booth with thousands of pieces of printed information daily. Instead, corporate brochures and other materials can be sent directly to interested persons. Waste and duplication should be eliminated, along with the time required to register thousands of requests by hand.

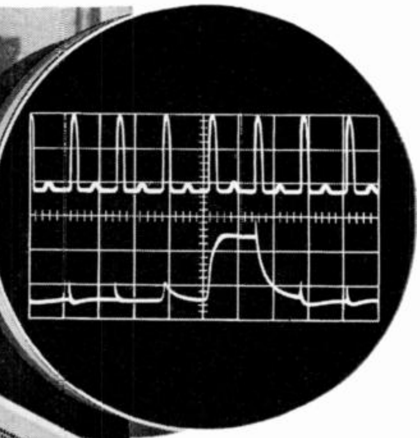
Papers Solicited for Electron Devices Meeting

Dr. W. M. Webster, General Chairman of the 1961 Electron Devices Meeting, has announced the appointment of Dr. I. M. Ross of Bell Telephone Laboratories as Program Chairman. R. K. Kilbon, of RCA, will handle publicity, and Dr. C. P. Marsden, of the National Bureau of Standards, will be Local Arrangements Chairman.

The meeting will be held at the Sheraton-Park Hotel, Washington, D. C., on October 26–28, 1961.

Papers dealing with applied research or development in the fields of electron tubes, semiconductor devices, masers, parametric amplifiers, integrated electronics, and energy conversion devices are desired. The abstracts of papers intended for presentation at this meeting are due on August 1, 1961. Contact: Dr. I. M. Ross, Bell Telephone Labs., Inc., Murray Hill, N. J., for further information.





Adjusting pulse width and height (of welding signal) from newly developed Welding Control apparatus developed by MAXIM CONTROLS COMPANY, Portland, Oregon.

By observing the dual-trace display on the Tektronix Type 516 Oscilloscope, the Project Engineer easily checks a welding gate output (lower trace) with respect to the trigger pulses (upper trace) and quickly notes any variations.

TEKTRONIX DUAL-TRACE OSCILLOSCOPE

Used in *Development* of High-Speed Welder

TYPE 516 DC to 15 MC

CHARACTERISTICS

4 Operating Modes
Both channels electronically switched — either on alternate sweeps or at a free-running rate of about 150 kc. Or each channel separately.

Vertical Amplifier
Frequency Response—dc to 15 mc (at 3 db down).
Risetime—23 nanoseconds.
Sensitivity—50 mv/div to 20 v/div in 9 calibrated steps.
Continuously variable uncalibrated from 50 mv/div to 50 v/div.
Constant Input Impedance—at all attenuator settings.

Sweep Range and Magnification
Linear Sweep—0.2 μsec/div to 2 sec/div in 22 calibrated rates.
Variable uncalibrated from 0.2 μsec/div to 6 sec/div.
Sweep Magnification—5X-magnifier extends calibrated sweep rate to 40 nsec/div.

Triggering Facilities
Fully automatic or amplitude-level selection (preset or manual) on rising or falling slope of signal, with AC or DC coupling, internal, external, or line—also, high-frequency sync to 20 mc.

Tektronix Cathode-Ray Tube
5-inch crt at 4 KV accelerating potential provides bright trace on 6 div by 10 div viewing area—each div equals 1 cm.

Amplitude Calibrator
11 square-wave voltages, from 50 mv to 100 volts, peak-to-peak, available from front panel.

Regulated Power Supplies
All critical dc voltages electronically regulated.

Size and Weight
13½″ high x 9¾″ wide x 21½″ deep—approximately 39 pounds.

Type 516 Oscilloscope (50-60 cycles) **$1000**

New, high-speed, precision welder developed at MAXIM CONTROLS COMPANY utilizes a controlled gate pulse—rather than capacitance decay—for joining high-temperature alloy materials, such as those used in manufacturing structural "honeycomb" cores.

In development of this new welder the Tektronix Type 516 Oscilloscope was used for critical timing and amplitude measurements. It was used by the Project Engineer for monitoring the time length of individual welds—since as many as six welds can be set to occur simultaneously or any number, sequentially—and for observing the constant amplitude and width of gate signals—thus assuring uniform bonds at speeds up to 2000 welds per second.

For your own research and development projects, consider the Type 516 Oscilloscope. Its dual-trace facility—with independent controls for each amplifier channel—permits you to position, attenuate, or invert the input signals as necessary for detailed analysis of their relative amplitudes, phase differences, time-delay characteristics. Its extremely reliable performance ideally suits the Type 516 for laboratory applications within the dc to 15 mc range.

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Type 516 MOD 101 (50-400 cycles) **$1035**
Type 516 MOD 108B (significantly improved writing rate at 6-KV on 6 div by 10 div viewing area—each div equals 0.85 cm) **$1075**

(prices f.o.b. factory)

For a demonstration of the Type 516 Oscilloscope in your own dual-trace (or single-trace) application, call your Tektronix Field Engineer.

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In Europe please write Tektronix Inc., Victoria Ave., St. Sampsons, Guernsey C.I., for the address of the Tektronix Representative in your country.

1961 EJCC Issues Call for Papers

The 1961 Eastern Joint Computer Conference, which will be held on December 12–14, 1961, at the Sheraton-Park Hotel in Washington, D. C., is soliciting papers on the theme, "Computers—Key to Total Systems Control."

This theme is to be treated in its broad sense; papers will be welcomed on all advances in computer hardware and concepts leading toward present and future control of industrial, government, defense, and business management systems. A few representative areas of interest follow:

- Business and Government Management
- Military Applications
- Space Projects
- Industrial Processes
- Real-Time Systems
- Self-Organizing Systems
- Man-Machine Systems
- Central/Satellite Computing Systems
- Data Acquisition
- Digital Data Communications
- Data Processing and Display
- Guidance and Control
- Information Retrieval
- Simulation and Gaming
- Numerical Analysis
- Mathematical Programming
- Computer Engineering
- Digital-Analog Devices
- Advanced Input-Output Devices
- Pattern Analysis and Recognition
- Automatic Programming
- Compilers and Translators
- Multiprogramming
- Operating Systems

Each person wishing to present a paper at the conference should submit two copies of both a 100-word abstract and a two-page summary to the Program Committee Chairman: B. G. Oldfield, IBM Federal Systems Div., 326 E. Montgomery Ave., Rockville, Md.

The deadline for submission of abstracts and summaries is June 20, 1961. All submissions should be identified by title-author, and affiliation. The full text of papers chosen for presentation must be submitted by September 1, 1961, in order to allow time for publication prior to the conference.

International Conference on Ionization of the Air

An International Conference on Ionization of the Air will be held on October 16–17, 1961, at the Franklin Institute in Philadelphia, Pa.

The Conference is being sponsored by the American Institute of Medical Climatology, a non-profit scientific association, founded in 1958 to promote the sciences of bioclimatology and biometeorology.

The technical program, now in preparation, will feature internationally recognized authorities in the field of air ionization. The topics to be covered are: physics of air ions, methods of producing and measuring air ions, relationship of various weather factors to natural ion levels, effect of air contaminants on ion levels, physiological response to air ions, psychological response to air ions, mechanisms of gaseous ion action on living matter, and possible medical applications of air ions.

The final session of the Conference will be a question and answer period at which time a panel consisting of the featured speakers will be available for discussion and questioning.

The program is now open to a limited number of prospective contributors. An abstract of 150 words for preliminary review by the program committee should be submitted before July 1, 1961, and completed manuscripts must be received before August 1. *Proceedings* of the Conference including all accepted papers will be available at a later date.

Correspondence concerning the Conference should be addressed to: American Institute of Medical Climatology, 1618 Allengrove St., Philadelphia 24, Pa.

Obituaries

Knox McIlwain (A'31–M'40–SM'43–F'48) died on March 31, 1961, in an automobile crash near Philadelphia, Pa. He was born in Philadelphia on September 4, 1897. He received the B.S. degree from Princeton University, Princeton, N. J., in 1918, and the B.S.E.E. and E.E. degrees from the University of Pennsylvania, Philadelphia, in 1921 and 1928, respectively.

Knox McIlwain

From 1921–1924 he was with the Engineering Department, Bell Telephone Company of Pennsylvania, and from 1924–1941 he was a professor at the Moore School of Electrical Engineering, University of Pennsylvania. He was associated with the Hazeltine Electronic Corporation from 1941–1957 as Chief Consulting Engineer.

Since early 1956 he had been with the Burroughs Corporation, Paoli, Pa. He served first as Assistant to the Vice President of Research and Engineering, where he was responsible at the staff level for all facets of the Corporation's contributions to the Air Force Intercontinental Ballistic Missile Program. In 1957, he was appointed Manager of the Special Products Division, and organized, directed, and was responsible for all engineering development and design projects in the fields of digital communications, weapons systems, air defense instrumentation, airborne control systems, telemetering, and automation. Later he became Manager of the Great Valley Laboratory, a major division of Burroughs Research Center, and conducted a great bulk of the Corporation's military development effort.

He was awarded over fifty patents, including an air navigation anticollision system and a passport identification system. He had numerous papers published, and included among his books are *Electrical Engineers Handbook*, *High Frequency Alternating Currents*, and *Principles of Color Television*.

In 1948 he was awarded a Certificate of Commendation by the United States Navy for his achievements during World War II. The certificate was accompanied by the citation: "This award is made for your outstanding supervision and great personal effort, as a Senior Department Head of the Hazeltine Electronics Corporation, in the design of test equipment which was of vital importance to the efficient operation of Naval radar identification equipment."

He was a past Vice President of the Institute of Navigation. Also, he was named RTMA representative of the Radio Technical Commission for Aeronautics in 1952, and one of his greatest contributions was his work with the National Television Systems Committee, Panel 8, in formulating color TV standards later approved by the FCC.

His active participation and interest in his profession were exemplified by his membership in numerous IRE committees; among them were Admissions, Board of Editors, Editorial Administrative, Editorial Review, Education, Papers, and Fellow. Also, he was a member of the following technical committees: Circuits, Electroacoustics, Electronics, Handbook, and Standards. In 1935–1936, he was Chairman of the Philadelphia Section of the IRE, and, in 1948, he received the IRE Fellow Award for his "contribution to the technical literature of radio and his activity in the field of radio aids to navigation."

Mr. McIlwain was a member of Eta Kappa Nu, Phi Beta Kappa, Sigma Psi, and Tau Beta Pi.

Oliver G. Tallman (SM'54) died on April 8, 1961, in the Pennsylvania Medical Center, Philadelphia. He was Director of the Engineering Directorate at the Rome Air Development Center, Rome, N. Y.

He was born in Asbury Park, N. J., on March 9, 1907, and attended New York University, New York, N. Y., and the Industrial College of the Armed Forces. In 1926 he entered the radio field, and specialized for several years in field testing and in the installation of marine radio telephone equipment. He jointed the Signal Corps Laboratories in 1941, again specializing in field testing, and developed the techniques of improved maintenance and troubleshooting of radar equipment.

In 1945 he transferred to the Air Force as Chief Engineer of the Florida Field Station, Watson Laboratories, and in 1947 was named Director of Technical Services at the Rome Air Development Center.

Active in community affairs since his arrival in Rome in March, 1951, he had served as Director of the 1960 United Fund Budget Committee, Chairman of the Salvation Army Advisory Board, and Chairman of the IRE National Symposium Scholarship Fund. He was a member of the Rome-Utica Chapter of the IRE, co-sponsor of the Professional Group on Engineering Management, and a current member of its Administrative Committee. Also, he had served as Chairman of the Professional Group on Military Electronics, as well as on its Administrative Committee and that of the Professional Group on Communications Systems.

HIGH PURITY ELEMENTS

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ANTIMONY	A-60	<1	<1	<1	2	<1								
✔ ARSENIC	A-60		<1											
BISMUTH	A-58		2	1		1	2						<1	
CADMIUM	A-60		<1	<1		1								
COPPER	A-58	No Impurities Detectable by Spectrographic Analysis												
GOLD	A-59		<1	<1		<1	1							
INDIUM	A-58		<1			1		<1	1					
LEAD	A-59	<1	<1	<1										
SELENIUM	A-58		<1		<1					1			<1	
SILVER	A-59		<1	<1		<1								
SULFUR	A-58											1	1	
TELLURIUM	A-58		<1	<1										
✔ THALLIUM	A-60		1	1		<1	<1					2	<1	
ZINC	A-59		<1	<1		<1								<1

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Chicago Spring Conference on Broadcast and Television Receivers

O'HARE INN, DES PLAINES, ILL., JUNE 19–20, 1961

Monday Morning, June 19

Registration

Session I

Chairman: *W. E. Swinyard, Hazeltine Electronics Corp.*

"Satellite Broadcasting," *S. C. Lutz, Hughes Aircraft Co.*

"Subscription Television System," *N. T. Watters, Zenith Electric Co.*

"Stereophonic Broadcasting," *C. G. Eilers, Zenith Electric Co.*

"Educational TV," *author to be announced.*

"A Solid-State Display Device," *S. Yando and S. Talesnick, General Telephone and Electronics Labs., Inc.*

Monday Afternoon

Session II

Chairman: *E. Hassler, Warwick Manufacturing Corp.*

"A Transistorized 27 Mc Citizens Band Transceiver," *J. Specialny, Jr., and C. Gray, Philco Corp.*

"Some Considerations in the Application of Electromechanical and Piezoelectric Filters," *H. J. Bensuly, Warwick Manufacturing Corp.*

"Subscription Television Decoder Controlling Device," *G. Morris and M. Hendrickson, Zenith Electric Co.*

"An Ultra-Low-Distortion Transistorized Power Amplifier," *H. M. Kleinman and C. F. Wheatley, RCA.*

"A Novel FM Multiplex Subcarrier Unit," *S. P. Ronsheimer, Hazeltine Electronics Corp.*

Tuesday Morning, June 20

Session III

Chairman: *C. B. Pierce, Wells-Gardner.*

"A New Compact VHF TV Tuner," *T. Gossard, W. Delp, and J. Stolte, Standard Kollsman Industries, Inc.*

"The Noise Figure of Transistor Converters," *R. R. Webster, Texas Instruments, Inc.*

"A Solid-State Electronic Tuning Receiver," *E. M. Aupperle and T. W. Butler, Jr., University of Michigan.*

"The Nuvistor Triode as a Video IF Amplifier," *K. W. Angel and J. T. Gote, RCA.*

"Dynamic Polar Display of Transfer Characteristics of Television Receivers," *P. H. Van Anrooy, Zenith Electric Co.*

Tuesday Afternoon

Session IV

Chairman: *F. Brewster, Motorola, Inc.*

"Impulse Noise Suppression in FM Receivers," *N. Parker, Motorola, Inc.*

"Design of Stereophonic Receiver for a Stereo System in the FM Band Using an AM Subcarrier," *A. J. DeVries, Zenith Electric Co.*

"A Noise-Immune Sync and AGC Circuit," *R. N. Rhodes and W. Deitz, RCA.*

"A Second Progress Report on TV Receiver Reliability," *E. H. Boden, Sylvania Electric Products, Inc.*

"Factors Affecting Over-all Performance on FM Stereophonic Receivers," *A. Csicsatka and R. Linz, General Electric Co.*

Fifth National Convention on Military Electronics (MIL-E-CON 1961)

SHOREHAM HOTEL, WASHINGTON, D. C., JUNE 26–28, 1961

More than 5000 of the nation's top engineers, scientists, and executives in the military electronics industry are expected to attend the Fifth National Convention on Military Electronics (MIL-E-CON 1961) at the Shoreham Hotel in Washington, D. C., on June 26–28.

This annual meeting is sponsored by the Professional Group on Military Electronics (PGMIL) of the IRE. Major General F. L. Ankenbrandt, USAF (Ret.), member of the Technical Staff of Defense Electronic Products, RCA, Camden, N. J., is Convention President for MIL-E-CON 1961.

At the opening session on Monday, June 26, J. H. Rubel, Deputy Director of Defense Research and Engineering, Office of the Secretary of Defense, will lead a panel discussion on "Trends in Weapons Systems Development." A substantial portion of military electronics receives its broad guidance from this phase of the military program. The Assistant Secretaries of the Army, Navy, and Air Force have been invited to participate in this panel discussion.

Rear Admiral Frank Virden, USN, Director, Naval Communications, will be the principal speaker at the Keynote Luncheon, also on Monday, sponsored by the Washington Chapter of the PGMIL. Dr E. G. Witting, Deputy Assistant Secretary of the Army (Research and Development), and Chairman of the PGMIL, will serve as Master of Ceremonies.

At a second luncheon on Wednesday, June 28, the M. Barry Carlton Award will be presented. This annual award is given to the author or authors of the best paper published in the IRE TRANSACTIONS ON MILITARY ELECTRONICS, the quarterly publication of PGMIL.

More than 100 papers on military electronics will be presented during the technical sessions which begin Monday afternoon. In addition to the 15 unclassified sessions, six classified sessions, sponsored by the Air Force Systems Command (AFSC), featuring invited papers, have been arranged by the Technical Program Committe, headed by Harry Davis, Office of the Secretary of Air Force for Research and Development.

The 1961 *Conference Proceedings* will be distributed free to all registrants at the Convention and will contain the texts of the unclassified papers, except those presented in the state-of-the-art sessions.

Displays of the latest components, instrumentation, equipment and systems of particular interest and significance in military electronics will fill the exhibit area, which has been substantially increased over last year's MIL-E-CON. Government exhibits will also be shown, including displays by the Federal Aviation Agency and NASA, in addition to the Army, Navy, Air Force and Marine Corps.

Another new feature of MIL-E-CON 1961 will be an evening of entertainment on Monday evening, replacing the customary banquet. It will be preceded by a reception in the Exhibition Hall.

The advance program follows:

Monday Morning, June 26

Research and Development Panel—Trends in Weapon System Development

Moderator: *J. H. Rubel, Deputy Director of Defense Research and Engineering; assisted by the Assistant Secretaries of the Army, Navy and Air Force.*

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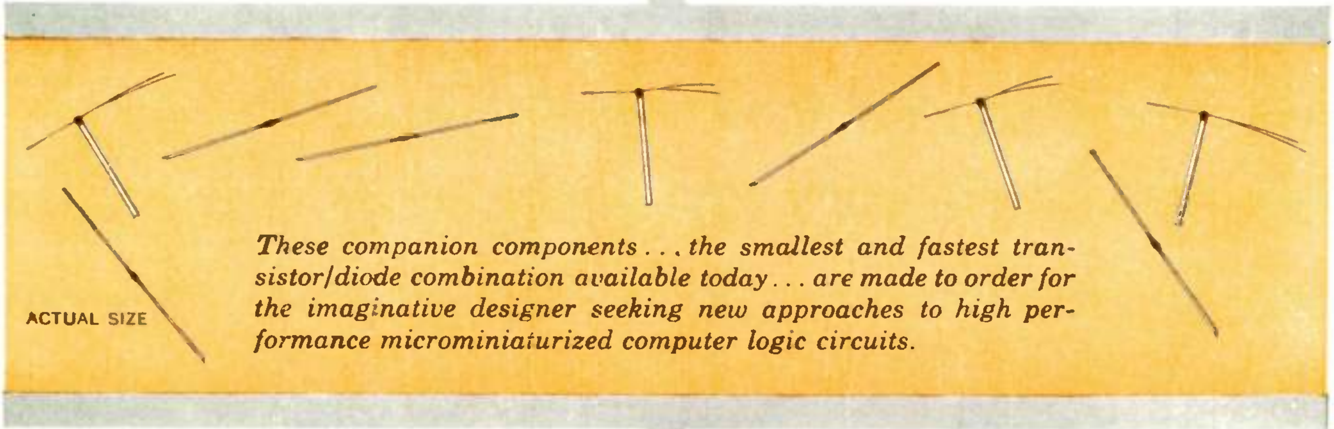

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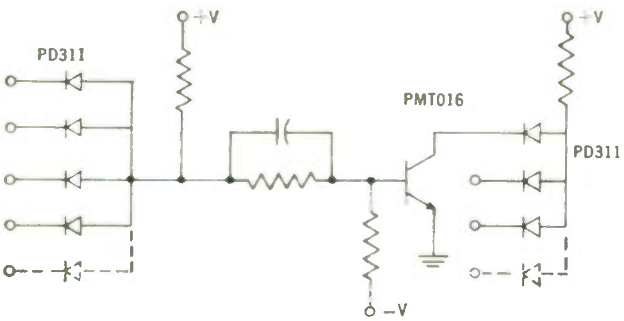

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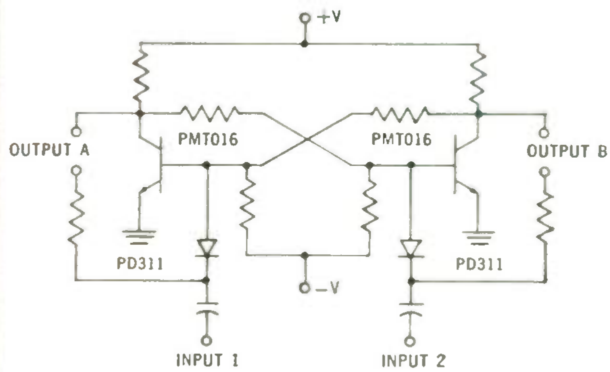

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This session sets the keynote for the Convention by outlining the trends in weapon systems development from which a great deal of military electronics will receive its broad guidance.

Monday Afternoon

Session 1.1 (Secret)—Command and Control in the Military Structure

Sponsor: *Air Force System Command.*

Moderator: *Dr. W. O. Baker, Vice President, Bell Telephone Labs.*

"The Impact of Automated Facilities on Military Command," *C. Zracket, Mitre Corp., Bedford, Mass., and W. J. Sen, Command and Control Dev. Div., AFSC, Hanscom Field, Bedford, Mass.*

"Methodology of Design," *Dr. R. Davis, Office, Naval Research, Head, Operation Research Div., Washington, D. C., and H. Bennington, Systems Dev. Corp., Santa Monica, Calif.*

"Tools and Techniques of Command and Control," *M. Rogoff, International Electric Corp., Paramus, N. J.*

"Problems of Command and Control Within the United States," *R. Clark, Office of Director of Defense Research and and Engrg., Washington, D. C.*

Session 1.2—Plasma Physics

Moderator: *Dr. Walter Kahn, Microwave Research Inst., Brooklyn, N. Y.*

"Plasma Physics in Review," *Dr. S. C. Brown, Research Lab. of Electronics, MIT.*

"Network Concepts in Plasma Wave Propagation," *Dr. N. Marcuvitz, Vice President for Research, Polytech. Inst. Brooklyn.*

"Magnetohydrodynamic Propulsion," *Dr. G. S. Janes, AVCO-Everett Research Labs.*

"Fusion Power Research," *Dr. R. G. Mills, Plasma Physics Lab., Princeton University.*

Session 1.3—Management and Packaging Techniques

Moderator: *R. I. Cole, Melpar, Inc., Falls Church, Va.*

"Military vs Contract Maintenace: A Case History," *H. W. Adams and A. S. Morton, Human Factors Dept., Mitre Corp., Bedford, Mass.*

"Power Dissipation in Microelectronic Transmission Circuits," *J. D. Meindl, U. S. Army Signal Research and Dev. Lab., Fort Monmouth, N. J.*

"Microminiaturization Design Techniques," *P. E. Boron, Hughes Aircraft Co., Culver City, Calif.*

"Advanced Management Systems for Advanced Weapon Systems," *Willard Fazar, Dept. of the Navy, Washington, D. C.*

"A Versatile Modular Packaging Concept," *Joseph Ritter, Bendix Corp., Towson, Md.*

Session 1.5—Instrumentation

Moderator: *Dr. Arnold Shostak, Office of Naval Res., Washington, D. C.*

"Remote Measurement of Wind Velocity by the Electromagnetic-Acoustic Probe," *R. W. Fetter, Midwest Research Inst., Kansas City, Mo.*

"Some Circuit Considerations for Application of the Avalanche Injection Diode," *A. P. Schmid Jr., Raytheon Res. Div., Waltham, Mass.*

"A Comparative Review of Phase Measurement Methods at Microwave Frequencies," *R. A. Sparks, Applied Physics Lab., The Johns Hopkins University, Silver Spring, Md.*

"Instrumentation for High Altitude and Space Vehicles," *Dr. S. C. Stephan, Jr., Bell Aerosystems Co., Bell Aerospace Corp., Buffalo, N. Y.*

"A Miniaturized High "g" Telemetering System," *Essad Tahan, Sylvania Electronic Systems, Applied Res. Lab., Waltham, Mass.*

Session 1.5—Radar Technology (Secret)

Moderator: *C. R. Kilgare, Applied Research Dept., Westinghouse Electric Co., Baltimore, Md.*

"A Radar Received Using Traveling-Wave Maser and Traveling-Wave Tube Amplifiers," *G. T. Kresan, Bell Telephone Labs., Whippany, N. J.*

"Bistatic Ranging," *G. A. Klein, The Mitre Corp., Electronic Warfare Dept., Bedford, Mass.*

"Recent Advances in Megawatt Klystrons for Broadband Systems," *Arlindo Jorge, Sperry Gyroscope Co., Electronic Tube Div., Great Neck, L. I., N. Y.*

"Radar Antenna Side Lobe Suppression," *L. Greenberg, Burroughs Corp., Paoli, Pa.*

"A Radar Simulation Study," *James Misho, The Mitre Corp., Lexington, Mass.*

"Space Tapered Antenna Arrays," *The Bendix Corp., Radar Research and Dev., Baltimore, Md.*

Tuesday Morning, June 27

Session 2.1—Oceanic Surveillance Technology (Secret)

Sponsor: *Office of Chief of Naval Operations.*

Moderator: *Dr. R. O. Burns, Office of the Chief of Naval Operations, Washington, D. C.*

"Oceanic Space Environment," *R. D. Fusselman, Capt., USN, U. S. Navy Hydrographic Office, Suitland, Md.*

"Methods of Subsurface Detection," *B. Rosenberg, Technical Analysis and Advisory Group, Office of the Chief of Naval Operations, Washington, D. C.*

"Ocean Surveillance," *L. M. Treitel, Bureau of Ships, Washington, D. C.*

"Ocean Signal Processing," *P. L. Stocklin, Office of Naval Research, Washington, D. C.*

Session 2.2—Radio and Radar Astronomy

Moderator: *Dr. J. P. Hagen, Director, Office for United Nations Conf., National Aeronautics and Space Administration, Washington, D. C.*

"Expanding the Horizons of Astronomy by Radio Techniques," *Prof. F. T. Haddock, The Observatory, University of Michigan, Ann Arbor, Mich.*

"Investigating the Ionospheres of Other Planets," *Prof. K. M. Siegel, Radiation Lab., Dept. of Elec. Engrg., The University of Michigan, Ann Arbor, Mich.*

"The 1000 Foot Dish and Its Capabilities," *Dr. W. E. Gordon, Cornell University, Ithaca, N. Y.*

"Radio Emission of the Planets," *Cornell Mayer, Naval Research Lab., Washington, D. C.*

"Venus Radar," *Dr. Eberhardt Rechtin, Jet Propulsion Lab., Pasadena, Calif.*

Session 2.3—Reliability and Electromagnetic Compatibility

Moderator: *E. J. Nucci, Office, Director of Defense Research and Engrg., Dept. of Defense, Washington, D. C.*

"The Reliability of Repairable Complex Systems," *R. S. Dick, International Electric Corp., Paramus, N. J.*

"Agree Reliability Testing of Air to Air Identification Equipment," *J. D. Burr and W. L. Allen, Hughes Aircraft Co., Engrg. Div., Culver City, Calif.*

"Techniques for Achieving Operational Reliability and Maintainability in Digital Computers," *T. B. Lewis, IBM Corp., Federal Systems Div., Owego, N. Y.*

"Semi-Conductor Reliability in Switching Systems," *W. P. Karas, Stromberg-Carlson Co., Rochester, N. Y.*

"The Progress of the Department of Defense Electromagnetic Compatibility Program," *R. F. Brady, Office of the Chief Signal Officer, Dept. of the Army; W. Dean, Jr., Assistant Head, Radio Frequency Spectrum Div., Office of Chief of Naval Operations; Henry Randall, Office of the Director of Defense Research and Engrg., Office of the Secretary of Defense; Col. O. J. Schulte, USAF, Chief of Electronics and Reconnaissance Div., Deputy Chief of Staff/Development.*

Session 2.4—Communications I

Moderator: *Ralph Clark, Assistant Director of Research and Engrg. for Communications, Office, Director of Defense Research and Engrg., Washington, D. C.*

"Recent Developments in the Field of Digital Data Communications," *F. R. Cronin, ACF Industries, Inc., Riverdale, Md.*

"Recent Developments in the Envelope Elimination and Restoration High Efficiency SSB System," *L. R. Kahn, Kahn Research Labs., Inc., Freeport, L. I., N. Y.*

"The Defense National Communications Control Center," *C. D. May, Jr., Defense Communication Agency, Naval Service Ctr., Washington, D. C.*

"System Performance of a Communications Control Center," *Jerome Hoffman and David Moses, Systems Development Corp., Paramus, N. J.*

"On the Survivability of Communications Systems," *T. G. Williams, Lt. Col., Rome Air Dev. Ctr., Griffiss AFB, N. Y.*

Tuesday Afternoon

Session 3.1—Military Systems (Secret)

Moderator: *D. J. McLaughlin, Applications Research Div., U. S. Naval Research Lab., Washington, D. C.*

"Weapon System Survivability: First Requirement of Command and Control," *J. P. Mullen, AF Cambridge Research Labs., Air Force Research Div., L. G. Hanscom Field, Bedford, Mass.*

"Evaluation of Gain Patterns and Side Lobe Response of Focused and Unfocused Synthetic Arrays," *R. C. Heimiller, Hughes Aircraft Co., Engrg. Div., Culver City, Calif.*

"A Pocket-Sized Reconnaissance Receiver and Analyzer," *J. C. deBroekert,*


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"Simulation of a Low Altitude Air Battle Over Varied Terrain," *A. A. McGee, IBM Corp., Federal Systems Div., Owego, N. Y.*

"Signal-Sorting at the Receiver in Radar Reconnaissance," *T. C. Jackson, Stanford University, Stanford Electronics Labs., Stanford, Calif.*

"The AN/APQ-80 Low Altitude Automatic Terrain Following System," *D. E. Corp, Cornell Aeronautical Lab., Inc., Buffalo, N. Y.*

Session 4.2—Computer Technology

Moderator: *Dr. Sam Alexander, National Bureau of Standards, Washington, D. C.*

"Evolution of Computer Concepts and Techniques," *Dr. W. H. Ware, RAND Corp., Santa Monica, Calif.*

"Review of Current Capabilities," *Dr. Sam Alexander, National Bureau of Standards, Washington, D. C.*

"Trend in Research on Computer Components," *R. D. Elbourn, National Bureau of Standards, Washington, D. C.*

Session 3.3—Radar I

Moderator: *A. Weinstein, Chief Surveillance Div., Office, Director of Defense Research and Engrg., Washington, D. C.*

"A Note on the Spatial Information Available from Monopulse Radar," *D. B. Anderson and D. R. Wells, Autonetics, Anaheim, Calif.*

"Breadboard COLIDAR (Coherent Light Detecting and Ranging) Systems," *M. L. Stitch, F. J. Meyers, J. H. Morse, and E. J. Woodbury, Hughes Aircraft Co., Engrg. Div., Culver City, Calif.*

"Comparison of Short ARC Tracking Systems for Orbital Determination," *R. Lieber and S. Shucker, RCA, Defense Electronic Products, Moorestown, N. J.*

"ECM System Analysis," *J. T. Oblinger, HRB Singer, State College, Pa.*

"Some Advances in CW Radar Techniques," *Dr. J. D. Harmer and W. S. O'Hare, Missile and Space Div., Raytheon Co., Bedford, Mass.*

Session 3.4—Communications II

Moderator: *Claud Beckham, Executive Assistant to Vice President—Defense, American Telephone and Telegraph Co., Washington, D. C.*

"Jamming Implications of Experiments on Interrupted Speech," *D. W. Tufts, Harvard University, Div. of Engrg. and Applied Physics, Cambridge, Mass.*

"Attenuation of Electromagnetic Radiation from Polaris," *J. D. Meyer, Electromagnetics Research, Lockheed Missiles and Space Div., Sunnyvale, Calif.*

"Solid State, S-Band Single Sideband Suppressed Carrier Transmitter," *W. J. Maciag, Sylvania Electronic Systems, Amherst, N. Y.*

"Status of Radiation Effects Studies," *E. E. Conrad, Bethesda, Md.*

"A New Digital Communication System—Modified 'Diphase'," *D. Douglas, G. Aaronson, and G. Meslener, RCA, Defense Electronic Products, New York, N. Y.*

Session 3.5—Aerospace Technology

Moderator: *Dr. C. T. Morrow, Aerospace Corp., Los Angeles, Calif.*

"Rocket Propulsion," *Hank Burlage, National Aeronautics and Space Administration, Washington, D. C.*

"High-Energy Rocket Propulsion," *C. J. Wang, Aerospace Corp., Los Angeles, Calif.*

"Attitude Control and Guidance," *J. F. Shea, AC Spark Plug Co., Milwaukee, Wis.*

"Payloads and Re-entry Vehicles," *L. D. Ely, Aerosapce Corp., Los Angeles, Calif.*

"Space Science and Instrumentation," *Martin Swetnik, National Aeronautics and Space Administration, Washington, D. C.*

Wednesday Morning, June 28

Session 4.1—Satellite Communication Session

Moderator: *Brig. Gen. G. P. Sampson, Director of Operations, Defense Communication Agency, Washington, D. C.*

"Communication Satellite Need Now and Future," *Ralph Clark, Assistant Director for Communications, Office, Director of Defense Research and Engrg., Washington, D. C.*

Limitations in Technology—Problem Areas," *Dr. Charles Morrow, Aerospace Corp., Los Angeles, Calif.*

"Various Present Approaches," 1) Passive Satellite—*Maj. Don Nowakowski, Air Force System Command, Washington, D. C.* 2) Active Satellite—*Sam Brown, Advent Management Agency, Washington, D. C.* 3) Lofti—*Emerick Toth, Radio Div., Naval Research Lab., Washington, D. C.*

"Future Prospects," *A. G. Kandoian, IT&T Co., Nutley, N. J.*

A panel discussion and question period will follow this session.

Session 4.2—Space Physics

Moderator: *Robert Jastrow, Chief, Theoretical Div., National Aeronautics and Space Administration, Washington, D. C.*

"Introductory Statement on Fundamental Scientific Problems in Space Research," *Robert Jastrow, NASA, Washington D. C.*

"Measurement of Fields and Particles in Space: Sun/Earth Relationships," *C. P. Sonnet, NASA, Washington, D. C.*

"Lunar and Interplanetary Spacecraft Development," *Orin Nicks, NASA, Washington, D. C.*

Session 4.3—Radar II

Moderator: *J. Whitman, Institute of Defense Analysis, Washington, D. C.*

"A Wide-Open Frequency-Reading Receiver," *J. L. Grigsby, Applied Technology, Inc., Palo Alto, Calif.*

"Ground Clutter and its Calculation for Airborne Pulse Doppler Radar," *G. R. Hetrich and S. D. Coleman, Air Arm Div., Westinghouse Electric Corp., Baltimore, Md.*

"Automatic Tracking Considerations for Ballistic Targets," *S. Adelman and S. M. Shinners, Sperry Gyroscope Co., Div. of Sperry Rand Corp., Surface Armament Div., Great Neck, N. Y.*

"Limitations on Dynamic Range and Multitarget Resolution in Search or Track Radars due to Distortion," *Dr. W. L. Rubin and J. V. DiFranco, Surface Armament Div., Sperry Gyroscope Co., Great Neck, L. I., N. Y.*

"A Target Angle Computation Method for Multi-Beam Radars," *H. C. Kreide and J. V. Popolo, Missile and Space Div., Raytheon Co., Bedford, Mass.*

"Pulse-to-Pulse Target Correlation with Range Only Data," *D. F. Allen, C. E. Dantas and H. C. Kreide, Missile and Space Div., Raytheon Co., Bedford, Mass.*

Session 4.4—Military Systems

Moderator, *A. Wimer, Chief Scientist, Air Force Systems Command.*

"Automation in Precision Plotting and Drafting," *C. W. Hargens, III, The Franklin Institute, Philadelphia, Pa.*

"A Position Difference System Using Satellite Doppler Signals," *Leonard Farkas, RCA, Missile and Surface Radar Div., Moorestown, N. J.*

"Utilization of UDOFTT (Universal Digital Operational Flight Trainer Tool) in Training Research," *D. W. Poole, U. S. Naval Training Device Ctr., Port Washington, N. Y.*

"AUDIT, An Automatic Unattended Detection Inspection Transmitter for the Monitoring of the Nuclear Test Ban," *C. C. Abt, Advanced Dev. Lab., Raytheon Weapons and Space Div., Bedford, Mass.*

"Automatic Detection of Changes in Reconnaissance Data," *Dr. Azriel Rosenfeld, Budd Electronics, Inc., Long Island City, N. Y.*

Wednesday Afternoon

Session 5.1—Ballistic Missile Defense (Secret)

Sponsor: Advanced Research Projects Agency.

Moderator: *A. Rubenstein, Director, Project Defender.*

"Review of Project Defender," *A. Rubenstein, ARPA, Washington, D. C.*

"Ballistic Defense Systems," *J. L. Crone, ARPA, Washington, D. C.*

"Re-Entry Physics," *K. Cooper, ARPA, Washington, D. C.*

"Radar," *J. Whitman, ARPA, Washington, D. C.*

"Ionospheric Research," *C. W. Cook, ARPA, Washington, D.C.*

"Kill Mechanisms," *Roy Weidler, ARPA, Washington D. C.*

"Infra-Red Background," *R. Zirkind, ARPA, Washington, D. C.*

Summary: *Dr. J. P. Ruina, Director, ARPA, Washington, D. C.*

Session 5.2—Modern Low Noise Receiving Systems

Moderator: *Dr. Robert Adler, Vice President and Associate Director of Research, Zenith Radio Corp., Chicago, Ill.*

"Noise Concepts Applied to Receiving Systems," *M. T. Lebenbaum, AIL Div. of Cutler-Hammer, Inc., Melville, L. I., N. Y.*

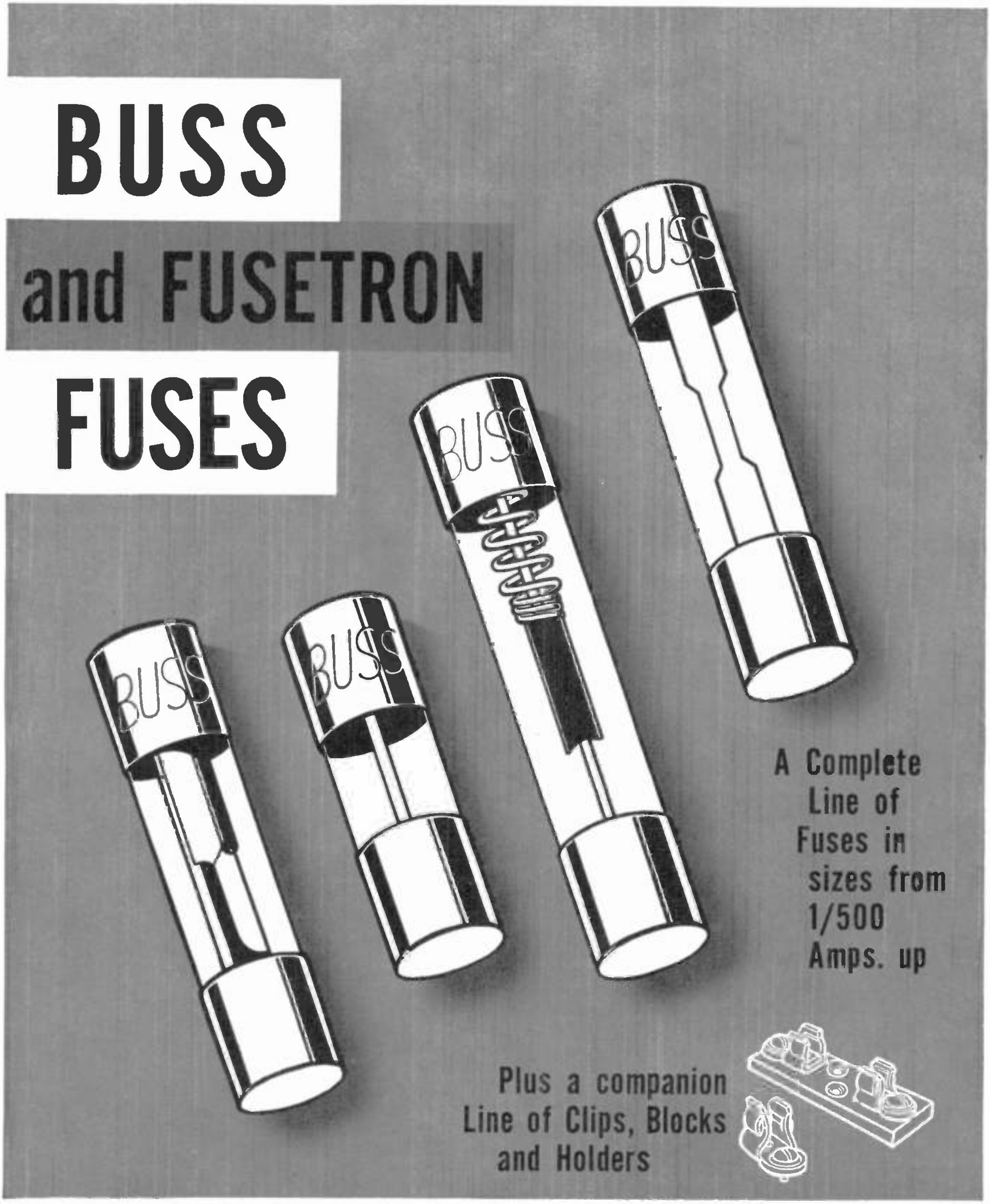
"Survey of Modern Low Noise Amplifiers," *Dr. Glen Wade, Spencer Lab., Raytheon Co., Burlington, Mass.*

"Recent System Application of Low Noise Amplifiers," *Dr. R. H. Kingston, MIT Lincoln Lab., Bedford, Mass.*

Session 5.3—Data Handling

Moderator: *Dr. L. Van Atta, Special Assistant to Director of Defense—Research and Engrg., Washington, D. C.*

"Polynomial Filtering of Signals," *F. W.*



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Nesline, Jr., Raytheon Co., Missile Systems Div., Bedford, Mass.

"Multiple Monitoring of Single Events," *G. R. Cooper, Professor of Elec. Engrg., Purdue University, Lafayette, Ind.,* and *P. M. Kelly, Aeronutronic Div. of Ford Motor Co., Newport Beach, Calif.*

"High Speed Data Sorting Using Magnetic Drums and a Large Electrostatic CRT Display," *E. B. Carne and B. G. Tragub, Ground Data Handling Equipment Lab., Melpar, Inc., Falls Church, Va.*

"Target Assignment Models," *P. J. Queeney, Systems Criteria Section, The Martin Co., Baltimore, Md.*

"Computer-Radar Environment Interplay," *R. B. Angus, R. R. Fidler, and P. A. Marino, Sylvania, Data Systems Operations, Needham, Mass.*

Session 5.4—Electron Devices

Moderator: *K. Garoff, Director, Electron Tube Div., US Army R&D Labs., Fort Monmouth, N. J.*

"A 2200 Megacycle Transmitter for Sattellites," *W. P. Quinlivan and S. E. Smith, Radiation, Inc., Melbourne, Fla.*

"Microwave Delayed Automatic Gain Control Applied to Traveling Wave Tube Systems," *L. Van Brunt and N. W. Hansen, IT&T Federal Labs., Nutley, N. J.*

"An X-Band Helix-Type TWT for Satellite Applications," *J. H. Herman, Electronic Tube Div., Sperry Rand Corp., Great Neck, L. I., N. Y.*

"Characteristics of a Backward Wave Oscillator Employing an External Feedback Circuit," *S. J. Ameen and Robert Plumridge, Raytheon Co., Microwave and Power Tube Div., Spencer Lab., Burlington, Mass.*

"A Quick Heating Rugged, UHF, Ceramic Pencil Triode for Missiles and Satellites," *S. W. Bogaenke and O. Johak, Jr., RCA, Harrison, N. J.*

Third National Symposium on Radio Frequency Interference

SHERATON-PARK HOTEL, WASHINGTON, D. C., JUNE 12–13, 1961

Monday Morning, June 12

Opening Remarks: *D. R. J. White, Symposium Chairman.*

Key Note Address: "The National Electromagnetic Compatibility Program," *Maj. Gen. J. Dreyfus.*

Session I—The RFI Problem from the Users' Viewpoint

Chairman: *E. Allen, FCC.*

"FAA Radar and ATC Interference," *H. Burton, FAA.*

"Missile Range Interference," *R. Jones, Frequency Coordinator, AMR.*

"Naval Radiation Hazards," *J. Roman, RCA Service Co.*

"Tactical Communications Interference."

"Military Radar Interference."

"Radio Astronomy Interference."

"Commercial-Industrial Interference."

Panel Discussion—Seven speakers and three invited panelists.

Monday Afternoon

Session II—Data Needs and Data Formats

Chairman: *S. Bailey, Jansky and Bailey, Inc.*

"Transmitters and Antennas," *H. Sachs, Armour Research Foundation.*

"Receivers," *K. Heisler, Jansky and Bailey, Inc.*

"Propagation Media," *W. Critchlow, National Bureau of Standards.*

Panel Discussion—Three speakers and five invited panelists.

Tuesday Morning, June 13

Session III—Instrumentation Needs and Instrumentation Limitation

Chairman: *H. Dinger, NRL.*

"Instrumentation Needs," *D. Ports, J. Hill, and K. Heisler, Jansky and Bailey, Inc.*

"Instrumentation Needs," *J. G. Holey and C. Blakley, Georgia Institute of Technology.*

"Instrumentation Limitation Below 1GC," *Dr. Haber, University of Pennsylvania, and Eckersley, Ark Electronics, Inc.*

"Instrumentation Limitations above 1GC," *Dr. F. Morris, Electro-Mechanics, Inc.*

Panel Discussion—Four speakers and four invited panelists.

Tuesday Afternoon

Session IV—Progress in Interference and Compatability Programs

Chairman: *Dr. R. Showers, University of Pennsylvania.*

"Interference Coordination Aspects of Satellite Communications Systems," *Dr. S. G. Lutz, Hughes Research Lab.*

"Dark Noise Generation of Super Power Tubes," *J. T. Coleman, RCA.*

"Predicting Power Transfer Between Large Aperture Antennas at Close Range," *E. Jacobs, University of Pennsylvania.*

"Modeling Techniques for Interference Measurements," *C. R. Miller, RADC, and J. Pullara, Melpar, Inc.*

"A Sampling Technique for the Measurement of Multimode Harmonic Power," *E. M. T. Jones and E. D. Sharp, Stanford Research Institute.*

The following series of tutorial papers will run concurrently with the main program, and will cover selected subjects in the fields of interference prevention and fixes, instrumentation advances, and system problems involving prediction and control.

Session I—Shielding, Filtering, and the Near-Field Problem

"Enclosure Shielding in Radio Interference," *C. B. Pearlston, Jr., Nortronics.*

"RF Shielding Analogies," *O. P. Schreiber, Technical Wire Products, Inc.*

"Effective Broadband Filtering for Interference Elimination in the Frequency Range from 10 Mcps to 10,000 Mcps," *H. M. Schlicke, Allen-Bradley Co.*

"Interference Aspects of Fresnel Region Phenomena," *B. Lindeman, Rome Air Development Center.*

"Alternate-Bonding and Radio Interference," *C. B. Pearlston, Nortronics.*

Session II—Instrumentation and Measurement Techniques

"Bandwidth Relationship in IF Amplifiers," *R. B. Schulz, Armour Research Foundation.*

"RF Susceptibility Testing Techniques for Airborne Electronic Equipment," *D. B. Clark, Douglas Aircraft.*

"Interference Instrumentation," *C. R. Miller, Rome Air Development Center.*

"Instrumentation for the Measurement of Extremely Low Levels of Radiated Interference," *J. P. Rutsey, The Hinchman Corp.*

"The Current Probe—A New Device in the Field of Radio Interference Measurement," *H. E. Vlfers, USASRDL, Fort Monmouth.*

"Alternate-Documenting Probable Errors of Measured C-E Parametric Data," *D. R. J. White, Don White Associates.*

Session III—Systems Problems

"Some Practical Approaches in the Control of Interference in Airborne Weapons Systems," *A. E. J. Dionne, Grumman Aircraft.*

"Irradiation-Susceptibility Nomograph," *F. Kugler and A. R. Kall, Ark Electronics, Inc.*

"Government Regulation of Unlicensed RF Equipment as a Means of Controlling RFI," *L. G. Whipple and H. Garlan, FCC.*

"Frequency Management in the Army Electromagnetic Compatibility Program," *C. A. Gregory, U. S. Army Radio Frequency Engineering Office.*

"A Discussion of the Site Effect Problem," *D. C. Ports and T. R. Evans, Jansky and Bailey, Inc.*

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30.8 Methods of Measurement of Recording and Reproducing Transducer Characteristics

4. Circuits

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J. G. Brainerd
C. H. Page
A. R. D'Heedene
E. H. Perkins
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W. H. Huggins
A. P. Stern
R. Kahal
J. J. Suran
B. K. Kinariwala
W. N. Tuttle
H. L. Krauss
L. Weinberg

4.2 Linear Lumped-Constant Passive Circuits

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J. A. Aseltine
G. L. Matthaei
R. Kahal
J. G. Truxal

4.3 Circuit Topology

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R. L. Dietzold
E. A. Guillemin
S. Goldman
J. Riordan

4.4 Linear Varying-Parameter and Non-Linear Circuits

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J. G. Kreer, Jr.
C. H. Page
J. R. Weiner

4.7 Linear Active Circuits Including Networks with Feedback Servomechanism

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E. J. Angelo, Jr.
J. M. Manley
W. A. Lynch
C. F. Rehberg

4.8 Circuit Components

A. R. D'Heedene, *Chairman*

4.9 Fundamental Quantities

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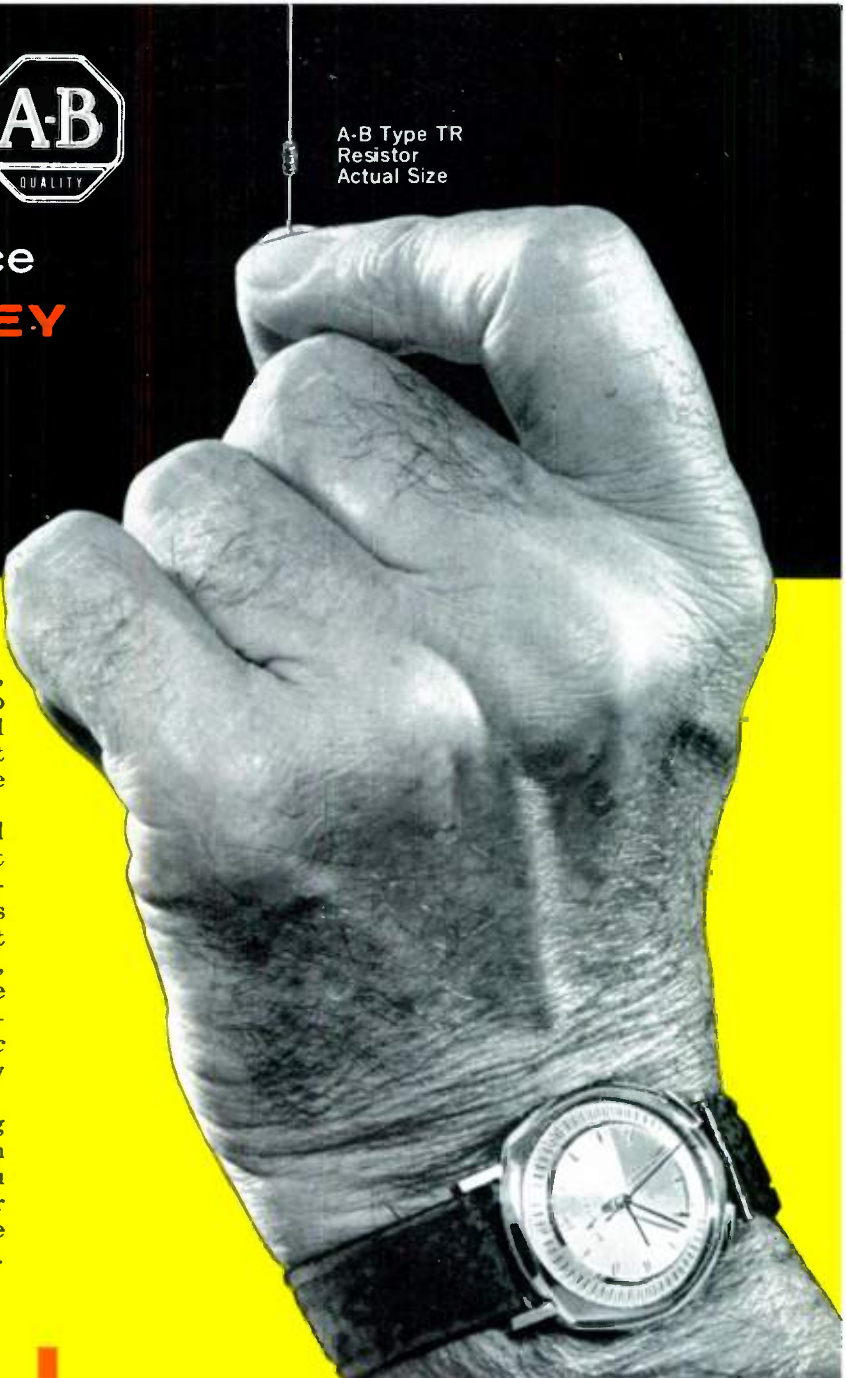
P. F. Ordung
J. D. Ryder

Accutron*...

new electronic timepiece uses ALLEN-BRADLEY Type TR Miniature Composition Resistors



With its miniature tuning fork and electronic circuit, Accutron introduces an entirely new principle to timekeeping—one which promises unprecedented wrist timepiece accuracy. Strapped to your wrist, it is guaranteed not to gain or lose more than one minute a month.

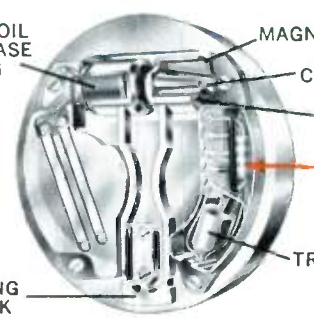
Allen-Bradley Type TR tiny resistors enabled Accutron designers to achieve the required circuit miniaturization for a wrist timepiece—*without sacrificing reliability*. This circuit controls the 360 pulses of power each second—31 million per day—that drive the tuning fork. Although incredibly small, these Type TR miniature composition resistors are made by Allen-Bradley's exclusive hot molding process that *guarantees complete freedom from catastrophic failures!* A-B Type TR resistors are conservatively rated 1/10 watt at 70°C.

There are also other Allen-Bradley space-saving potentiometers, capacitors, and h-f filters that can help solve your miniaturization problem. And you obtain the same reliability for which the larger Allen-Bradley components have earned a world-wide reputation. For full details, send for Publication 6024.

* TRADEMARK BULOVA WATCH CO., INC.

A-B HOT MOLDED COMPOSITION RESISTORS

DRAWING OF ACCUTRON SHOWS BASIC MECHANISM

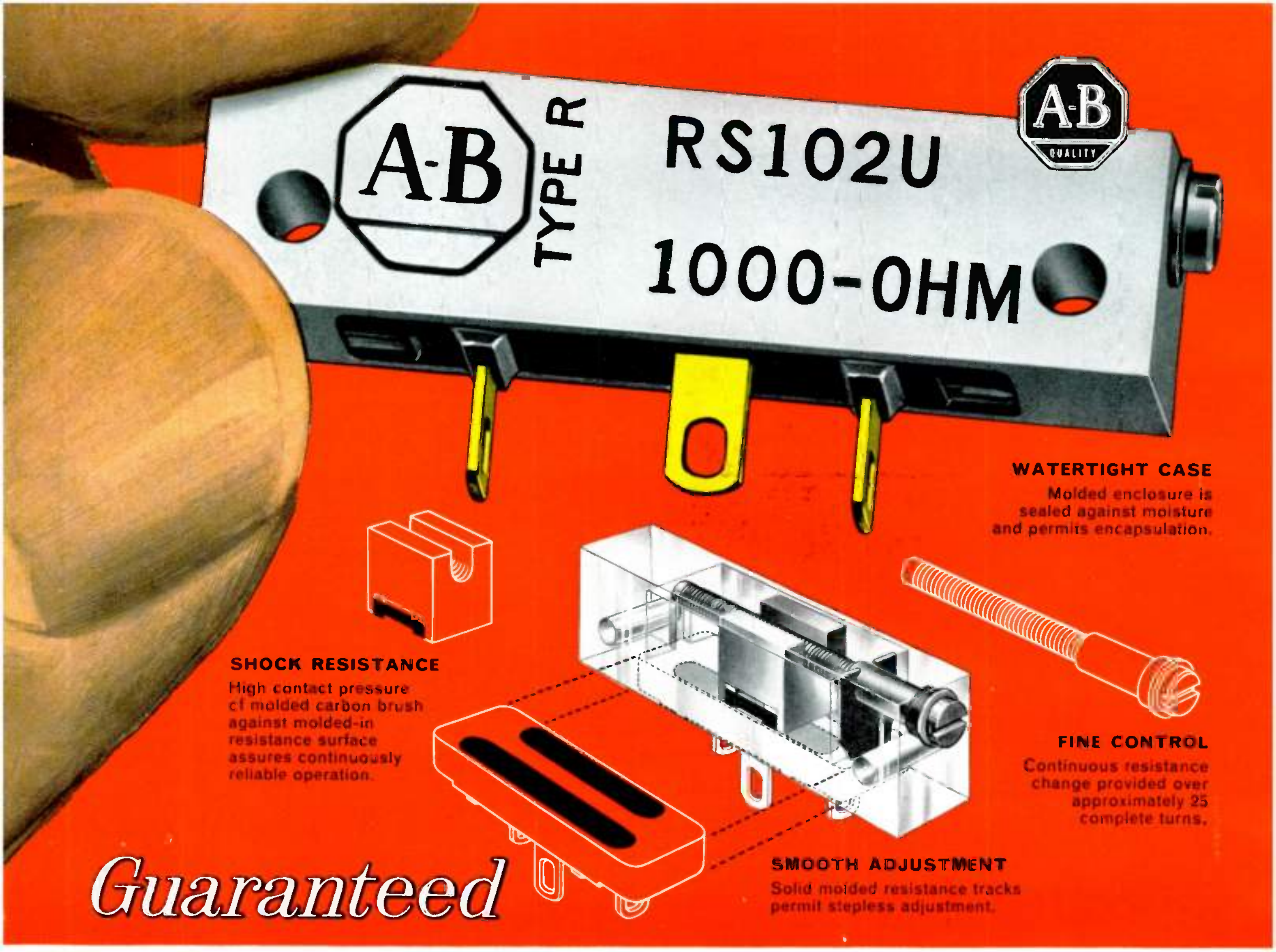


ALLEN-BRADLEY

Quality Electronic Components

Allen-Bradley Co., 222 West Greenfield Avenue, Milwaukee 4, Wisconsin • In Canada: Allen-Bradley Canada Ltd., Galt, Ontario

WRH



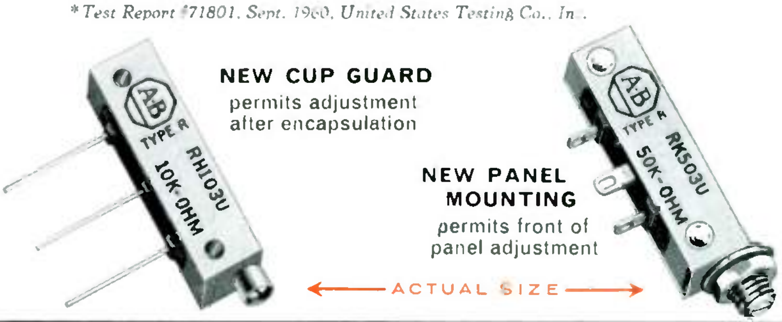
Reliable Performance and Stable Settings

UNDER EXTREME ENVIRONMENTAL CONDITIONS

In critical applications, Allen-Bradley Type R adjustable fixed resistors are *without equal*. For example, in recent tests* Type R resistors successfully withstood acceleration, shock, and vibration *five times* better than the latest MIL Spec requirements. Such wide margin of safety is your assurance of complete reliability. Virtual indestructibility is obtained through an *exclusive* Allen-Bradley process in which the solid resistance elements and the insulating mounting are hot molded into one integral unit. The moving element is self-locking for absolutely stable settings. Also, the Type R control allows "stepless" adjustment of its resistance.

The molded case of the Type R control is watertight and dust-tight. Rated ¼ watt at 70°C, these Type R controls are available in values from 100 ohms to 2.5 megohms.

** Test Report #71801, Sept. 1960, United States Testing Co., Inc.*



3-61-E

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QUALITY ELECTRONIC COMPONENTS

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7.1 Tubes in Which Transit-Time is Not Essential

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7.3.1 Methods of Test for TR and ATR Tubes

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7.3.2 Gas-Filled Radiation Counter Tubes

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7.4 Camera Tubes, Phototubes, and Storage Tubes in Which Photo-Emission is Essential

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S. E. Webber

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7.5.2 Operating Measurements of Microwave Oscillator Tubes

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29.1 NOMENCALTURE

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29.2 METHODS OF MEASUREMENT

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17.11 AM-FM Broadcast Receivers

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15.3 Double Sideband AM Transmitters

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15.5 Single Sideband Radio Communication Transmitters

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19.5 Flutter

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31.2 System Reliability Engineering

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31.1 Symbology

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31.4 Methods of Measurement

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31.5 Reliability Prediction

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28.0.1 General Definitions

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R. M. Ryder

28.0.2 Cryogenic Devices

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S. K. Ghandhi, *Chairman* (AIEE)

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C. W. Mueller
R. L. Pritchard
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D. E. Sawyer
A. C. Sheckler
A. P. Stern
R. L. Trent

28.4.2 Methods of Test for Transistors for Linear CW Transmission Service

A. Coblenz, *Chairman*

28.4.3 Definitions and Letter Symbols of Semiconductors

B. J. Rothlein, *Chairman*

28.4.4 Methods of Test for Semiconductor Devices for Large-Signal Applications

R. L. Trent, *Chairman*

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R. S. Hill
W. D. Reiner
V. R. Saari

28.4.5 Methods of Test for Bulk Semiconductors

E. N. Clarke, *Chairman*

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D. C. Cronemeyer
I. Drukaroff
A. Kestenbaum
M. F. Lamorte
J. Mandelkorn
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28.4.6 Photo Devices

D. E. Sawyer, *Chairman*

M. Adelman
J. Bramley
G. Dermit
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A. J. Hodges
M. L. Lamorte
F. A. Reid

28.4.7 Transistor Internal Parameters

R. L. Pritchard, *Chairman*

R. B. Adler
J. B. Angell
J. M. Early
W. M. Webster

28.4.8 Semiconductor Devices Symbols

S. K. Ghandhi, *Chairman*
B. J. Rothlein

28.4.9 Semiconductor Diode Definitions

C. H. Knowles, *Chairman*

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D. C. Dickson
B. Jacobs
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28.4.10 Semiconductor Diode Measurements

B. Jacobs, *Chairman*

A. Bakanowski
R. Doyle
D. R. Fewer
J. Fishel
J. H. Forster (alternate)
J. Gillette
C. Irwin (alternate)
E. F. Platz
A. C. Sheckler

28.4.12 Definitions and Methods of Test for P-N-P-N Devices

J. M. Goldey, *Chairman*
W. T. Matzen, *Secretary*

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28.4.13 High Frequency Diodes

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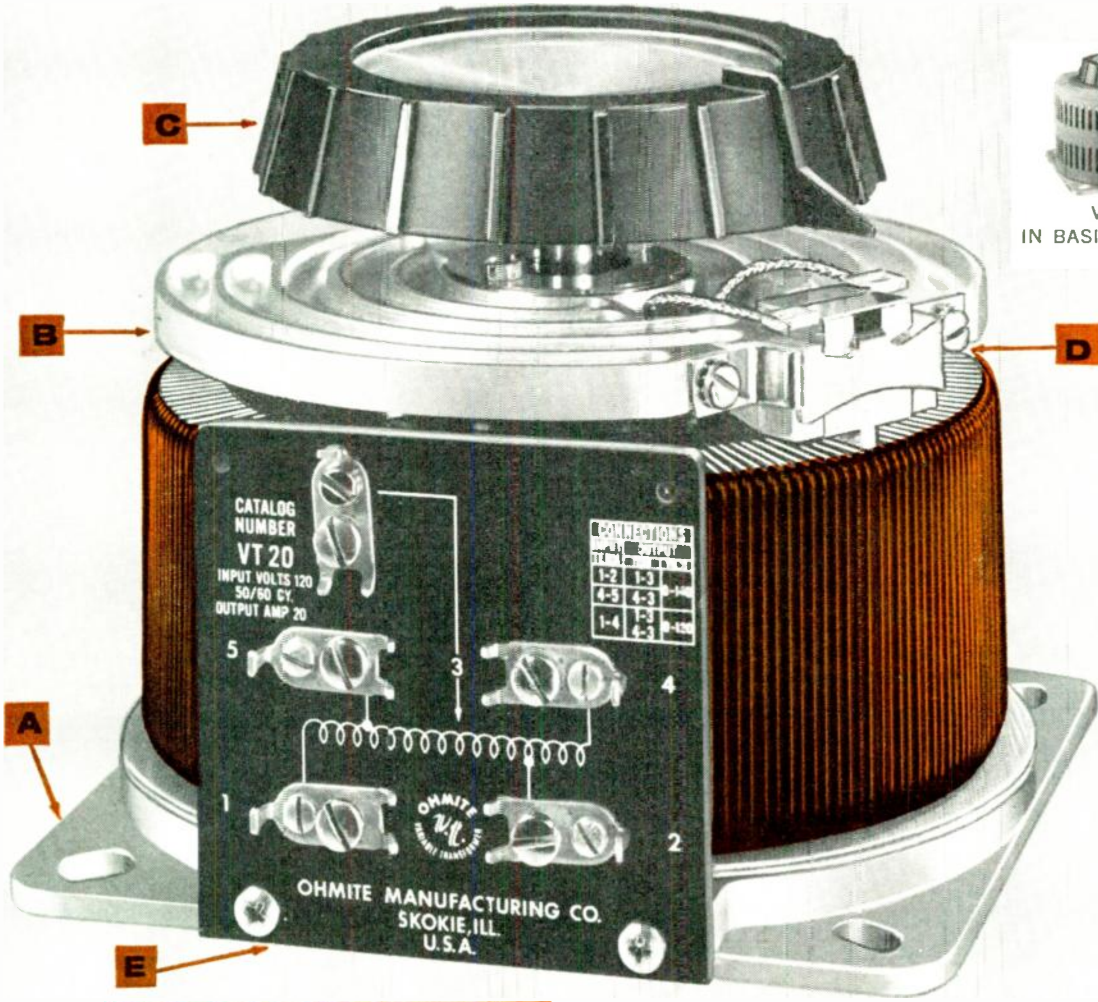

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28.4.14 Esaki Tunnel Diodes

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28.5.1 Non-Linear Dielectric Definitions

H. Diamond, *Chairman*

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E. A. Sack
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E. E. Loebner
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28.5.2 Dielectric Electro-Optic Definitions

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28.5.3 Definitions, Formation and Utilization of Electrostatic Images

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28.6 Magnetic Devices

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28.6.1 Basic Terminology, Materials and Measurements

H. A. Perkins, Jr., *Chairman*

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28.6.2 Magnetic Materials and Measurements

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28.6.3 Magnetic Memory and Logic Devices

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28.6.4 Microwaves

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28.7 Masers and Variable Reactance Transducers

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28.8 Thermoelectric Devices

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28.9 Microsystems Electronics

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21.5 New Proposals and Special Assignments

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21.7 Letter Symbols

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23.3 Video Systems and Components Methods of Measurement

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23.4 Video Signal Transmission Methods of Measurement

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F. Davidoff (alternate)
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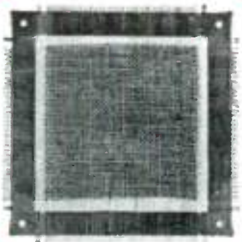

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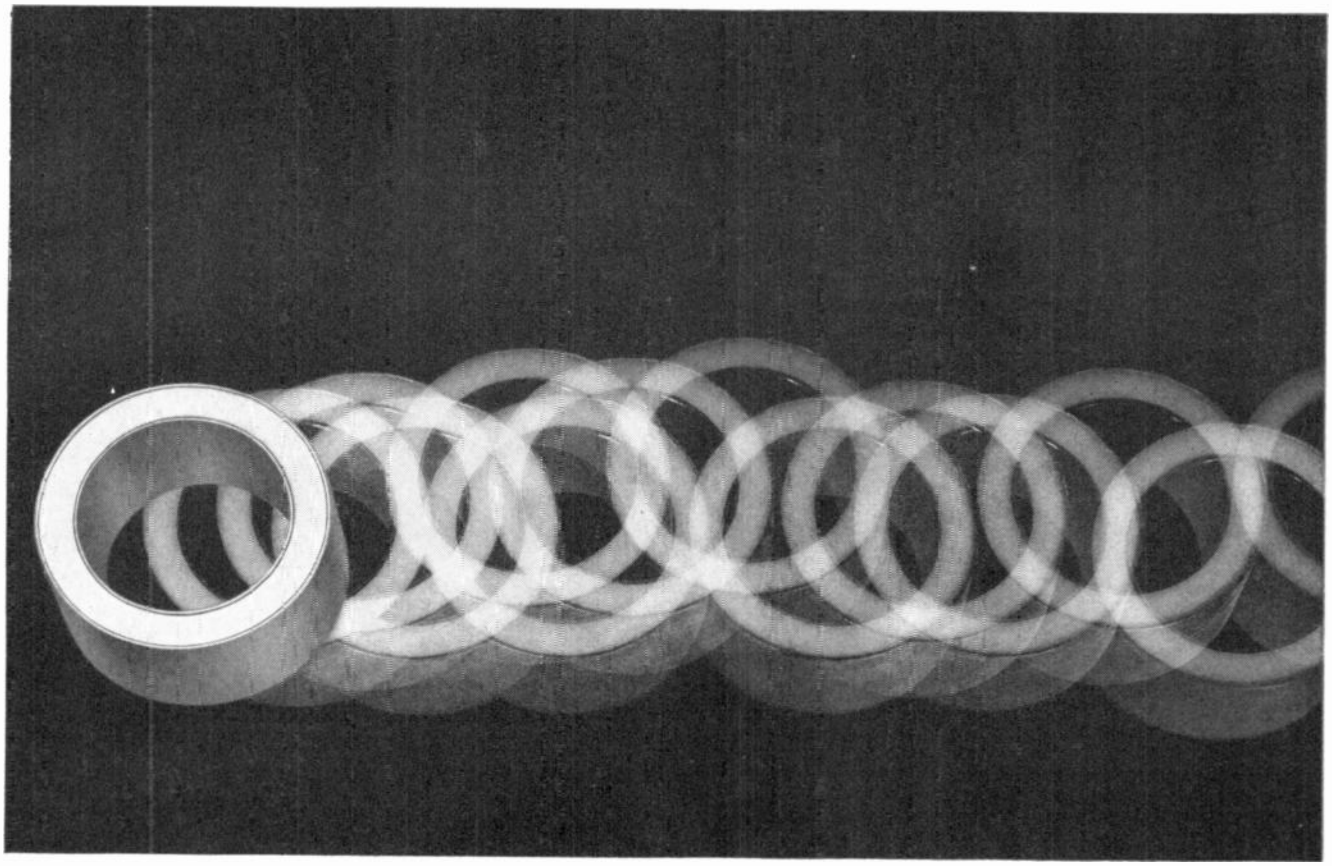

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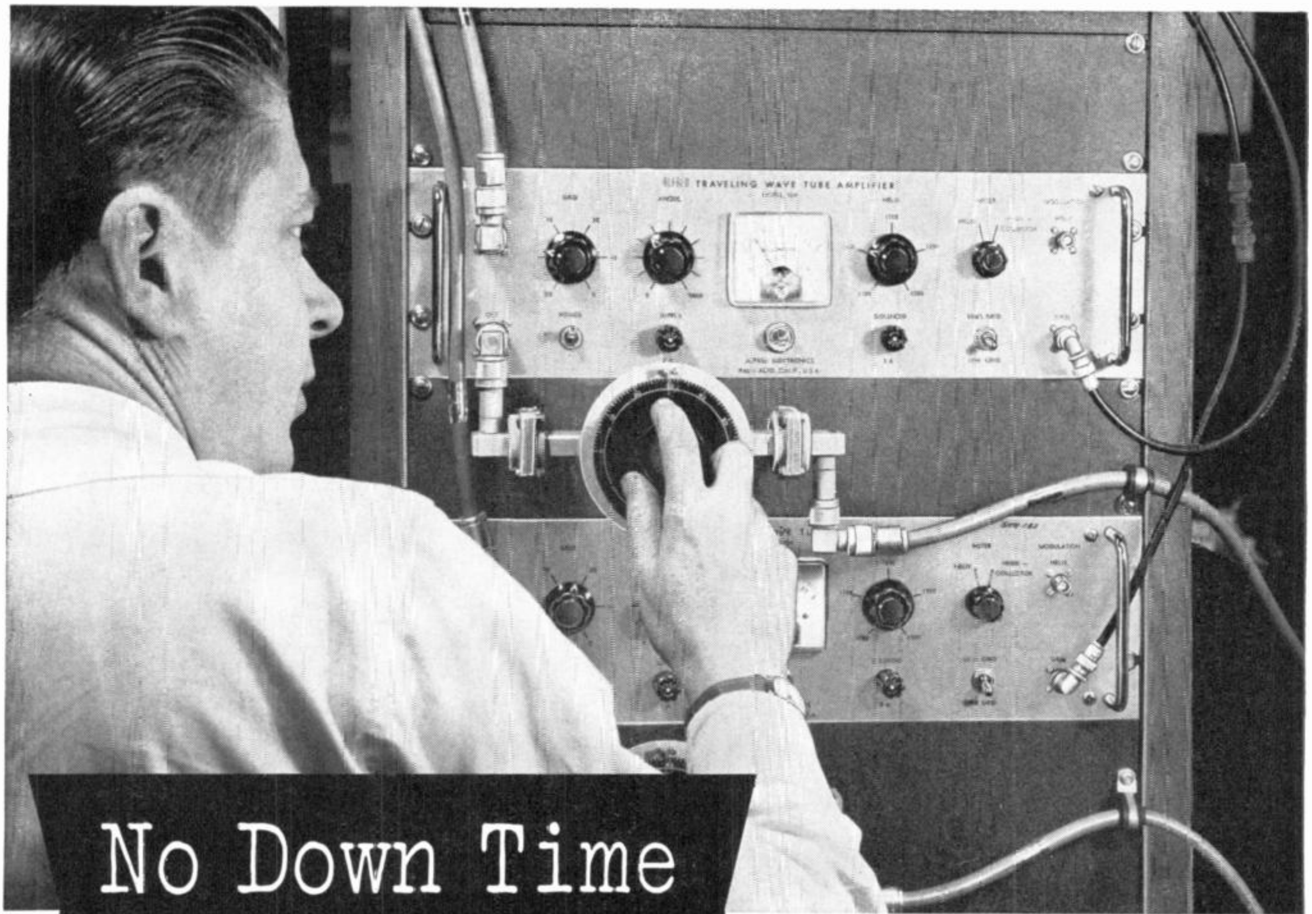

No Down Time Is Normal...

Alfred Electronics Model 504 Microwave Amplifiers, in use at Hughes Aircraft Co., Culver City, California

with ALFRED Microwave Amplifiers

These Alfred TWT microwave amplifiers have seen continuous service at Hughes for over 9 months. There has been practically no down time even for replacement of TWT tubes. Used in the RF portion of a missile testing system, the Alfred units provide high gain, wide band, flat response and low spurious modulation from 8 to 12.4 kmc. Hughes engineers praise the functional layout of the Model 504, its simple operation and reliable performance.

In short, Hughes finds the Alfred 504 Microwave Amplifiers good, sound, straightforward reliable instruments. We think you will too.

KEY SPECIFICATIONS—ALFRED MICROWAVE AMPLIFIERS

	Model	Frequency Range KMc/s	Gain db (min.)	Power Output (min.)	Price
General Purpose Amplifiers for AM, Pulse and Phase Modulation	505	1 to 2	30	10 mw	$1550.00
	501	2 to 4	30	10 mw	1490.00
	503	4 to 8	30	10 mw	1550.00
	504	8 to 12.4	30	10 mw	1490.00
	549	10.5 to 16	30	10 mw	3190.00
Medium Power Amplifiers	5-6752	1 to 2	30	1 W	$1990.00
	512	2 to 4	30 20	1 W pulsed 100 mw CW	2050.00
	502	2 to 4	30	1 W	1550.00
	5-6868	2 to 4	30	10 W	2750.00
	506	4 to 8 4.5 to 7.5	27	.5 W 1 W	2290.00
	5-542	4 to 8	30	1 W	3190.00
	*528	7 to 11	30	5 W	3490.00
	509	8.2 to 11	27	.5 W	3190.00
	510	8 to 12.4	20	100 mw	2290.00
	*527	8 to 12.4	30	2 W	3490.00
	*526	12.4 to 18	25	1 W	4950.00
High Power Amplifiers	5-6826	2 to 4	30	1 Kw Pk .01 duty cycle	$6895.00
Low and Medium Noise Figure Amplifiers	Amplifiers with low and medium noise figures are available either as packaged units or unit·zed for remote operation o' TW tube and solenoid. Standard units provide coverage from .5 to 12.4 KMc with noise figures from 7 db up.				

*New Instrument

MANY MODELS TO CHOOSE FROM

The 504 is just one model from the industry's most complete line of microwave amplifiers. For technical details and a demonstration arranged at your convenience, contact your nearby Alfred representative or write direct. Please address Dept. 86 . . .

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Industrial Engineering Notes*

Association Activities

The EIA Microwave Section has ended its long and intensive activity in the Federal Communications Commission's proceeding on allocation of frequencies of space communications with the filing of its technical findings. The filing, a companion piece to policy comments entered into FCC's Docket 13522 late in February, reiterated the Microwave Section's contention that space and earth communications systems can share the same frequencies, and gave technical explanations to back it up. The section's major conclusion was that "satellite communications systems can exist on a co-channel basis with surface multi-channel systems by engineering such that an interfering signal to the surface system is no greater than 10 db above the receiver noise level for 0.01 per cent of the time." The exhaustive study, conducted under Section Chairman R. G. Jones (Motorola, Inc.), also concluded that radar systems would interfere most with satellite communications, and could not share spectrum space. The same was found true for airborne radar. Interference by tropospheric scatter systems would be harmful, but could be engineered to protect a stationary orbiting satellite receiver. The FCC now has before it the comments of EIA and eight other firms and organizations. A lengthy study is expected before the Commission lays down rules for frequencies to be used for global communications via satellites.

Governmental and Legislative

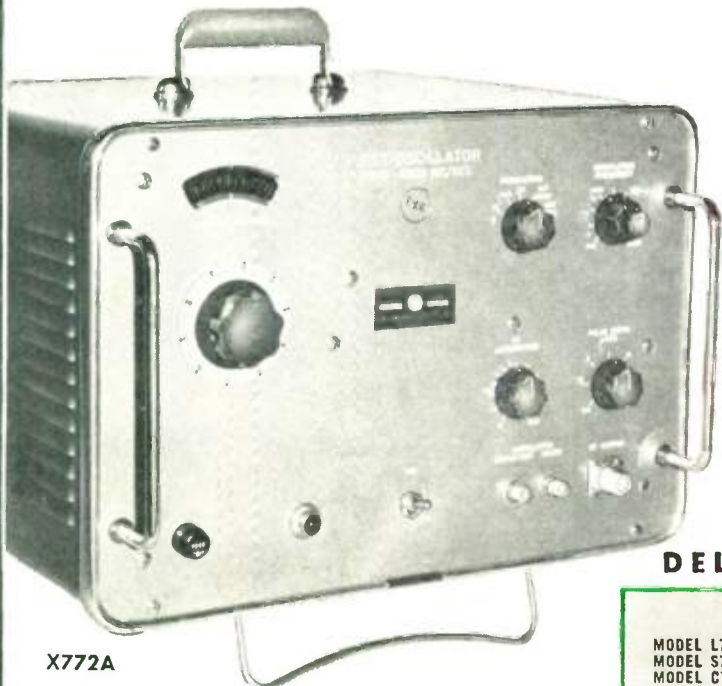
The Federal Communications Commission launched a new inquiry into the problems of establishing world-wide communications via satellites—this one concerned with regulating commercial space communications systems. The proceeding, which the FCC said would not touch on matters involved in its current space frequencies inquiry, will probe into administrative and regulatory problems which must be resolved before authorization of space communications systems operated by commercial firms. The Commission asked for industry's detailed comments based on an assumption that a single or a limited number of satellite communications systems would best serve the public. It set forth these specific questions: 1) What plan for participation is best designed to provide equitable, non-discriminatory use of satellite facilities by existing and future international communication common carriers and others? Would the plan include participation by manufacturers of satellite communication and launching equipment? 2) How would such a plan comply with existing laws and policies? Which laws and rules of communications are applicable, and what changes should be made to implement the plan? 3) To what extent would the participants in such a plan be subject to regulation by the FCC as common carriers or otherwise? The FCC notice set a May 1 deadline for industry comments. . . **The Federal Communications Commission began a formal inquiry into the methods by which one of the seven commercial VHF television channels at New York City and at Los Angeles could be made available for non-commercial educational broadcasting.** The Commission noted that, despite the achievements of educational groups in placing 54 educational TV stations on the air, "the vast populations of the New York City and Los Angeles areas—over 13 per cent of the total population of the country—are still not served by non-commercial educational TV stations. In these circumstances, and recognizing the abundance of resources for educational programming in those cities, it is urgently desirable in the public interest to inquire into available means by which it could enhance the opportunities for providing such services," the Commission's announcement said. Deadline for comments is May 1. . . . **Research and development in the field of semiconductors—transistors, crystal diodes, and related devices—amounted to more than $70 million in 1959—the latest year for which figures are available—the Business and Defense Services Administration of the U. S. Department of Commerce reported.** Private industry underwrote more than $54 million of this total, and U. S. Government agencies, primarily the Department of Defense, the remaining $16 million, BDSA said. Of 60 semiconductor manufacturers surveyed, the 3 largest accounted for approximately 39 per cent of the total factory shipments in 1959. Some 34,700 production workers were engaged in semiconductor manufacturing in the period January to April, 1960. Scientists and engineers numbered approximately 6,200, and other workers, 7,800. More than 60 per cent of the manufacturers were planning to expand their facilities. The data are contained in a publication, "Semiconductors: U. S. Production and Trade," prepared by BDSA's Electronics Division. The material is based in part on information gathered in the course of BDSA's industrial mobilization activities, and also on certain studies by the Department of Defense. Markets, prices and foreign trade in semiconductors are treated in the study. The report is available from the Superin-

(Continued on page 50A)

* The data on which these Notes are based were selected by permission from *Weekly Reports*, issues of April 3, 10, 17, and 24, 1961, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

SIGNAL SOURCES

direct reading, self contained...



X772A

4 MODELS COVER RANGES FROM 0.95 KMC TO 11.0 KMC

MODULATED AND CW OUTPUT

AMPLE POWER FOR NORMAL REQUIREMENTS

SERIES 772 SIGNAL SOURCES

- Single control tuning
- ±1% frequency accuracy
- Pulse or square wave modulation, internal or external
- 10 mw to 100 mw max. CW power output
- Regulated internal power supply

Truly unique in the industry, the FXR family of signal sources provides full coverage from 0.95 KMC through 11.0 KMC. Power output is more than ample for mcst test requirements. The sources provide for the use of internal or external modulation, either pulse or square wave, or external FM. Design features include an internal, regulated power supply and frequency tuning dial accuracy of ±1% throughout the range. This frequency tuning accuracy is always assured by automatic variation of the klystron reflector voltage simultaneous with positioning of a broadband, non-contacting tuning plunger within the oscillator cavity. Each model is a compact self-contained unit ready for laboratory or field use.

DELIVERY FROM STOCK

	FREQUENCY RANGES	PRICE
MODEL L772A MODEL S772A MODEL C772A MODEL X772A	0.95 to 2.0 KMC 1.9 to 4.0 KMC 3.95 to 8.2 KMC 7.0 to 11.0 KMC	\$1340. \$1340. \$1340. \$1340.
FREQUENCY ACCURACY	±1%, all models	
POWER OUTPUT	10 mw to 100 mw max CW output power variable by front panel control through use of an internal level-set attenuator	
MODULATED OUTPUTS	Internal: CW, pulse or square wave External: Pulse, square wave or FM	
EXTERNAL MODULATION REQUIREMENTS	Pulse: positive pulse of 30 v. amplitude across 100 K ohms. Pulse width from 1 microsecond to square wave Reflector: sine wave or sawtooth FM, sensitivity from 100 to 200 kc/v	
CONNECTORS	RF Output—Type N jack External Pulse—Type BNC jack Reflector Modulation—Type BNC jack	
POWER REQUIREMENTS	115/230 v. AC, 50 or 60 cycles, 150 w.	
DIMENSIONS	11" high x 16" wide x 15" deep	
NET WEIGHT	45 lbs.	

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Industrial Engineering Notes

(Continued from page 48A)

tendent of Documents, Government Printing Office, Washington 25, D. C. Price 15 cents.

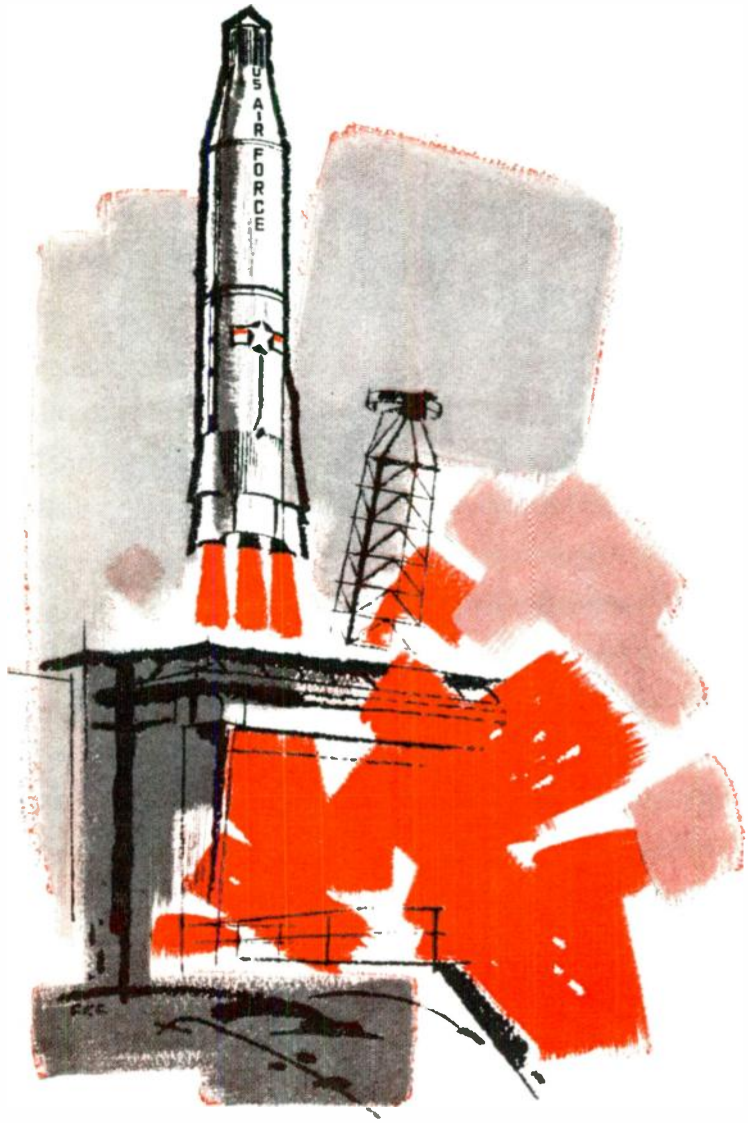
FCC Actions

Authorizing a new broadcasting service and opening a vast new market for radio receivers and broadcasting equipment, the Federal Communications Commission on April 19 adopted technical standards for stereophonic FM radio broadcasting by multiplexing techniques. The new service was made possible at this time by the intensive technical studies and tests conducted by the National Stereophonic Radio Committee established and financed by EIA. The adopted standards were described by FCC as a "composite" of stereophonic transmission standards proposed by the Zenith Radio Corporation and General Electric Company. Stereophonic broadcasting will be accomplished by subcarrier multiplex transmission in conjunction with main channel operation of FM stations offering the service. To receive the stereo broadcasts, FCC said, FM stereo receivers must be purchased or, if an FM receiver and stereo phonograph combination is already owned, a multiplex subchannel adapter must be obtained. The Commission anticipated that receiving equipment would be on the market "in the relatively near future." The Commission said the Zenith-GE combination of standards was adopted after "other systems were weighed and rejected, after technical analysis and an exhaustive field test program, because of inferior technical quality or unacceptable impairment of main channel coverage." The FCC's long-awaited stereo radio decision was the culmination of a staff study of technical data accumulated during an intensive study by the NSRC. The Committee of industry experts on the technical aspects of stereo sound was established, with the blessing of FCC, in January, 1959 at the suggestion of the late Dr. W. R. G. Baker, former President of EIA and then Director of the EIA Engineering Department.

Industry Marketing Data

Japanese transceiver equipment for the Citizen's Radio Service may become the third largest export to this country, ranking behind transistor radios and tape recorders, a Japanese trade magazine predicts. "Transceiver fever is reported to be high in the United States," states an article in the March edition of Japan Electronics. "Many makers began cashing in on this reported budding boom in America by producing transceivers for export to that country." With the release by the Federal Communications Commission of a band for private radio service, the article says, there were "increasing inquiries and, in some cases, actual orders from America on transceivers . . . and an increasing number of Japanese manufacturers started to consider or to produce transistorized transceivers." The transceiver "is likely to become one of (the) star export productions, following TR radios and tape recorders," the magazine forecast. The popularity of the Citizens Radio Service in the U. S. prompted the Japanese Radio Wave Control Commission, in conjunction with EIA of Japan, to study the feasibility of also releasing the 27 Mc band for private use, the article reported. . . . **Japanese exports of electronic products to the United States during 1960 totaled $94 million, a 24 per cent increase over the $75.6 million total of 1959, the Electronics Division, Business and Defense Services Administration, U. S. Department of Commerce, reported.** The value of exports of radio receivers in 1960, which accounted for 74 per cent of the total shipments, registered a gain of 11 per cent over 1959. Exports of radios with 3 or more transistors last year increased by 4 per cent in quantity, but declined by 4 per cent in value from the preceding year; exports of other radios increased appreciably. Other products showing substantial gains were sound recorders and reproducers, radio-phonographs, speakers, receiving tubes, and other electronic components. The 1960 exports of television receivers to the U. S. totaled 10,000 valued at $507 thousand. Exports to the U. S. were equivalent to 48 per cent of total Japanese exports of electronic products to the world in 1960 compared with 56 per cent in 1959. . . . **Shipments of electronic components increased about 4 per cent during the fourth**

(Continued on page 54A)



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MEET THE MOST SPECIALIZED AND STRINGENT DEMANDS•Cannon MS Plugs are built for rugged service! From general duty ground use to specialized missile applications, these plugs fulfill the requirements of MIL-C-5015... are also suitable for many commercial and industrial applications where quality and dependability are required. Our full line of environmental resisting MS plugs gives you the optimum in interchangeability, variety of contact arrangements, and shell types and sizes. The MS series, MS-A, MS-B, MS-C, MS-E, MS-R, MS-K, are available from authorized Cannon Distributors everywhere; or write:

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MICROWAVE DEVICE NEWS from SYLVANIA

AVAILABLE IN QUANTITY!

BROADBAND FERRITE ISOLATORS

COAXIAL DEVICES / WAVEGUIDE DEVICES

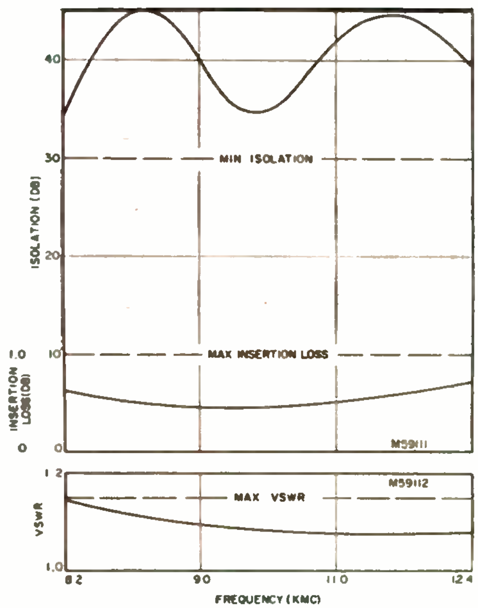
—from 1-11 GC / *—from 3.95-26 GC*

Sylvania offers a comprehensive line of coaxial and waveguide isolators—12 types, in all—featuring extraordinary isolation to insertion loss ratios as high as 30 to 1. Designed for wideband applications in such end-products as ECM and laboratory test equipment, they're extremely well suited for the reduction of VSWR and the elimination of anomalies caused by long line effects in oscillator outputs.

Your Sylvania Sales Engineer will be pleased to tell you more about the full line of *competitively priced* broadband coaxial and waveguide isolators. Ask him. For technical data, write Electronic Tubes Division, Sylvania Electric Products Inc., 1100 Main St., Buffalo 9, N. Y.

BROADBAND COAXIAL ISOLATORS Frequency	Types	Isolation (Min.) (db)	Insertion Loss (Max.) (db)	VSWR
1.0-2.0	FD-1537	10	1.2	1.2
2.0-4.0	FD-151P	15	1.0	1.2
4.0-8.0	FD-1519	10	1.0	1.25
8.0-11.0	FD-1522W	30	1.0	1.4
BROADBAND WAVEGUIDE ISOLATORS				
3.95-5.85	FD-492	18	1.0	1.15
5.85-8.2	FD-502	20	1.0	1.15
7.05-10.0	FD-512	24	1.0	1.2
8.2-12.4	FD-522	30	1.0	1.15
10.0-15.0	FD-7530	30	1.0	1.15
12.4-18.0	FD-911	24	1.0	1.15
18.0-26.0	FD-531A	24	1.0	1.15
18.0-26.0	FD-531AF1	24	1.0	1.15

Typical performance characteristics of FD-522



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GENERAL TELEPHONE & ELECTRONICS

GENERAL SYSTEM

Industrial Engineering Notes

(Continued from page 50A)

quarter of 1960 in sharp contrast to the 5 per cent decline during the third quarter, according to a report released by the Electronics Division of the Business and Defense Services Administration. Total shipments for the year were 10 per cent over 1959 levels, reflecting the general increase in electronics activity. The quarterly rise was due to increased requirements for military and industrial type electronic components. Shipments of consumer type components declined significantly. Output of receiving tubes and television picture tubes, which are predominately used in consumer electronic equipment, declined 12 per cent during the quarter while shipments of all other major groups of electronic components increased. The value of shipments of semiconductor devices increased sharply despite the continuing decline in average unit prices. Unit output of transistors increased over 30 per cent although average unit prices declined over 11 per cent. Unfilled orders at the end of 1960 were about 5 per cent above those of a year earlier and amounted to about 11 weeks' production at the fourth quarter, 1960 rate. The tables appearing in the Supplemental Information section of this edition give details by component category. They were derived from the quarterly Survey of Production Capabilities for Electronic Parts conducted jointly by the Electronics Production Resources Agency of the Department of Defense, and the Electronics Division, BDSA. The data presented, however, represent estimated total industry shipments and unfilled orders rather than total shipments and unfilled orders reported in the survey, since adjustments were necessary where the coverage was not complete. . . . **Factory sales of transistors in February gained by more than 1 million units and $2.7 million over totals for January, according to month-end statistics released by the EIA Marketing Data Department.** The number of transistors sold at the factory in February totaled 13,270,428 valued at $25,699,625. The month before, 12,183,931 units worth $22,955,167 were sold. During the first two months of this year, 25,454,359 transistors were sold at the factory, compared with 19,134,292 sold during the same period in 1960. Revenue from sales during January and February of this year totaled $48,654,792, against a total of $49,546,150 during the two-month period last year.

Military and Space

Electronic checkout techniques similar to those used in missile launchings are being tested on Army aircraft to determine if the airplanes are safe for flight, the Department of Defense announced. Army-sponsored research into the feasibility of the concept is being carried out by the Bendix Corporation, York, Pa. Known as Project ALARM, the concept uses strategically placed sensors to forecast the condition of critical components.

The Highest Sensitivity

and the lowest noise.....

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

fig. 1 Gain characteristics

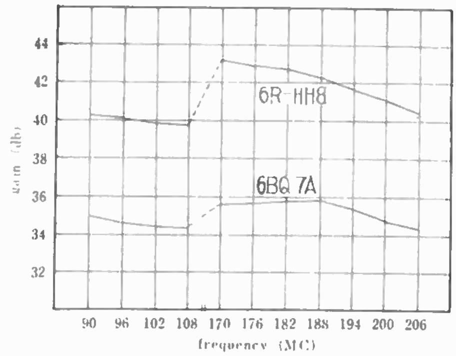
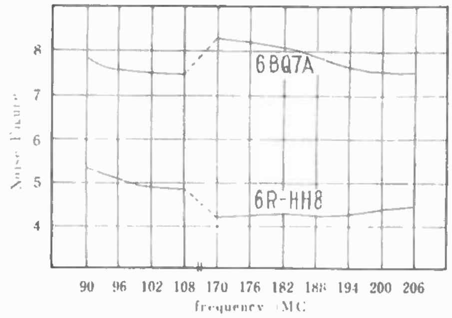
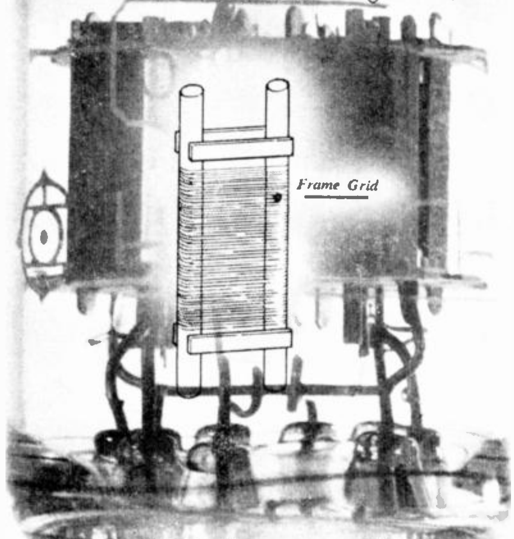


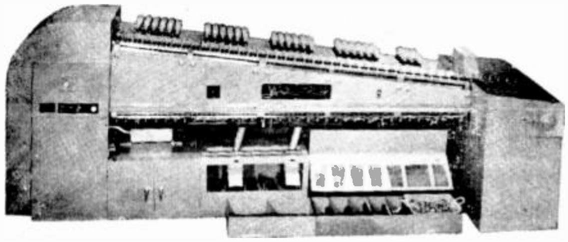
fig. 2 Noise characteristics





Frame Grid (6R-HH8)

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



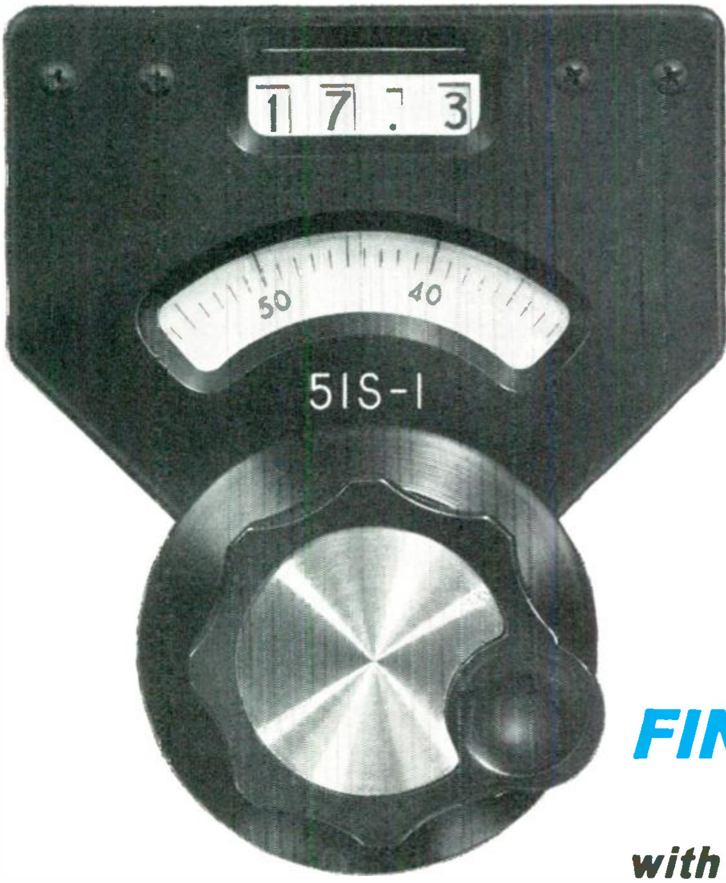
Automatic tube testing equipment

Hitachi, Ltd.

Tokyo Japan

Cable Address: "HITACHY" TOKYO

NEW FROM COLLINS — THE 51S-1



FINGERTIP TUNING

with accuracy. The latest in a series of general coverage HF receivers features **single sideband** and AM reception with: **extreme dial accuracy,** visual setting within one kc throughout the range – **high frequency stability,** particularly suited to receiving pre-assigned frequencies – **optimum selectivity,** made possible by **Collins Mechanical Filters. Highest sensitivity** for difficult monitoring assignments – all in **one compact, lightweight, easily installed unit.**

Write for descriptive brochure.

COLLINS

COLLINS RADIO COMPANY • CEDAR RAPIDS, IOWA • DALLAS, TEXAS • BURBANK, CALIFORNIA

IRE People

Appointment of **Howard A. Bond** (SM'57) as Vice President of Systems and Development in Dresser Electronics, SIE Division, has been announced in Houston, Tex. He will be responsible for the Division's expanding military electronics product development activities as well as for its industrial and military electronic systems management and engineering.

The SIE Division of Dresser Electronics now provides electronic equipment and systems for military applications, pipeline and process control systems, and geophysical instrumentation equipment.

Mr. Bond has been serving on one of the panels of the Scientific Advisory Board of the National Security Agency and as an advisor to the Deputy Director for Research and Development. He was formerly Manager of Reconnaissance Systems at Stromberg-Carlson with responsibility for management of a $50,000,000 electronic reconnaissance system under development for the United States Air Force since 1957.

For more than ten years previously, he was engaged in research and development for the U. S. Government, first with the Naval Security Agency, and later as chief of a radio frequency division of the National Security Agency.

From 1944 to 1947, he served in the U. S. Army Signal Corps and the U. S. Army Security Agency, and now holds the rank of major in the U. S. Army Reserve, as a member of the U. S. Army Security Agency.

Mr. Bond holds a degree in electrical engineering from Northwestern University, Evanston, Ill., and attended the graduate school at the University of Maryland, College Park. He is a member of Tau Beta Pi, honorary engineering fraternity, and is active in the IRE Professional Groups on Engineering Management, Military Electronics, Antennas and Propagation, and Communication Systems.

❖

The appointment of **Dr. Donald B. Brick** (S'50–A'54–SM'56) as Manager of the newly-formed information processing group at the Applied Research Laboratory of Sylvania Electric Products, Inc., has been announced.

D. B. Brick

The Applied Research Laboratory is responsible for the investigation and creation of new electronic systems and techniques to meet the requirements of government and industry.

Dr. Brick's group will be responsible for research in bionics, speech and pattern recognition, statistical decision theory, artificial intelligence information retrieval and related areas in the data and information processing fields.

With Sylvania since 1955, he was formerly a Research Fellow at Harvard University, Cambridge, Mass., where he worked on electromagnetic scattering, antennas, and microwave circuits. During the past five years, he has directed a variety of work in electromagnetic theory, applied physics, countermeasures and counter-countermeasures, missile guidance and radar systems. Among his more recent projects has been the study of the use of artificially generated clouds for communication purposes.

He is a graduate of Harvard University, and received the Bachelor of Arts degree, cum laude, in engineering sciences and applied physics in 1950, the Master's degree in 1951, and the Ph.D. degree in applied physics (electrodynamics) in 1954.

He is a member of Sigma Xi, honorary scientific fraternity; the Harvard Engineering Society, the American Physical Society and the American Association for the Advancement of Science.

❖

Dr. Liborio J. Castriota (S'47–M'52–SM'59) has been promoted to the position of Manager of Engineering at PRD Electronics, Inc., Brooklyn, N. Y., it was recently announced.

L. J. Castriota

In 1947, he received the B.E.E. degree from New York University, College of Engineering, New York, N. Y., and the M.S. degree from Harvard University, Cambridge, Mass., in 1948. During the same year, he joined the research staff of the Microwave Research Institute of the Polytechnic Institute of Brooklyn, Brooklyn, N. Y., where he worked on microwave interference studies and RF noise generation investigations.

Beginning in 1950, he headed an interference research unit at the Rome Air Development Center, Rome, N. Y., and was engaged primarily with radar and communications interference problems.

Two years later, he returned to the Microwave Research Institute, where he undertook research activities on low frequency telephony location circuitry, continuous lookthrough radar jamming systems development and receiver noise studies.

During the period from 1955–1957, his services were again utilized at the Rome Air Development Center, this time as a USAF research consultant. In 1957, he was appointed leader of the Linear Networks Section and was engaged in microwave dis-

(Continued on page 60A)

IRE People

(Continued from page 58A)

tributed parameter networks studies. As a member of the Research Faculty, he also lectured in the Graduate School, Polytechnic Institute of Brooklyn.

The following year (1958), he was awarded the Doctorate from the Polytechnic Institute of Brooklyn. In 1959, he joined PRD, where he became involved in all phases of operations with the company especially with the development of military and industrial microwave calibration systems and special test equipment for field and laboratory applications.

Dr. Castriota is a member of Sigma Xi and of RESA.

❖

C. E. Rutherford, President of Rutherford Electronics Company, Culver City, Calif., has announced the appointment of **John J. Davis** (S'59–M'60) as project engineer on the new B-10 Transistorized Pulse Generator.


J. J. Davis

Prior to coming to Rutherford, Mr. Davis was at Summers Gyroscope, Aeronutronic Division of the Ford Motor Company, and the Ramo-Wooldridge Corporation.

He is a registered engineer in California, a member of Eta Kappa Nu, the Audio Engineering Society, and is also currently a broadcasting consultant and chief engineer at KCBH, Beverly Hills, Calif. He attended the University of Southern California, Los Angeles, earning the B.S. and M.S. degrees in electrical engineering.

❖

Alton C. Dickieson (SM'44–F'60) has been named Director of Transmission Development of the Bell Telephone Laboratories, effective April 1, 1961.

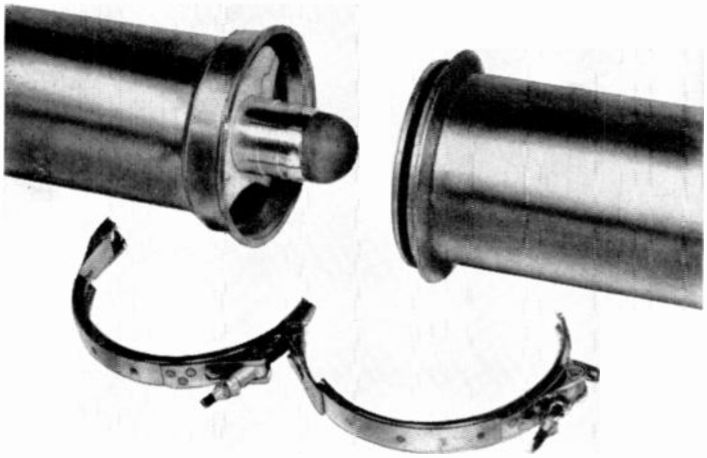

A. C. Dickieson

He was born in New York, N. Y., and studied electrical engineering at the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. He began his Bell System career in 1923 in the engineering department of Western Electric Company, and transferred to Bell Laboratories when it was established in 1925. For most of his career he specialized in long distance transmission systems. During World War

(Continued on page 64A)

proven in use

DIELECTRIC Rigid Transmission Lines Satisfy Every Performance Requirement



Choose From Two Hard-Drawn, High-Conductivity Lines . . . QUICK-CLAMP TYPE 40 and EIA (RETMA) TYPE 70

Available in:

Standard sizes: ⅞, 1⅝, 3⅛, 6⅛, 9 3/16 inches

Special sizes: 9, 10, 12, 16 inches

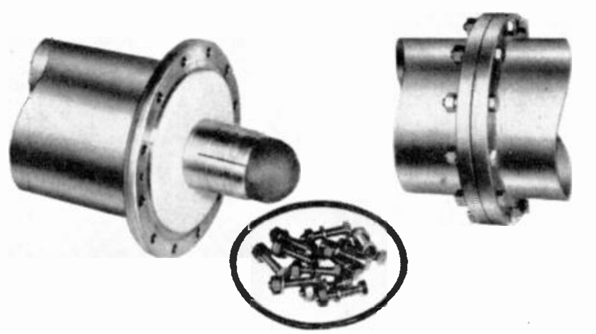
Any length up to 30 feet

All standard impedances from 25 to 200 ohms

Copper or aluminum outer conductors

Both DIELECTRIC Lines—Type 40 Quick-Clamp and Type 70 EIA Bolted-Flange are:

- **Engineered to exhibit low VSWR, small attenuation of power**
- **Designed for high average and high peak power transmission**
- **Engineered to provide minimum r-f leakage at connections**
- **Notable for their proven low-noise connections**



TYPE 40 QUICK-CLAMP LINE

Unusually flexible in application and installation, DIELECTRIC'S Type 40 Line offers such specific advantages as:

- flange connections that swivel to any rotational position
- a Marman Clamp that assures quick and positive assembly with the tightening of only two bolts
- piloted flanges that eliminate the need to align the outer conductor by centering rings or other devices

A captive O-ring for sealing not only provides a perfect gas seal, but renders damage or jamming during assembly virtually impossible. Electrical contact is dependent only on tightening the clamp.

TYPE 70 EIA (RETMA) LINE

When specifications stipulate an EIA (RETMA) bolted-flange line, DIELECTRIC Type 70 offers the same electrical properties as the Type 40 line. Positioning of line sections during installation is made accurate by alignment pins. An O-ring placed between the flanges is provided for gas sealing.

PRESSURIZED COMPONENTS ACCESSORY TO TYPE 40 AND TYPE 70 LINES

90° ELBOWS • 45° ELBOWS • FLEXIBLE SECTIONS
BREAKAWAY SECTIONS • ADAPTERS
QUICK STEP REDUCERS • GAS STOPS
END SEALS • END COVERS
COUPLINGS (unpressurized) • FIELD FLANGES

Prompt shipment of standard lines and components is usual because a large stock of manufactured parts is maintained for fast assembly. And in addition, DIELECTRIC'S complete r-f laboratory and production facilities are available for the design and manufacture of other lines and components for special requirements.

For more detailed data on the above lines and their components, write for Catalog 61-4. Also, DIELECTRIC designs, develops and manufactures a wide range of components and equipment for the communications industries. If you'd like to know how we can help you solve a problem in this field, simply dial DIELECTRIC.

*For applications in the broadcast and television field, transmission line products are available from the Radio Corporation of America. For all other applications, contact DIELECTRIC directly.

DIELECTRIC'S areas of capability include coaxial, waveguide and open wire techniques . . .
TRANSMISSION LINES & COMPONENTS • NETWORKS
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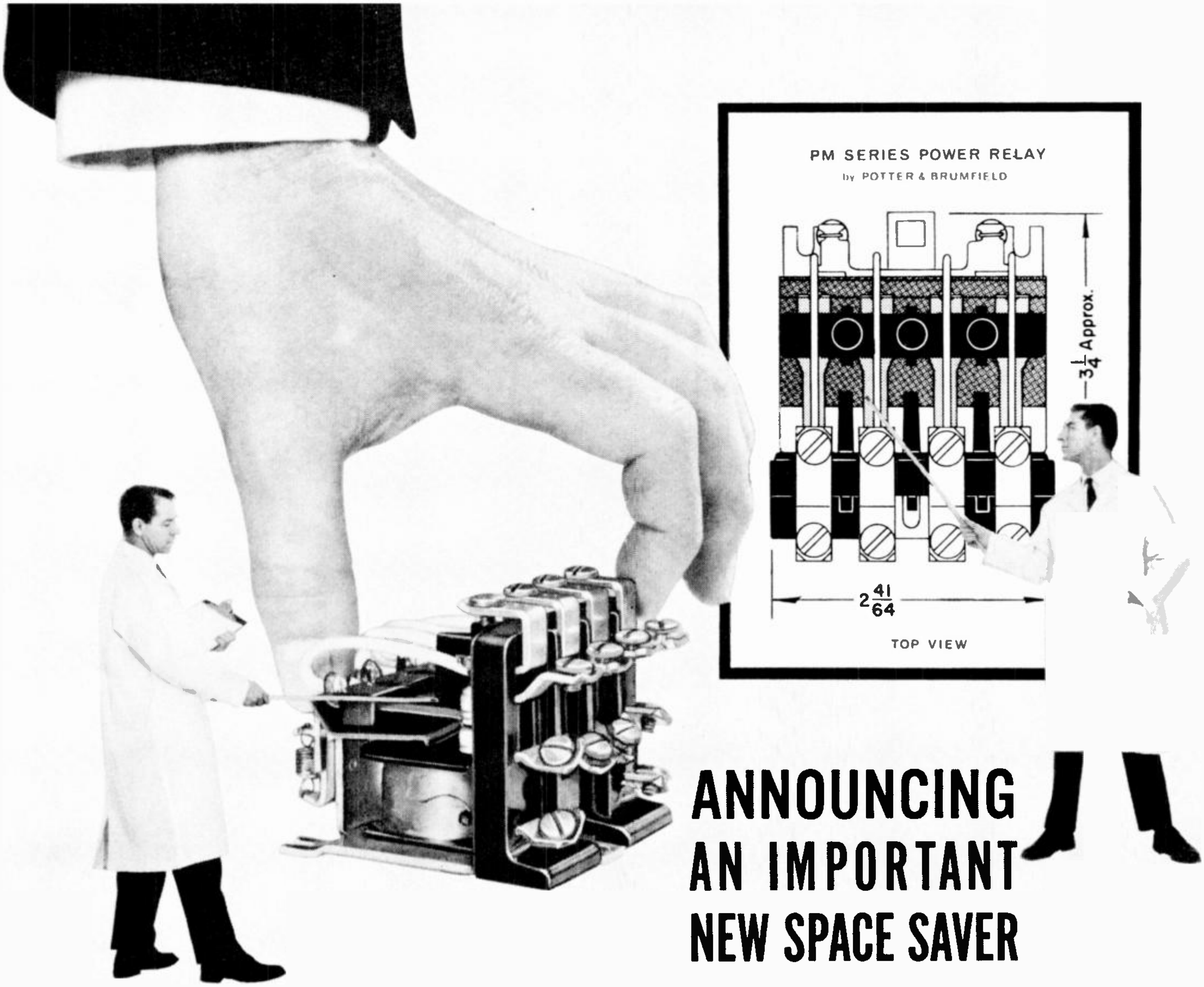
dial DIELECTRIC for solutions to communications problems

DIELECTRIC PRODUCTS ENGINEERING CO., INC.

RAYMOND, MAINE **Tel. No. OL 5-4555**



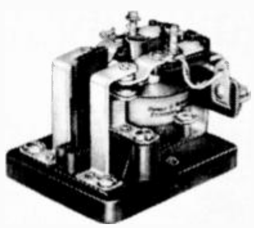
ANNOUNCING AN IMPORTANT NEW SPACE SAVER

P&B compact 4PDT power relay switches one H.P. per moveable arm

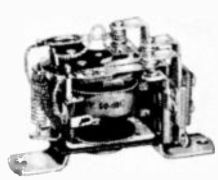
Save panel space! This new 4-pole relay is only 3/16" wider than our PR Series, America's most popular 2-pole power relay! Yet, it is engineered for reliable heavy-duty switching . . . and you can confidently expect 10 million mechanical operations.

PM Series relays are rated at 16 amperes (or 1 H.P.) at 115 volts, 50/60 cycles resistive . . . and special relays can be supplied for loads up to 25 amperes, at 220 volts, 50/60 cycles resistive. Heavy screw terminals are arranged for fast, easy hook up. An adapter plate is available for mounting PM relays in the same location used for 2-pole relays.

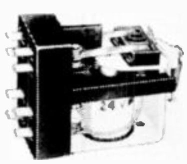
For full information, write today or call your nearest P&B representative.



PR Series



MR Series



AB Series

A whole family of power relays for a wide range of applications carry the P&B symbol of quality. Call P&B first for all your power relay requirements.

PM ENGINEERING DATA

GENERAL:

Description: Heavy-duty AC power relay.
Insulating Material: Molded phenolic.
Insulation Resistance: 100 megohms minimum.
Mechanical Life: 10 million operations minimum.
Contact Life: 100,000 operations minimum at rated load.
Breakdown Voltage: 2,000 volts rms minimum between all elements and ground.
Ambient Temperature: −55°C to +55°C.
Weight: Approximately 14 ozs.
Pull-In: 78% of nominal voltage.
Terminals: Heavy-duty screw type with No. 8-32 BH screw.

CONTACTS:

Arrangements: 4PDT or 4PST—normally open.
Material: 1/4" dia. silver-cadmium-oxide.
Rating: 16 amps @ 115 volts, 50/60 cps resistive.
8 amps @ 220 volts, 50/60 cps resistive.
1 H.P. per moveable, 115 or 220 volts AC single phase.
25 amps @ 220 volts, 50/60 cps resistive available on special order.

COILS:

Voltage: 6 to 230 volts AC 50/60 cycles.
Power: 14 volt-amps average at nominal voltage.
Duty: Continuous.

P&B STANDARD RELAYS ARE AVAILABLE AT YOUR LOCAL ELECTRONICS PARTS DISTRIBUTORS

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IRE People

(Continued from page 60A)

II he was in charge of anti-submarine warfare activities and of carrier systems for the Signal Corps. In 1954 he was responsible for planning the communications systems for the original Distant Early Warning (DEW) Line stations in Alaska. He was named Director of Transmission Systems Development in 1951.

Mr. Dickieson is a Fellow of the AIEE.

❖

Charles H. Doersam (SM'51) has been appointed Product Development Manager in Engineering at Potter Instrument Company, Inc., Plainview, N. Y., according to a recent announcement.

C. H. Doersam

He joined the company in October, 1960, as Assistant to the Chief Engineer.

Prior to this association he was employed by Sperry Gyroscope Company, and earlier by Fairchild Camera and Instrument Company.

Mr. Doersam received the Bachelor of Science degree in 1942 and the Master's degree in mechanical engineering in 1944, both from Columbia University, New York, N. Y.

❖

Promotion of Vice President **Norman H. Enenstein** (S'46–A'49–M'55) to Director of Technical Administration of Litton Industries, Inc., was recently announced. He had been Director of the Data Systems Laboratory at Canoga Park, Calif.

Before joining Litton in 1958, he was Vice President and Director of Engineering for a Los Angeles electronics firm. He is a graduate of UCLA with the Bachelor's degree in meteorology and holds the M.S. and Ph.D. degrees in electrical engineering from the California Institute of Technology, Pasadena.

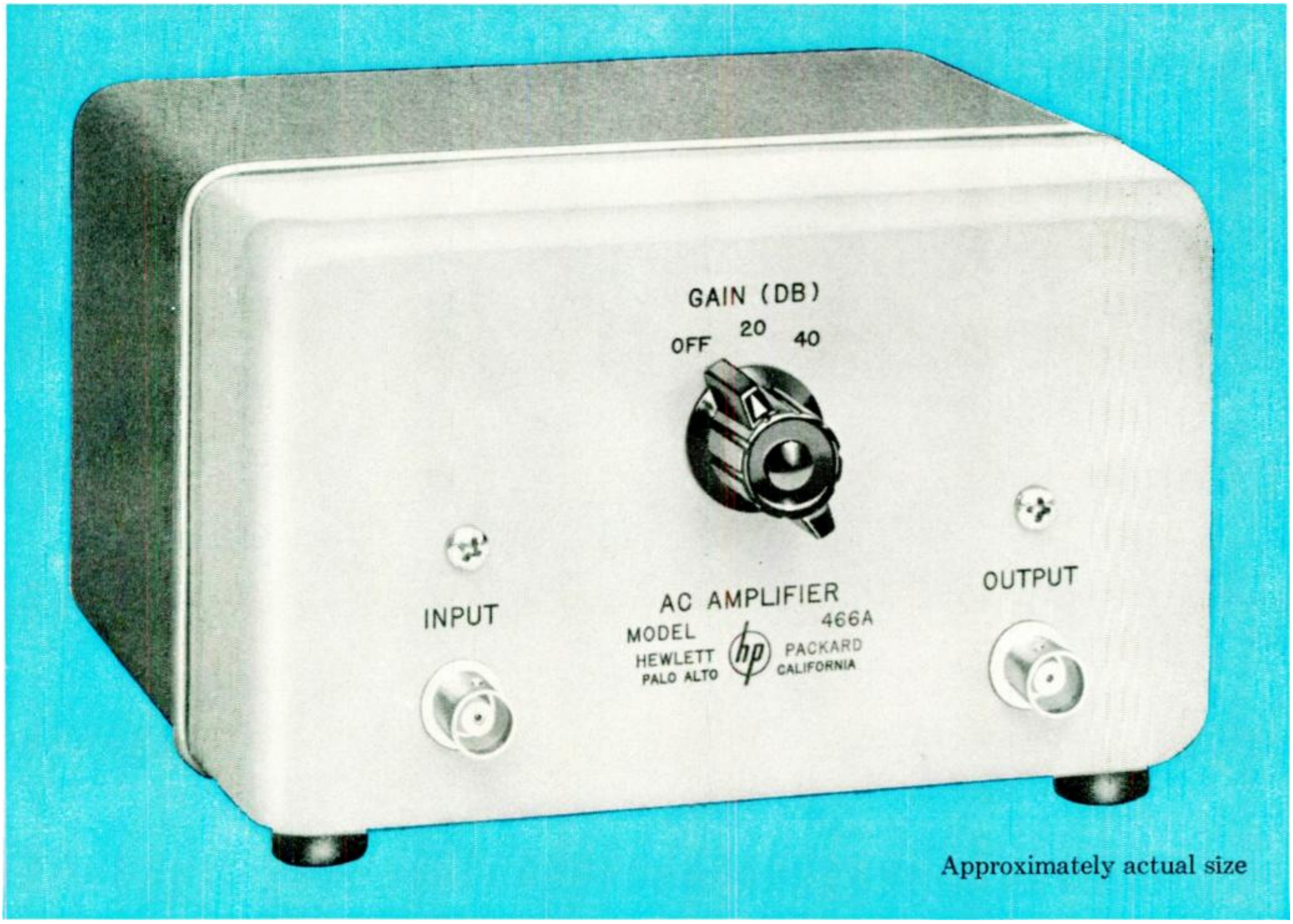
❖

George A. Franco (S'54–M'56) has been named to receive the annual General Dynamics/Electronics Award for Science and Technology for 1960—a commendation plaque and $2,000.

G. A. Franco

He is manager of the Radio Communication Laboratory in General Dynamics/Electronics' Research Division. The award was given in recognition of his work in developing a new technique for high speed data transmission that is exceptionally resistant to jamming. This technique, called DEFT

(Continued on page 69A)



Approximately actual size

This 3 lbs. of transistorized new AC amplifier gives you 20 or 40 db gain, increases scope or VTVM sensitivity 10 or 100!

This new hp 466A AC Amplifier is just 4" high, 6" wide and 6" deep. Yet it can become one of the most helpful instruments on your bench, or in the field. It is ac or battery powered; battery operation gives you hum-free performance and easy portability. Response is flat within approximately ½ db over the broad range of 10 cps to 1 MC, distortion is less than 1%, and gain is stabilized by substantial negative feedback to virtually eliminate effects of transistor characteristics and environment.

For a demonstration on your laboratory or field application, call your hp representative or write direct.

Specifications

Gain:	20 and 40 db, ±0.2 db at 1000 cps.	**Distortion:**	Less than 1%, 10 to 100,000 cps.
Frequency Response:	±0.5 db, 10 cps to 1 MC; ±3 db, 5 cps to 2 MC.	**Power:**	Ac line power normally supplied, but battery operation available. (12 radio type mercury cells, battery life about 160 hours.) Specify battery operation if desired.
Output Voltage:	1.5 v rms across 1500 ohms.		
Noise:	75 μv rms referred to input, 100,000 ohm source.	**Dimensions:**	6¼" wide, 4" high, 6¼" deep. Weight: approx. 3 lbs.
Input Impedance:	1 megohm shunted by 25 μμf.		
Output Impedance:	Approximately 50 ohms.	**Price:**	$150.00 f.o.b. factory. (Either ac or battery operation.)

Data subject to change without notice.

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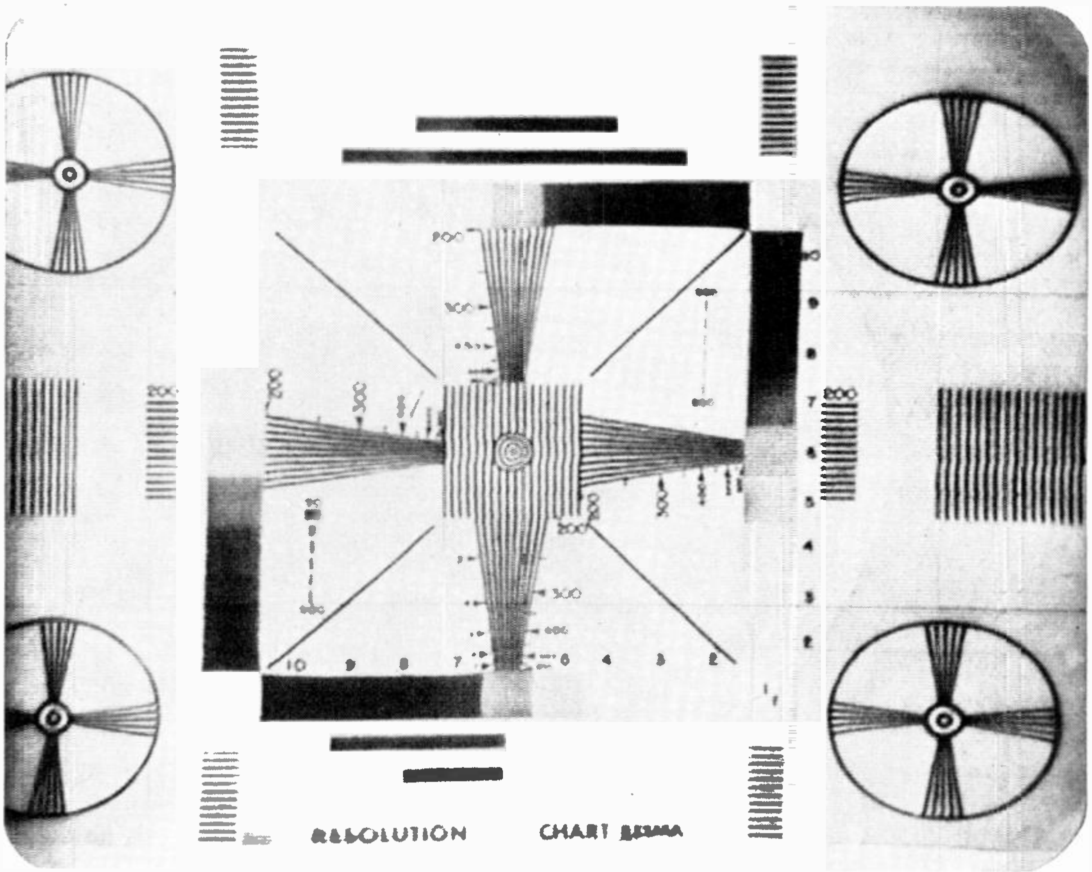
Field Representatives in all principal areas

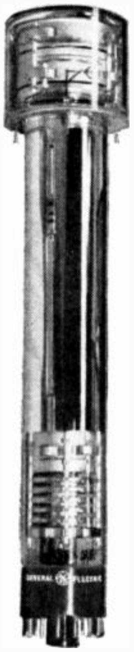
Another First From General Electric



Unretouched photo of TV monitor test pattern shows 600-line resolution accomplished at 1000 cycles, 44 G's vibration with a General Electric super-ruggedized image orthicon selected at random from stock.

Super-Ruggedized Image Orthicons Exceed

600-LINES AT 1000 CYCLES, 44G's



General Electric, which first introduced the Magnesium Oxide thin-film semi-conductor target, now offers a complete line of super-ruggedized image orthicons.

Outstanding performance of new General Electric IO's is the result of a new integral mounting technique recently developed by General Electric engineers.

Initial tests indicate this new technique surpasses all previous methods of ruggedizing image orthicons.

Built to withstand severe vibration conditions encountered in tank, aircraft, missile and space-vehicle applications, the new IO's greatly exceed military specifications of 350-line resolution at 50 to 500 cycles and 5 G's.

Representative samples are evaluated to as high as 1000 cycles, 44 G's vibration.

A complete line of the new IO's is available for immediate delivery. For more information on General Electric super-ruggedized image orthicons, and other special purpose tubes, contact the General Electric Co., Camera Tube Section, Cathode Ray Tube Dept., Building 6, Electronics Park, Syracuse, New York.

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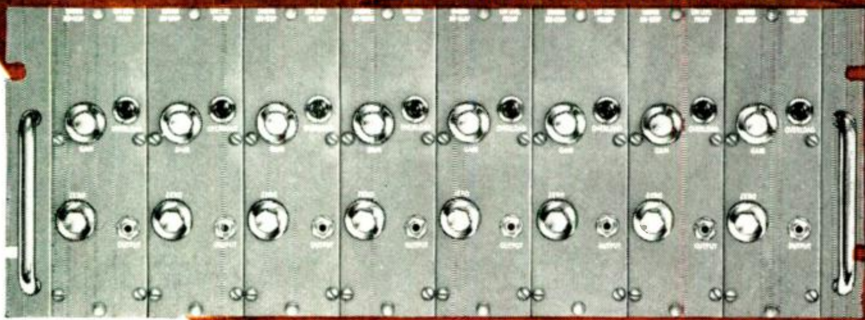
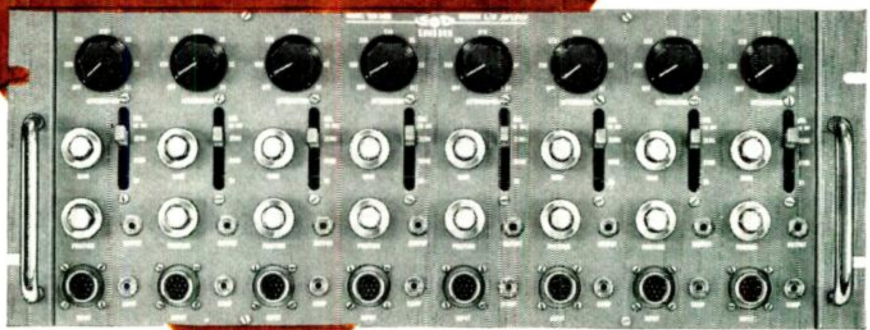
CATHODE RAY TUBE DEPARTMENT

GENERAL ELECTRIC

Model 860-1500P—handles low level DC data signals in the presence of high common mode



Model 658-3400 — drives high frequency optical galvanometers to 5 KC



COMPACT 7″ HIGH 8-CHANNEL UNITS ARE COMPLETELY TRANSISTORIZED, HAVE FLOATING INPUT ISOLATED FROM OUTPUT

Sanborn precision amplifiers

Data Preamplifier — Model 860-1500P

Designed for precise, economical amplification of signals with source impedance of zero to 10,000 ohms, such as thermocouples, strain gage bridges, etc. in presence of severe ground loop noise, and for driving digital voltmeters, scopes, tape recorders and similar devices. Each plug-in unit is only 2″ x 7″ x 14½″ deep; 64 channels with blower require only 60″ of rack-panel space. Separate 868-500 Power Supply required for every 8 preamplifiers. Power consumption 2.5 watts per channel.

Noise	3 uv peak-to-peak
Gain	100 (10 mv in gives 1 v out) (Model 860-1500PA with gain of 1000 also available)
Output	± 1 v across 300 ohms, DC-70 cps; ± 1.5 v to 40 cps. Output impedance 100 ohms. (10 v across 10K available on special order.)
Linearity	± 0.1% of full scale output (2 v)
Common Mode Performance	120 db rejection at 60 cps, 160 db at DC, with 5000 ohms unbalance in source. Inphase tolerance 220VAC.
Input Impedance	Greater than 200,000 ohms
Gain Stability	± 0.1% for 24 hours
Drift	± 2 uv referred to input
Rise Time	to 99.9% less than 25 ms

Optical Galvanometer Amplifier — Model 658-3400

Eight channels of amplification and common power supply. Each channel provides for sensitivity, compensation, damping and current limiting. Inputs floating and guarded, impedance 100,000 ohms on all ranges. All amplifier elements except output transistors are plug-in assemblies.

Sensitivity	± 10 mv input gives ± 400 ma output into 20 ohm load (max.). Eleven attenuator steps to X2000 in 1-2-5 ratio, smooth gain control.
Common Mode Performance	± 500 volts, max; rejection 140 db min at DC.
Gain Stability	Better than 1% to 50°C and for line voltage variation from 103-127 volts.
Frequency Response	0 to 5 KC within 3 db; can accomodate wide range of galvanometers.
Output	Output networks available for wide range of galvanometers.
Power Consumption	125 watts for 8 channels.

Your Sanborn Sales-Engineering Representative (offices throughout the U. S., Canada and overseas) will provide detailed information and application assistance. Call him or write plant in Waltham, Mass.

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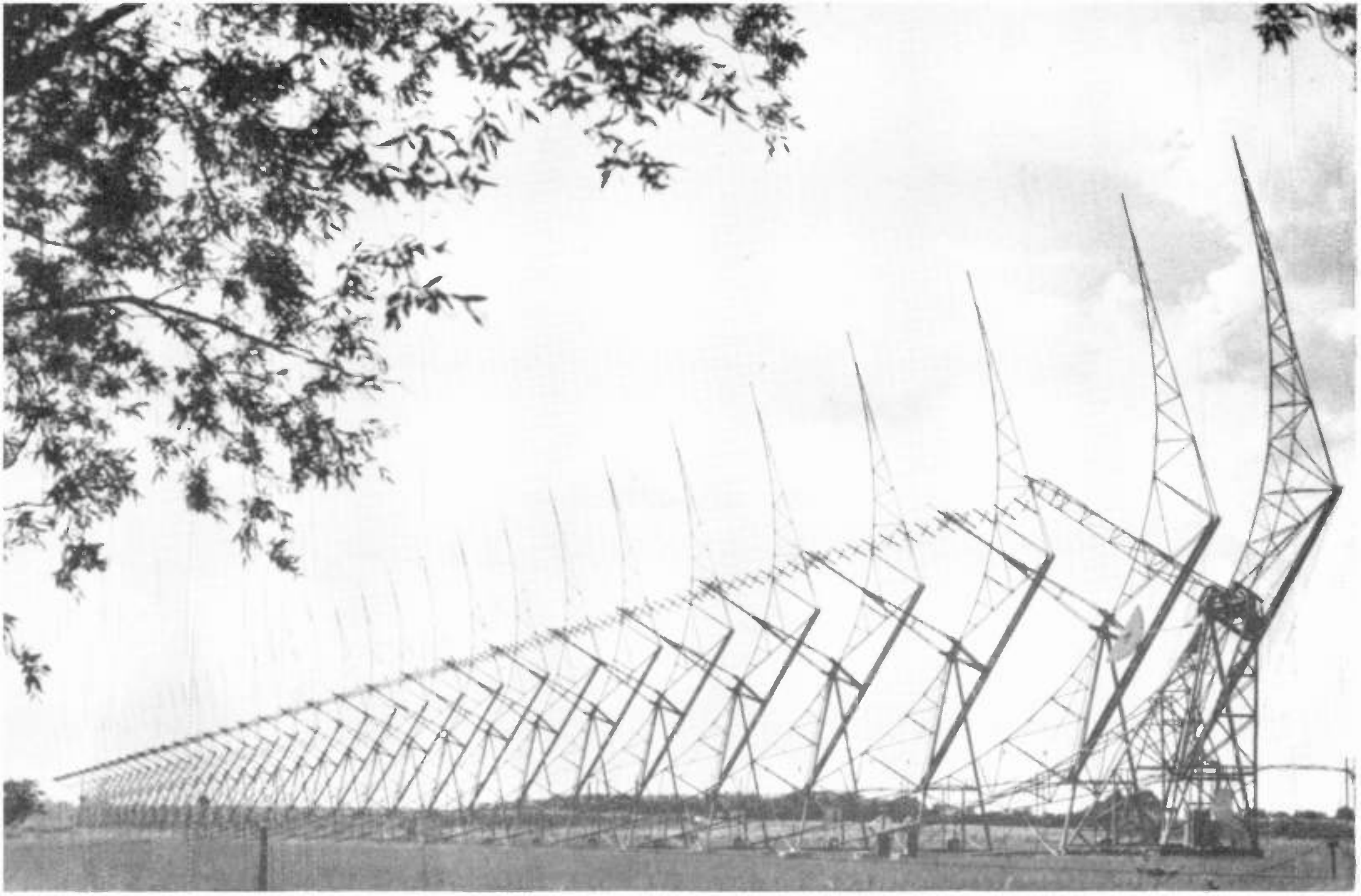
INDUSTRIAL DIVISION
175 Wyman Street, Waltham 54, Massachusetts



The giant radio-telescope aerial at Mullard Radio Astronomy Observatory, Cambridge

What's happening in Europe?

From Europe's center of electronics research and development—universities, industries, government agencies—comes a prolific stream of new ideas and practical applications.

A complete, up-to-date picture of this vigorous progress is reflected in WIRELESS WORLD, Britain's leading technical magazine for electronics, radio and television, enjoying by far the widest readership among Britain's scientists and technologists in this field.

WIRELESS WORLD articles are backed by the authority of top specialists, and maintain a careful balance between theory and practise. WIRELESS WORLD's coverage includes illustrated reviews of latest equipment, news of important conferences and exhibitions, detailed reports of industrial developments.

WIRELESS WORLD was the world's first magazine devoted to radio, and this year celebrates its fiftieth anniversary. It is an essential medium for keeping contact with European advances.

TYPICAL RECENT FEATURES:

- Report on the London Physical Society Exhibition
- Beam Indexing Color TV Tubes
- Accurate Hydrographic Radar
- Increasing Dynamic Range in Magnetic Recording
- Permeability Tuners for Television
- The "Bootstrap-Follower" Circuit
- Report on London Scientific Radio Conference (U.R.S.I.)

Wireless World

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IRE People

(Continued from page 64A)

(Dynamic Error Free Transmission), was described in a technical paper which he presented at the 1961 IRE International Convention.

This is the second of the awards, which are given annually by General Dynamics/Electronics in recognition of outstanding achievement in science and technology, and for which all employees of this division of General Dynamics Corporation are eligible.

In presenting the award to Mr. Franco, J. D. McLean, President of General Dynamics/Electronics, told him that his work "opened new horizons in the field of data transmission, and established a solid foundation for future developments. . . ."

Mr. Franco has been with General Dynamics since 1956. He is a native of Phillipsburg, N. J., and a graduate of Lehigh University, Bethlehem, Pa., with the degree in electrical engineering. During World War II he served as a radio operator and gunner in the U. S. Air Force in Europe. In 1950 he re-entered military service as a lieutenant in the U. S. Army, and served two years in Korea. He is a member of the American Radio Relay League.

❖

Airborne Instruments Laboratory (AIL), a division of Cutler-Hammer, Inc., has announced the appointment of **Winfield E. Fromm** (A'41–M'44–SM'50) to the position of Director, Research and Systems Engineering Division. Prior to this appointment, he was in charge of AIL's Space Technology and Research Department.

W. E. Fromm

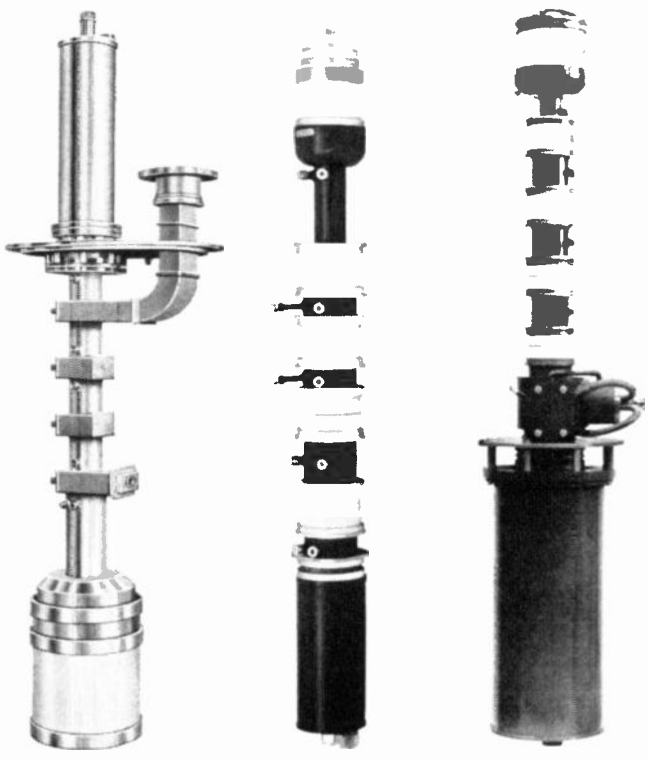
He received the B.S. degree in electrical engineering from Drexel Institute of Technology, Philadelphia, Pa., in 1940. He received the M.E.E. degree in 1948 from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y.

In December, 1941, after one and one-half years in Kansas City, Mo., with Transcontinental and Western Air, Inc., as an aircraft radio engineer, he joined Columbia University's Division of War Research (OSRD) at Quonset Point, R. I., for work on magnetic airborne detection of submarines. In 1942 the Quonset Point Laboratory became Columbia University's Airborne Instruments Laboratory at Mineola, N. Y. From 1942 to 1943, he was a technical advisor on antisubmarine warfare to the Army Air Force at Langley Field, Va., and to the Navy for six months in the South Pacific.

He remained with AIL when it became an independent laboratory in 1945. His activities have included work on special re-

(Continued on page 72A)

Most companies think there's only one way to design a klystron.

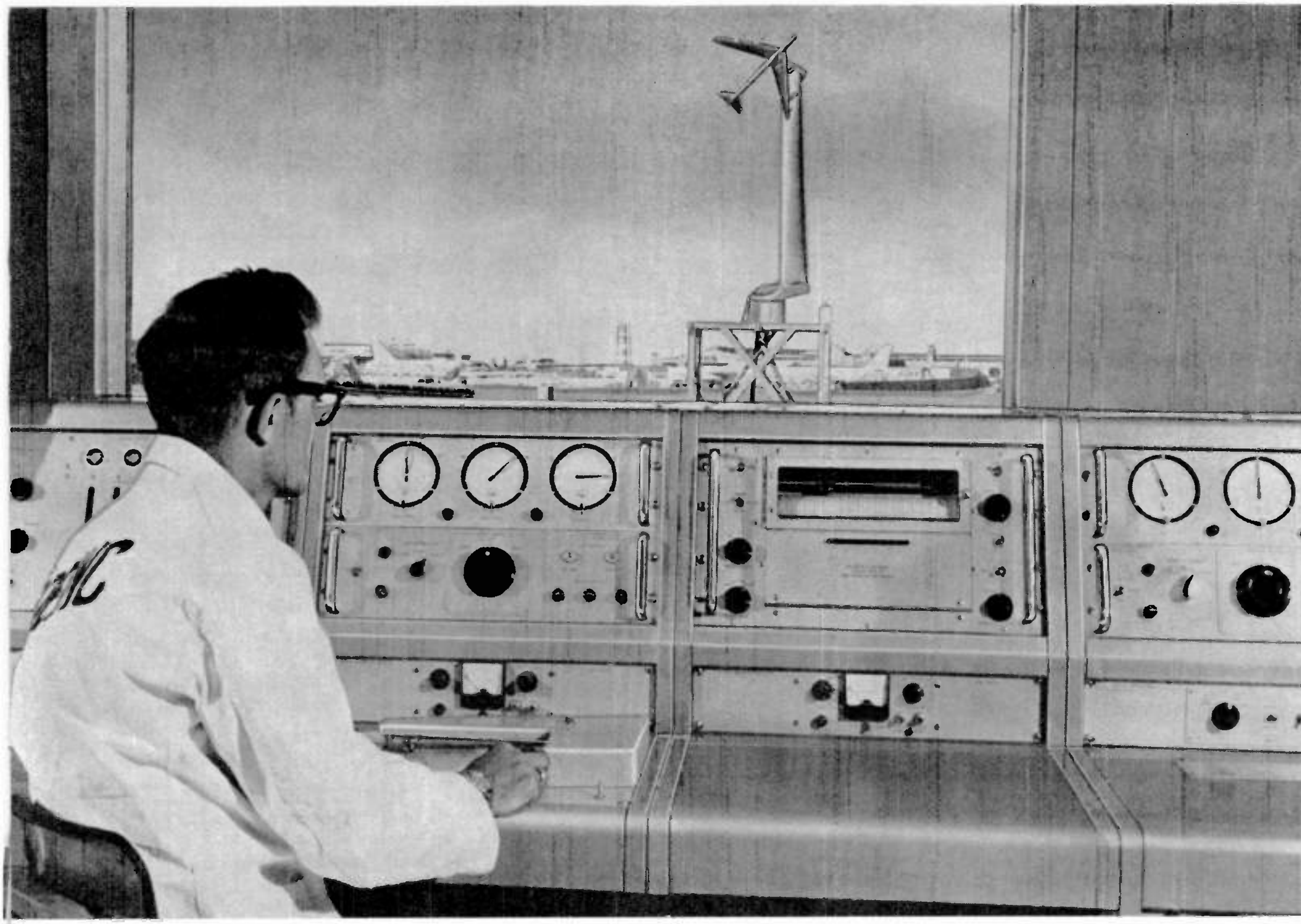


But Eimac knows there are <u>three</u>.

Some rf ranges and requirements call for an internal cavity klystron. Others, an external one. For still others, a combined internal-and-external cavity is best. That's why Eimac designs klystrons *all three ways.* (And why it has more high power klystrons operating throughout the free world than all other makers combined.) Fact is, that's how Eimac designs *every* tube: to meet your specific needs.

For data on Eimac klystrons shown above (4KP40,000SQ, internal cavity; 4K50,000LQ, external cavity; 5K210,000LQ, combined internal-and-external cavity) contact your Eimac representative or write: Power Klystron Division, Eitel-McCullough, Inc., San Carlos, Calif.

Eimac



Scientific-Atlanta antenna pattern recording console at Boeing Airplane Company, Wichita, Kansas, with Scientific-Atlanta model range tower in background.

Advancing the Art of Aircraft Antennas

Boeing uses versatile Scientific-Atlanta equipment to design and evaluate antennas for B-52H bombers

A Boeing B-52H global bomber packs more total firepower than that expended by all the Allied and Axis bombers in World War II. Each B-52H will carry four Skybolt missiles plus a potent assortment of other weapons. Equipped with penetration aids, including electronic countermeasures (ECM) and decoys, the B-52H can strike as many as five military targets on a single mission. It is produced for the Air Force's Strategic Air Command at the Wichita, Kansas, Division of Boeing.

Obviously, the design of antennas for such an aircraft demanded nothing short of "state of the art." As it turned out, Boeing engineers *advanced* the state of the art in the design of ECM antennas for B-52Hs and B-47s. They were aided significantly by a new antenna test facility, consisting predominately of Scientific-Atlanta equipment—including pattern recorders, wide range receivers, signal sources, and a model range tower.

The foremost advantage of Scientific-Atlanta instrumentation is *versatility*. Complete frequency coverage is provided with recordings proportional to voltage, power, or db in either rectangular or polar coordinates. Owing to the equipment's wide frequency coverage, sensitivity, and flexibility, many other laboratory measurements can be made including calibration of microwave attenuators, insertion loss and gain measurements. These, and other features which enable Boeing to derive data faster and easier, have resulted in significant savings of research time.

There's one other point that should be mentioned. At Boeing's antenna test facility, Scientific-Atlanta's equipment has operated with good reliability. Whenever help is needed, Scientific-Atlanta engineers are there in a hurry.

Scientific-Atlanta will accept full responsibility for the design, construction, and manufacture for any antenna test facility that suits your needs. For details, write

SCIENTIFIC-ATLANTA, INC.
2162 Piedmont Road, N.E. • Atlanta 9, Georgia
Phone: TRinity 5-7291

Excellent employment opportunities exist for electronic, microwave and mechanical engineers.

IRE People

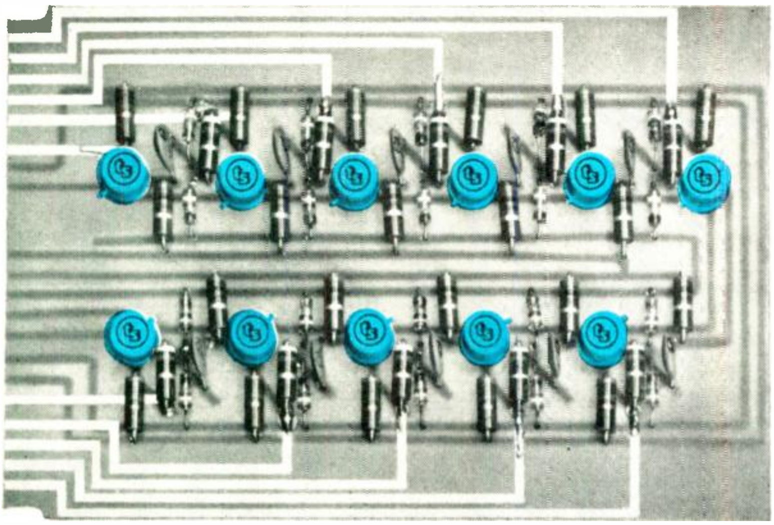
(Continued from page 69A)

ceivers, automatic antenna pattern measurement systems, and advanced magnetic airborne detection systems. In 1949, he became Assistant Supervising Engineer of the Special Devices Group in AIL's Research and Engineering Division and was associated with numerous projects in the fields of microwave instrumentation, microwave components (including new types of rotary joints and receiver front ends), microwave techniques (including STRIPLINE of which he was a co-inventor), and countermeasures systems. In 1955, he was appointed Department Head of the Department of Special Systems and Components. Projects in this department have encompassed numerous engineering fields, including antennas, microwaves, electronics, passive and active countermeasures, and other special electronic and microwave systems. Project STAR (Space Technology and Advanced Research) began in this department, and when it became a full AIL space program early in 1958, he was appointed Program Director. Project STAR is engaged in high priority systems projects in the field of space technology. Areas of activity include advanced development, engineering, and fabrication of highly-reliable equipment in the fields of microwaves, electronics, data processing and analysis.

(Continued on page 74A)

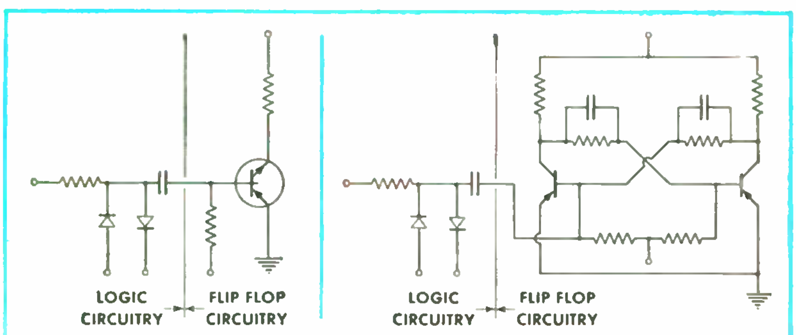
SLASH COMPONENT REQUIREMENTS with Dynaquad

new Tung-Sol 4-layer PNPN Bistable Transistor

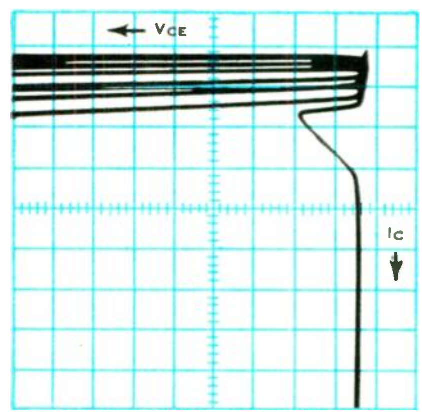


Here is a shift register panel which demonstrates the enormous component savings and the substantial reduction in backboard wiring and circuit complexity that can be achieved through the use of Tung-Sol Dynaquad transistors. This component advantage is typical of the assembly economy (especially with printed circuitry) that can be realized in many other applications, including: computer memory and readout; core drivers; relay activators; sweep generators; and high energy switching. For full technical details write: Tung-Sol Electric Inc., Newark 4, New Jersey.

1 printed circuit board assembly performs the job of 3. 10-bit shift register designed with Tung-Sol Dynaquad transistors. Just one assembly is required where 3 are necessary when designed with conventional components.



7 components replace 14. Comparison of a single stage of the 10-bit shift register designed with Dynaquad transistors (left) and conventional components (right) shows the circuit simplicity and component reduction obtained with Tung-Sol's new germanium multilayer alloyed junction transistor.



Dynaquad is a three-terminal device featuring regenerative switching characteristics. One terminal—the base—serves as the control gate for initiation of the regenerative action. It permits turn-on and turn-off by bursts of drive power. In this way, a small signal controls large amounts of energy in a ratio not approached by conventional 3-layer junction transistors. Trace shows Dynaquad collector characteristics with base current turn-on.

2N1966	2N1967	2N1968

Typical electrical characteristics and ratings.

P_C	collector dissipation at 25°C	120	MW
BV_{CES}	collector breakdown voltage	− 50	volts
I_{CS}	sustaining current	15	Ma
I_B (on)	base turn-on current	0.1	Ma

Technical assistance is available through: Atlanta, Ga.; Columbus, Ohio; Culver City, Calif.; Dallas, Tex.; Denver, Colo.; Detroit, Mich.; Irvington, N.J.; Melrose Park, Ill.; Newark, N.J.; Philadelphia, Pa.; Seattle, Wash. In CANADA: Abbey Electronics, Toronto, Ont.

TUNG-SOL®

IRE People

(Continued from page 72A)

and ground support. Emphasis in the program is on analysis of the operational problem, system design of the space and ground equipment, and reliability.

He has presented papers on microwave STRIPLINE developments, rotary joints, antennas, and countermeasures and space electronic systems at various technical symposia. He is also the author of "The Magnetic Airborne Detector," which appeared in "Advances in Electronics," Academic Press, New York, N. Y., 1952. In 1951 he was a special lecturer for a graduate course in microwaves at Hofstra College, Hempstead, N. Y.

Mr. Fromm is a member of Phi Kappa Phi, Tau Beta Pi, and Eta Kappa Nu. Also, he is a member of the American Rocket Society and the American Management Association.

❖

The appointment of **Robert R. Goldsborough, Jr.** (M'53–SM'55) as Manager of the Engineering Operations Department at the Reconnaissance Systems Laboratory of Sylvania Electric Products, Inc., has been announced.

He previously was manager of the USD-7 program for development and design of airborne electronic equipment at Sylvania's Mountain View Operations.

The Reconnaissance Systems Laboratory, a part of Sylvania Electronic Systems, a major division of the company, is engaged in research and development of advanced reconnaissance systems for the armed services.

Mr. Goldsborough joined the company in 1933 as a Senior Engineer at the Electronic Defense Laboratories in Mountain View, Calif. He was later named head of the systems test and integration section, and, in 1958 was transferred to the Waltham Laboratories of Sylvania Electronic Systems to assume new responsibilities as manager of the USD-7 program. He returned to Mountain View the following year when the program office was transferred to the West Coast.

A 1946 graduate of the U. S. Naval Academy, Annapolis, Md., he has since taken graduate work at Arkansas State Teachers College, Conway, and Stanford University, Stanford, Calif. He is a member of the American Rocket Society and the Naval Academy Alumni Association.

❖

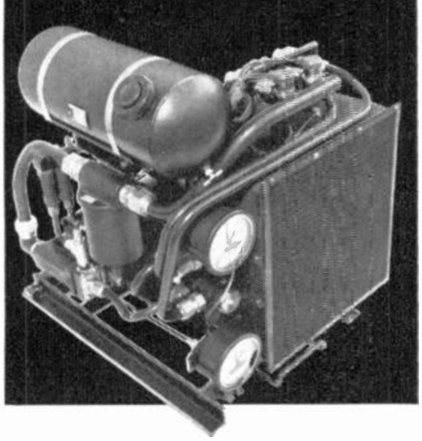
Radiation, Incorporated has announced the appointment of **John T. Hartley** (S'54–M'56) as Vice President, Corporate Marketing.

He has been with the company since 1956, and is now responsible for an integrated corporate sales effort for all divisions and subsidiaries. He will supervise evaluation and assignment of product line areas, and the establishment and operation of Radiation's District Sales Offices. He reports directly to the President.

(Continued on page 77A)

Environmental conditioning

for missile ground support systems



AiResearch electronic cooling units for U.S. Army Hawk missile mobile ground radar equipment require only half the space originally allotted. These lightweight production units, with a heat rejection capacity of 10 KW, measure 20″ x 21″ x 24″.

A complete system package, the liquid-to-air unit includes an accumulator, pump, heat exchanger, fan, switches and valves.

Contact AiResearch early in your planning and design stage for greater reliability, smaller unit size and weight. AiResearch is the leading designer and manufacturer of advanced electronic conditioning equipment and systems for missile and ground support applications.

Environmental conditioning equipment has been produced for the following electronic systems:

Detection • Communication • Control • Ground Support • Guidance

Write for literature today.

THE GARRETT CORPORATION

AiResearch Manufacturing Division

Los Angeles 45, California

CUSTOM ENGINEERED

CRYSTAL FILTERS

100 kc to 30 mc

Bliley crystal filters are custom engineered for your production requirements, and may be designed for a wide range of frequency and selectivity characteristics. Dependent upon shape factor and frequency range, bandwidths are obtainable from .01% to 8% of center frequency. Call upon Bliley for a qualified analysis and quotation on your crystal filter application.

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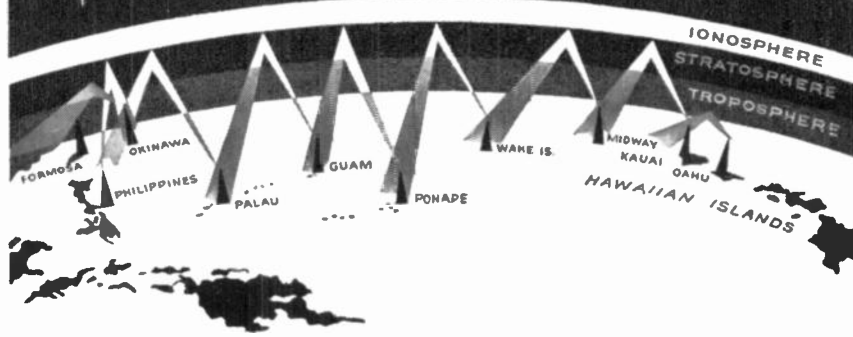
Bliley **BLILEY ELECTRIC COMPANY**

UNION STATION BLDG., ERIE, PENNSYLVANIA

Another reason...
the world becomes smaller



7,500-mile Pacific Scatter Communication System linking major command posts from Hawaii to Formosa was recently designed and built for the U. S. Army Signal Corps by

Page COMMUNICATIONS ENGINEERS, INC.

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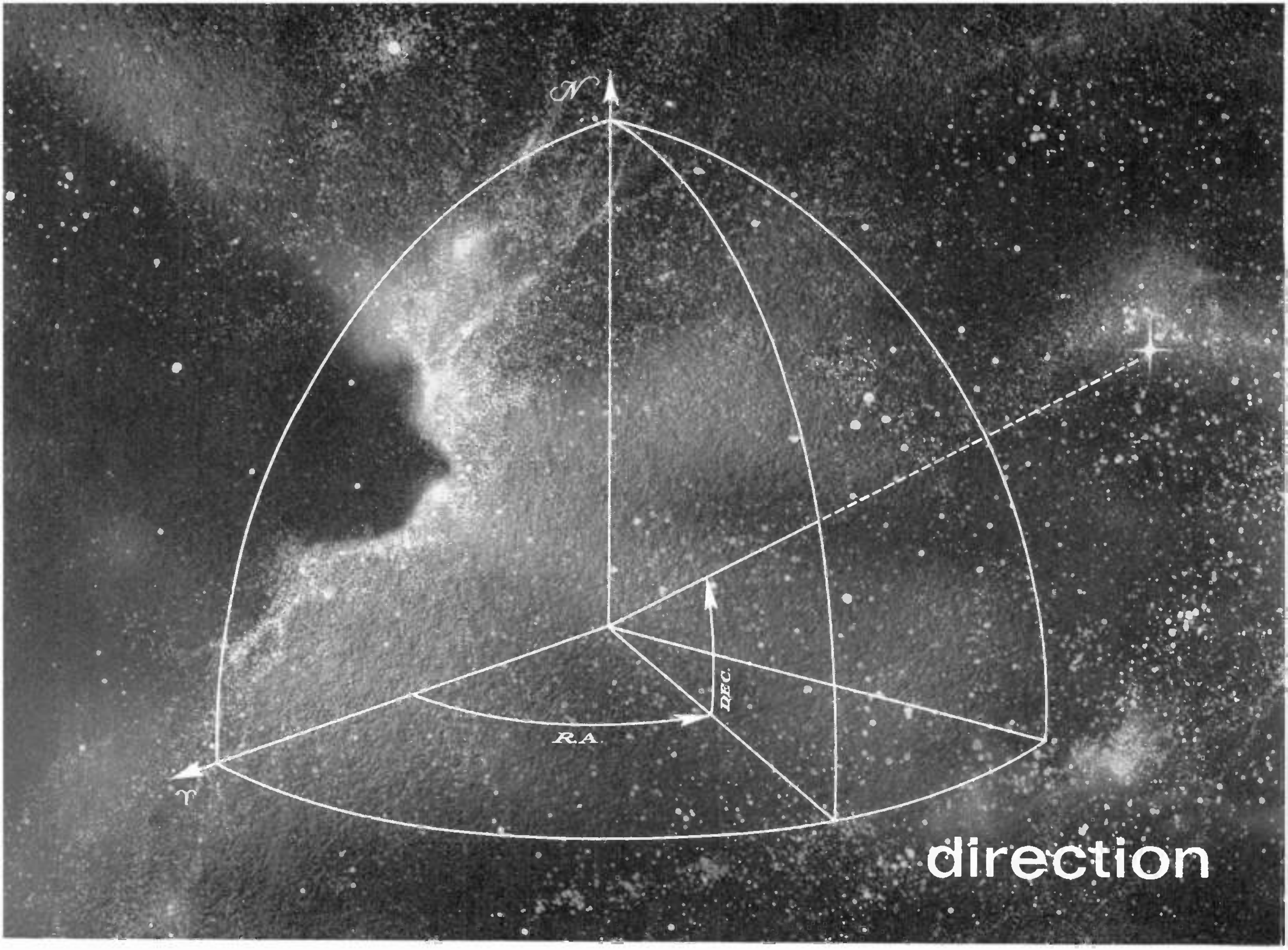


Once off the firm footing of earth, the most critical need of any vehicle is for precise *direction*. The straight course of a sub, a ship, a jet . . . the precision track of missile or space vehicle . . . these result from a directional reference of superior accuracy; the kind provided by gyros made at Sperry.

Whatever the application, gyros by Sperry have a common denominator: stability. Sperry is dedicated to, concentrates upon, stability – absolute directional accuracy, absolute repeatability. The result is seen in the widespread technological successes achieved at the direction of the Sperry gyroscope.

Sperry gyros are the precise directional reference in these and many other applications: pinpoint navigation for the Polaris subs; automatic steering for advanced vessels such as the N. S. Savannah; automatic flight controls for the DC-8 and other aircraft; precision bomb-nav system for the B-58 "Hustler"; miniaturized gyros for space vehicles (symbolized in main illustration).

SPERRY

(Continued from page 74A)

A Navy veteran, he served with the Atlantic Fleet as a staff communications officer, and worked in radar and CIC communications. A native of Jacksonville, Fla., he graduated from Alabama Polytechnic Institute, Auburn, with the B.S. degree in chemistry and the bachelor's degree in electrical engineering. He joined Radiation from a position as instructor in electrical engineering at Auburn.

He became part of Radiation Incorporated's research staff in 1956, then rose to staff engineer, technical advisor in the Advanced Programs Department, and later assistant manager of the Advanced Programs Department. He was made Director of Marketing in March, 1959. Prior to being named Marketing Vice President, he served as executive assistant to the Vice President, Operations, and on special assignment as Program Manager for the Company's Dyna-Soar project.

Mr. Hartley is a member of the American Management Association, and the honorary fraternities, Eta Kappa Nu, Tau Beta Pi, and Phi Kappa Phi.

❖

Ryan Electronics has named **Herbert H. Hauser** (SM'59) as technical assistant to the Chief of Advanced Design to serve as technical liaison for European Affairs.

H. H. Hauser

He came to the United States from France in 1959 to serve as Chief of the Technical Liaison Department of Nord Aviation's Washington, D. C., office.

His experience includes seven years in the design of countermeasure receivers and transmitters, instrument landing system receivers, broadband microwave circuits, transistorized power supplies, and servo systems. He holds patents for a frequency meter and a multichannel receiver.

Before coming to the United States, he held several key posts at Compagnie Generale de TSF (CSF), the largest electronics corporation in France.

He speaks French, German, and English. He was graduated from the University of Paris with the Bachelor of Science degree in mathematics, and the Master's degree in electronic engineering. He later studied business management at the Sorbonne Law School.

Mr. Hauser is a member of the American Rocket Society, the Institute of Aerospace Sciences, the Institute of Electrical Engineers (Britain), and the Societe Francaise des Radio Electriciens (France).

❖

Dr. J. Richard Hechtel (SM'59) has joined the staff of the Research Laboratory

(Continued on page 79A)

IS POWER STABILITY A REQUIREMENT FOR PARAMETRIC PUMPS?

by IRVING R. VERSOY
Engineering Manager, Instrument Division
LABORATORY FOR ELECTRONICS, INC.

Parametric Amplifiers, in contrast to conventional or transistor amplifiers, utilize A-C sources of power rather than D-C sources. These A-C power supplies are commonly called pumps and the operation of the amplifier depends on converting some of the pump energy into signal energy.

There are two main requirements for any amplifier: that it have a stable gain characteristic, and that it introduce into the system no more than the absolute minimum of noise. Examination of these two design factors indicates that the power source of the amplifier must have specific characteristics. Considering the requirements of the parametric amplifier, we find that the pump must have both power and frequency stability for optimum results.

Oversimplified, the effect of pump power on parametric amplifier gain can be compared to the effect of varying the B+ supply of a more conventional amplifier and indeed, within limits, variation of pump power can be used as an amplifier gain control. A typical curve of pump power vs amplifier gain is shown in figure 1.

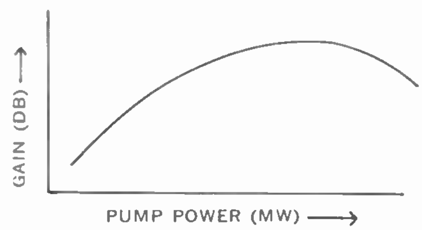


Figure 1

From this figure it is obvious that optimum pump power is required for maximum gain, and also that slight variations in pump power can have a significant effect on amplifier gain.

The noise figure of a parametric amplifier is given approximately by:

$$NF = 10 \log\left(1 + \frac{\omega_1}{\omega_2}\right)$$

Where ω_1 = signal frequency
ω_2 = idler frequency

Additionally the pump frequency, ω_3 is related to ω_1 and ω_2 by the following possible relations.

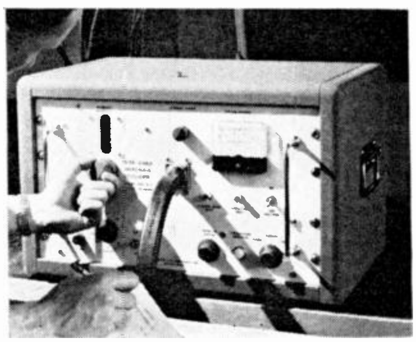
$$\omega_3 = \omega_2 - \omega_1$$
$$\omega_3 = \omega_2 + \omega_1$$
$$\omega_3 = \omega_1 - \omega_2$$

In the case where $\omega_1 = \omega_2$, the minimum achievable noise figure is 3 db. However, when $\omega_2 >> \omega_1$, the noise figure can become better than 1 db.

For this reason, it is generally taken as a guide figure that ω_2 = 4 to 6 ω_1. For lower frequency, amplifiers UHF through S-band, the X-Band pump is chosen. At higher frequencies, a compromise must be made between available power sources and minimum noise figure.

So far, little has been said concerning the requirement for frequency stability of the pump. This occurs because the bandwidth of the amplifier is a function of the idler circuit Q; the higher the Q, the narrower the bandwidth. Increase of bandwidth can be attained only at the expense of more pump power. It is, therefore, very desirable to be able to set the pump on frequency and forget it, particularly if the pump power is also stable.

LFE's Model 814 series of Stable Microwave Oscillators are ideally suited to attaining the ultimate in both stable gain and pass band.



Here are the pertinent specifications:

Frequency:
Short-term stability
5 parts in 10^8 peak deviation
(i.e. 500 cycles peak @ 10 kmc)
Long-term stability 1 part in 10^6 per hour
Power Stability
At any set frequency
± 0.25 db/hour into a constant load

These figures are based on normal environmental conditions.

Short-term frequency stability implies a period of time. Actually, however, our definition is the measure of periodic or transient perturbations that take place about the operating frequency, F_O, over a disturbance or modulating frequency band of 15 c/s to 10 kc/s.

This is determined by the response characteristic of the metering amplifier in the stability tester used, the LFE Model 5009. This corresponds to an averaging period of a minimum of 100 microseconds and a maximum of 0.066 seconds. As a matter of interest, noise, a non-periodic function, introduces the greatest amount of fm residual. As expected, the principal periodic contribution is from power line frequencies.

Long-term drift is brought about almost equally by variations in temperature and humidity, and drift of the D-C amplifier. The components most affected are the invar dual-mode discriminator cavity, and the diodes associated with it. The specification figure is a maximum to be expected over a period of one hour under usual environmental conditions.

In the 814 Series, there are only two primary operating controls, the main one is directly calibrated in frequency to ±0.1%. Loss of stabilization is signalled by a warning light. An additional feature of the 814 Series is a built-in 1 kc AM modulator, with provision for FM modulation by an external modulator. Wherever frequency and power stability really count, you can count on LFE's Stable Microwave Oscillators.

SIGNIFICANT USES:

- Maser and parametric amplifier pumps
- MTI laboratory and production tests
- Secondary microwave frequency standards
- Microwave spectroscopy
- Backscatter measurements
- Q measurements
- Simplified dielectric constant and tangential loss measurements.

SERIES 814
STANDARD MICROWAVE
OSCILLATOR LINE

Oscillator	Frequency in mc/s	Nominal Power (milliwatts)
S-BAND		
814-S-1	2500-3050	75
814-S-2	2950-3600	80
814-S-31	3700-4300	1
C-BAND		
814-C-1A	5100-5900	60
814-C-10	5400-5900	200
814-C-31	5925-6425	1
X-BAND		
814-X-1	8500-10,000	80
814-X-2	9000-10,500	55
814-X-21	8500-10,000	500
K-BAND		
814-K-21	12,400-14,000	100
814-K-22	14,500-17,500	100
817-K-25	24,000-27,000	40

In addition to the standard models whose power and frequency ranges are shown above, packaging and engineering modifications are available to meet the physical requirements of systems in which an 814 is a component. For Bulletin 814, or for particular information, write to Mr. Perry Pollins. If you are also interested in measuring stability, ask for a bulletin on the Model 5009 Stability Tester.

LABORATORY FOR ELECTRONICS, INC.
714 BEACON STREET, BOSTON

IRE People

(Continued from page 77A)

of Litton Industries' Electron Tube Division as a senior scientist.

For the last three years he has been head of the microwave tube branch, Naval Ordnance Test Station, China Lake, Calif. Previously, he was employed for seven years as research scientist and head of the microwave tube department for Telefunken, GmbH, at Ulm, Germany. He also was associated with C. Lorenz A.G., Oberesslingen/Stuttgart, Germany, for two years.

Since receiving the doctorate at the University of Technology, Munich, Germany, in 1940, he has specialized in reflex klystrons, traveling wave tubes, triodes and the interaction of electrons.

Dr. Hechtel is associated with the Research Society of America and the Professional Group for Electron Devices.

❖

PRD Electronics, Inc., Brooklyn, N. Y., has announced the promotion of **Dr. Samuel Hopfer** (SM'58) to the position of Manager of Research, it has been disclosed.

S. Hopfer

He is a graduate of West Virginia University, Morgantown. After completion of his undergraduate work in 1943, he became a physics instructor in the ASTP Program at Westminster College, New Wilmington, Pa. Following the termination of this program, he continued his graduate studies in physics at Cornell University, Ithaca, N. Y., on a teaching fellowship. He received the M.A. degree in 1946, at which time he joined PRD. In 1954, he received the Ph.D. degree in physics from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y.

He was appointed to Section Head in 1949, became Head of the Microwave Research Department in 1956, and was promoted to Manager of the Advanced Engineering Department in 1960.

Since joining PRD, he has been engaged in the fields of microwave theory and instrumentation. He has made substantial contributions to the theory of coupling in multimode systems, the development of broadband transmission systems, the engineering of Ammonia Maser Oscillators, and the design of numerous novel microwave instruments.

He is an Adjunct Professor at the Polytechnic Institute of Brooklyn, a member of the American Physical Society and of the Sigma Xi and Sigma Pi Sigma Societies. He has delivered numerous papers, many of which have been published in the IRE Transactions, in various Symposium Records, PRD Reports and other scientific journals.

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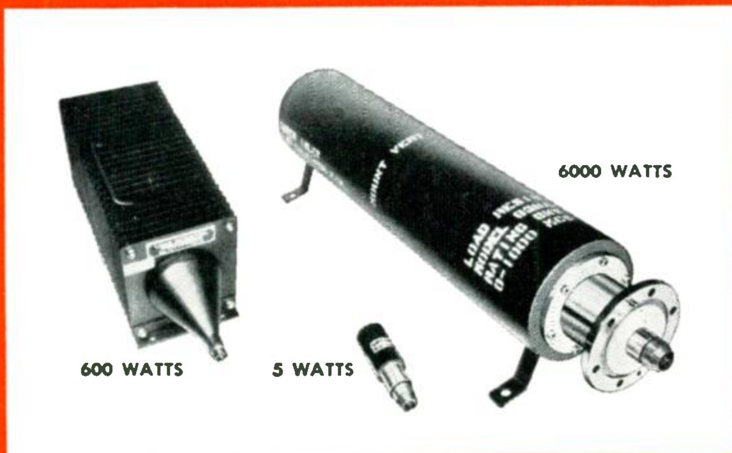
(Continued on page 138A)

MicroMatch®

RF LOAD RESISTORS COVER THE RANGE:

TO 6000 WATTS AND 3000 MCS.



MicroMatch RF Load Resistors provide the virtually reflectionless terminations needed for accurate RF power measurement. They serve many useful purposes as non-radiating RF power absorbers, particularly in lieu of antenna systems during the measurement and alignment phase of transmitter operation.

Other useful functions are in conjunction with feed-through wattmeters to form excellent absorption-type wattmeters, and as a load for side-band elimination filters or high power directional couplers.

SPECIFICATIONS		RF LOAD RESISTORS	
MODEL NO.	FREQUENCY RANGE (mcs)	RF POWER DISSIPATION (watts)	RF CONNECTORS
601	0-3000	5	N, C or BNC
603	0-3000	20	N, C or BNC
633	0-3000	50	N, C or HN
634	0-3000	150	N, C or HN
635	0-3000	200	N, C or HN
636	0-3000	600	N, C or HN
638	0-2000	6000	3⅛" flange

Many other special models have been designed and manufactured to meet your particular space and input connection requirements.

For more information on RF Loads, Directional Couplers, Tuners, and RF Wattmeters, write:

MC Jones

M. C. JONES ELECTRONICS CO., INC.

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SUBSIDIARY OF THE Bendix CORPORATION

SPRAGUE PIEZO-ELECTRIC CERAMIC ELEMENTS

ELEMENTS FOR ALL APPLICATIONS AS WELL AS COMPLETE TRANSDUCER ASSEMBLIES FOR MOST APPLICATIONS, SUCH AS UNDERWATER SOUND AND VARIOUS ORDNANCE AND MISSILE DEVICES.

Sprague-developed mass production and quality-control techniques assure lowest possible cost consistent with utmost quality and reliability. Here too, complete fabrication facilities permit prompt production in a full, wide range of sizes and shapes.

Look to Sprague for today's most advanced ceramic elements—where continuing intensive research promises new material with many properties extended beyond present limits.

YOUR INQUIRIES ARE INVITED

WRITE FOR LITERATURE

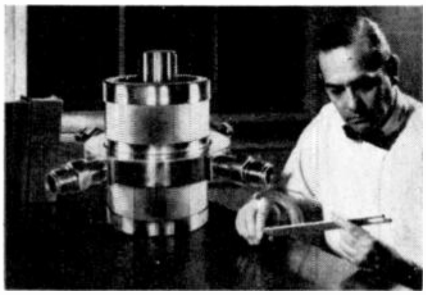
SPRAGUE ELECTRIC COMPANY
235 Marshall Street, North Adams, Mass.

SPRAGUE®
THE MARK OF RELIABILITY

RCA SUPER POWER TUBES

For Widest Choice at Highest Power

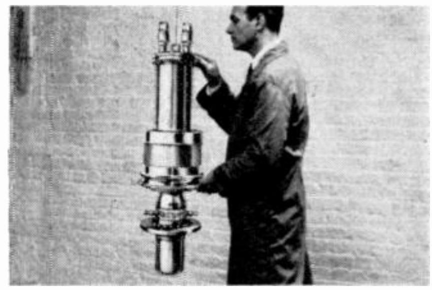
RCA SUPER POWER TUBES are now available in a new, wide choice of ratings for CW, Pulse, and Hard-Tube-Modulator Applications.



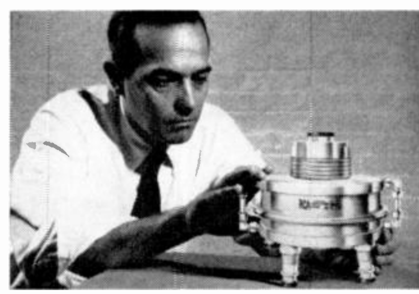
RCA Super Power Tubes have made operation with conventional circuit tuning techniques a practical reality in high-power UHF.

Electron optics—ingeniously coupled with mechanically and electrically rugged envelope structure—enable you to develop unprecedented levels of power with straight-forward circuitry. Cumbersome magnetic field beam-focusing is not required. X-ray shielding is not required in essentially all applications. Phase stability (less than 0.25° per 1% change in supply voltage) meets exacting requirements in every application where phasing is a major parameter. Where sophisticated modulation is required, RCA Super Power Tubes take the extra load in stride—regardless of complexity of data impressed on the carrier.

Extensively proved in commercial and military applications, RCA Super Power Tubes today can be supplied in a variety



of types, and ratings. Variants of many types also can be supplied for specific application. For complete information, get in touch with the RCA Field Office near-



est you. Or write: Marketing Manager, RCA Industrial Tube Products, RCA Electron Tube Division, Lancaster, Pa.

Type	Prototype	Description	Power Output at Frequency	Max. Freq. (Mc)	Power Gain	Typical Plate Kilo-volts	Typical Pulse Plate Kilo-volts	Typical Duty Factor	Typical Pulse Width µSec
CW APPLICATION									
6448		Beam Power Tetrode	14Kw/400Mc	1000	35	6.5	—	—	—
6806	6448	Beam Power Tetrode	25Kw/400Mc	1000	85	8.5	—	—	—
2029	6806	Beam Power Tetrode	25Kw/400Mc	1000	85	8.5	—	—	—
A-2548*	2041	Beam Power Tetrode	50Kw/425Mc	575	50	10.0	—	—	—
A-2690*	6952	Beam Power Tetrode	70Kw/450Mc	900	50	10.0	—	—	—
6949		Shielded-Grid Triode	500Kw/425 Kc	75	250	17.5	—	—	—
A-15157*	A-2342*	Shielded-Grid Triode	500Kw/110Mc	150	25	17.0	—	—	—
A-2335-C*		Double-Ended Triode	75Kw/550Mc	1000	15	9.0	—	—	—
A-15161*	A-15037*	Double-Ended Triode	300Kw/425Mc	600	10	9.0	—	—	—
PULSED RF APPLICATION									
2041	A-2515-H*	Beam Power Tetrode	180Kw/450Mc	575	100	24**	—	0.06	2000
4605	6952	Beam Power Tetrode	2Mw/425Mc	575	100	—	50	0.004	13
6952		Beam Power Tetrode	2Mw/425Mc	575	100	—	50	0.004	13
A-2589*	2041	Beam Power Tetrode	180Kw/575Mc	1000	90	—	26	0.06	2000
A-2590*	6952	Beam Power Tetrode	1Mw/940Mc 200Kw/940Mc	1000 1000	40 40	— —	40 25	0.005 0.05	20 200
A-2606*	2041	Beam Power Tetrode	1.25Mw/425Mc	900	100	—	50	0.003	13
A-2645*	6952	Beam Power Tetrode	275Kw/425Mc	900	100	25**	—	0.06	2000
A-2669-A*	6952	Beam Power Tetrode	275Kw/425Mc	900	100	20**	—	0.06	2000
4603	6949	Shielded-Grid Triode	1.5Mw/ 50Mc	100	125	—	32	0.09	2000
6950/2039	A-2342*	Double-Ended, Shielded-Grid Triode	1.5Mw/200Mc	250	30	—	30	0.05	2000
2054	A-2346-N*	Double-Ended Triode	5Mw/440Mc	600	25	—	33	0.06	2000
7835	A-2346-f*	Double-Ended Triode	5Mw/250Mc	300	35	—	34	0 006	25
A-15025-A*	A-2346-f*	Double-Ended Triode	5Mw/250Mc	300	35	—	34	0.02	20
A-15040*	A-15037*	Double-Ended Triode	5Mw/425Mc	600	30	—	30	0.008	20
A-15038*		Coaxitron	5Mw	Broadband 385-465	20	—	25	0.008	30
A-2344*	A-2335*	Double-Ended Triode	5Mw/900Mc	200-1300	20	—	50	0.01	10
HARD-TUBE MODULATOR APPLICATION									
A-15034-C*	6949	Shielded-Grid Triode	11Mw	—	—	40	—	0.05	2000
A-15042	5831	Beam Triode	44Mw	—	—	40	—	0.005	20

*Developmental Type Number **Screen-and-Grid-Pulsed Amplifier Service

RCA Electron Tube Division Field Offices

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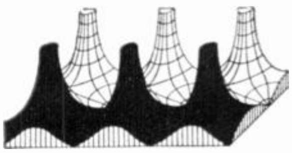
RADIO CORPORATION OF AMERICA

June, 1961
Vol. 49 No. 6

Proceedings of the IRE

Poles and Zeros



International Scene. Poles and Zeros, since its inception in 1956, has commented eight times on some aspect of IRE international activities. These comments have covered such significant items as the establishment of sections outside the United States, the first meeting of the Board of Directors held outside of the United States, cooperation with international scientific bodies, and the number of members outside of the United States. Regular readers (if any) of Poles and Zeros may have wondered at this repetition. The emphasis on international activities on this page has been a purposeful one. IRE as an organization, although founded in the United States, not only recognizes the world-wide interest in electronics but also recognizes that scientific knowledge has no respect for geographic boundaries.

It is significant that the increasing emphasis on IRE activities outside the United States has its origin among foreign members and nonmembers who recognize the service that IRE renders to the profession through its publications, meetings and sections. A specific example of this recognition is revealed by the editorial "Why in the World in Europe?", written by Bruce B. Barrow, as an introduction to the March 1961 issue of the TRANSACTIONS ON COMMUNICATIONS SYSTEMS. This editorial is reprinted in this issue of the PROCEEDINGS and is called to your attention for the importance it places on IRE in international affairs as viewed by an eminent member of the IRE living in the Netherlands

The issue of the TRANSACTIONS ON COMMUNICATIONS SYSTEMS, referred to above, is devoted to the papers presented at the International Symposium on Data Transmission held at the Delft Institute of Technology, The Netherlands, in September, 1960. This was the first symposium sponsored by the IRE outside of North America. It originated through the initiative of the Benelux Section of the IRE. The Benelux Section was effectively supported by the Netherlands Radio Society, the Royal Institute of Engineers (Telecommunications Section), and the IRE Professional Group on Communications Systems. Over 500 persons registered for the symposium; about one-half were from The Netherlands and the other half were from thirteen other countries. The symposium marks a distinct extension of the IRE international activities.

Other recent emphasis on international events is found in the December, 1960 issue of the TRANSACTIONS of the Professional Group on Circuit Theory. The issue was jointly sponsored by the Professional Group and by the Union Radio Scientifique International in memory of Balthasar van der Pol (a native of The Netherlands, whose pioneering contributions to nonlinear circuit theory began nearly forty years ago). Contributions to the issue come from six nations: Holland, France, Japan, England, the USSR, and the USA.

Other IRE activities on the international scene are worth recording here. For many years IRE has participated directly in the activities of the International Electrotechnical Commission (IEC), the International Radio Consultative Committee (CCIR), and the Union Radio Scientifique International (URSI). IRE appoints representatives to serve on the United States National Committee or delegation for each of these bodies. The Professional Group on Automatic Control represents IRE as a member of the Automatic Control Council. This council is a member of the International Federation of Automatic Control. It held its first international congress in Moscow in 1960. The Professional Group on Electronic Computers is a member of the American Federation of Information Processing Societies, a member body of the International Federation of Information Processing Societies. It held its first international meeting in Paris in 1959, and it will hold its second meeting in Munich in 1962. The Professional Group on Biomedical Electronics has participated in three international conferences of the International Federation for Medical Electronics. The Professional Group will be the host for the fourth conference to be held in July, 1961. The Professional Group on Education will be a cosponsor of an International Conference on Electrical Engineering Education to be held in September, 1961.

In further recognition of the increasing interest in IRE activities and a demand for greater accessibility of IRE services and publications from abroad, the Board of Directors at its meeting last January established an Ad Hoc Committee on IRE International Activities Outside of Existing Regions. Among other assignments the Ad Hoc Committee is charged with the responsibility of assisting in the establishment of IRE Sections abroad. Poles and Zeros will present further information on the activities and accomplishments of this committee.

Membership. An interesting point revealed by the most recent Report of the Secretary is the character of the growth of IRE. Most emphasis on growth of any organization is usually placed on total membership figures, and IRE has pointed out this phase of its growth regularly. The Secretary's report shows the membership totals for the years 1958, 1959, and 1960 as 71,361, 79,166, and 88,479, respectively. A more detailed examination reveals, however, that this steady increase has resulted from increases in all grades except that of Associate. When one recalls that the qualification for Associate grade is only that of "interest in radio engineering or an allied field" the figures above emphasize the truly professional character of the annual membership growth.—F.H., Jr.

W. G. Shepherd

Director, 1960-1962

Dr. William G. Shepherd (A'42–SM'49–F'52) is Professor of Electrical Engineering at the University of Minnesota in Minneapolis. He was born in Fort William Ontario, Canada, on August 28, 1911, and attended public and high schools in Minneapolis, Minn. He entered the University of Minnesota in 1929, where he was awarded the B.S. degree in electrical engineering in 1933 and the Ph.D. degree in physics in 1937.

He joined the technical staff of the Bell Telephone Laboratories in 1937 and remained until 1947. His principal research activities at the Bell Telephone Laboratories were concerned with nonlinear microwave circuits and electron tubes, particularly reflex klystrons for radar applications. In 1947 he returned to the University of Minnesota as Professor of Electrical Engineering. He has been actively engaged in research in the field of physical electronics, particularly studies of electron emission From 1954 to 1956 he served as Associate Dean of the Institute of Technology, and since 1956 has served as Head of the Department of Electrical Engineering

He has served on a number of IRE committees, including the Electron Devices Committee, the IRE Fellow Committee, and the Awards Committee. In 1953 he served as Chairman of the 1953 IRE Electron Tube Conference and in 1954 as Local Arrangements Chairman for the Semiconductor Devices Conference. From 1955 to 1958 he was a member of the Administrative Committee of the Professional Group for Electron Devices

He has been active in the affairs of the International Scientific Radio Union, serving first as Chairman of the U. S. Commission 7, and presently as International Chairman of Commission VII on Radio Electronics. In 1957 he organized the program of the U S. Commission 7 held at Boulder, Colo., and in 1960 the program of the International Commission VII in London.

Dr. Shepherd was awarded a Certificate of Appreciation by the U. S. Navy Department in 1948. He is a member of Sigma Xi and a registered engineer in the State of Minnesota.

Scanning the Issue

Why in the World in Europe? (Barrow, p. 1008)—As noted in *Poles and Zeros* this month, the IRE, through its Benelux Section and the Professional Group on Communications Systems, sponsored its first large international meeting outside North America last September in Holland. This event was important evidence of the quickening growth of the IRE as an international organization. In Europe, for example, the IRE has formed new Sections in the Benelux countries, in Geneva and in Italy just in the last two years. Why in the world Europe? The answer to this important question is provided by the Secretary of the Benelux Section in an editorial reprinted from the March issue of the PGCS TRANSACTIONS.

Frequency Allocations for Space Communications (Joint Technical Advisory Committee, p. 1009)—One of the most important matters now facing the communications field concerns the impending use of earth satellites to provide long-range broad-band communication channels. A major question that must first be answered is how to provide room for this important new communication medium in an already crowded frequency spectrum. The question is of world-wide importance, growing urgency, and involves a wide range of technical considerations. The ITU has, therefore, called a special international conference in New Delhi in 1963 to discuss spectrum allocations for satellite relays. Meanwhile, in the U.S.A., the FCC is already deeply engaged in an inquiry into the matter. As reported in the April *Poles and Zeros* column, the Joint Technical Advisory Committee of the IRE and EIA recently formed a special subcommittee to study the technical problems imposed by satellite relays, including the critical question of whether sharing frequencies with other services is technically feasible. Although the study is still continuing, in order to provide the FCC with initial comments by March 1 the JTAC issued a preliminary report on that date. Because of the widespread interest in communication satellites and the importance of the JTAC studies, special arrangements have been made to present in this issue the principal findings of the report.

A Low-Noise Microwave Quadrupole Amplifier (Ashkin, p. 1016)—Twenty months ago the PROCEEDINGS reported an important new type of fast cyclotron wave parametric amplifier, called the quadrupole amplifier, which gave unusually good low-noise performance at UHF frequencies. Indeed, of the electron beam parametric amplifiers thus far proposed, it has proved to be the most successful. The excellent experimental work described in this paper shows that the quadrupole amplifier can achieve an equally low noise figure at much higher frequencies than previously reported, namely, a double-channel noise figure at 0.79 db at 4137 Mc. This important result will be of great interest not only to tube specialists but to anyone interested in low-noise devices.

Tapered Distributed RC Lines for Phase-Shift Oscillators (Edson, p. 1021)—It is well known that the losses in a multisection lumped RC network can be greatly reduced by tapering the impedance and increasing the number of sections. However, the analysis of such a network is exceedingly complicated. The author develops the much simpler procedure of considering the limiting case in which the number of sections becomes infinite and the network becomes a smoothly tapered distributed RC transmission line. He then works up a set of normalized curves describing the behavior of tapered lines. These curves make it readily possible to estimate the performance of a network containing a finite number of sections and, in particular, are shown to have practical application to the design of phase shift networks for oscillators.

Submillimeter Wave Radiometry (Long and Rivers, p. 1024)—Radiometers serve as a valuable tool for detecting and measuring the temperature of thermal and radio sources in the heavens and studying atmospheric attenuation. Needless to say, there is a lively interest today in sensitivity improvement of radiometers and in their utility over the spectrum. This paper presents a timely review and comparison of existing radiometric techniques at centimeter and millimeter wavelengths, extrapolating the latter to a virtually unexplored region of the spectrum—the submillimeter region.

Results of a Long-Range Clock Synchronization Experiment (Reder, *et al.*, p. 1028)—Several organizations in the U.S.A. and England are jointly studying the feasibility of using VLF transmissions for world-wide synchronization of atomic clocks with an error of less than 5 microseconds. Under the proposed plan, synchronization would be established by flying an atomic clock from the master to the slave clocks. From then on, synchronization would be maintained by VLF signals controlled by the master clock. A brief progress report is presented, giving results on recent flight tests and propagation experiments and indicating that there is no longer any doubt that world-wide synchronization can be accomplished to within a 1-microsecond error.

High-Frequency Power in Tunnel Diodes (Dermit, p. 1033)—A general expression has been developed for the maximum power output of tunnel diodes operating at high frequencies within the nearly linear range of their negative resistance characteristic. From this expression a number of very interesting results are derived which will be of timely and general interest to device engineers, circuit designers, and a broad segment of readers interested in microwave device applications.

Design Theory of Optimum Negative-Resistance Amplifiers (Kuh and Patterson, p. 1043)—An ideal negative resistance can, in theory, provide infinite gain-bandwidth. In reality, however, parasitic elements impose a practical limitation. What maximum gain-bandwidth products, then, can be expected and what methods of synthesis will lead to an optimum design? The bases for answering these questions are provided in this timely contribution

A General Steady-State Analysis of Power-Frequency Relations in Time-Varying Reactances (Smart, p. 1051)—This paper contains some interesting results concerning power distribution among different frequencies in time variable circuits. The analysis is of great generality, exceeding that of the famous Manley-Rowe relations, which are shown to be special cases of the results presented here. This generality makes the analysis an important and timely contribution to the study of many important energy conversion and power amplification devices.

Error Statistics and Coding for Binary Transmission Over Telephone Circuits (Fontaine and Gallager, p. 1059)—The authors explore whether it is feasible to employ error codes to lower the error rates for transmitting digital data over telephone circuits. They find that it is much better to use a coding that will detect an error and then to retransmit that portion of the message, than to use a coding that will correct the error. In view of the rising use of digital communication, the data reported here come at a most timely moment.

Communications Satellites Using Arrays (Hansen, p. 1066)—This paper examines whether satellites might to advantage employ a relatively new type of array which, because of the special way its radiators are interconnected, will always return energy in the direction of arrival (*i.e.*, towards the earth) even though the direction (*i.e.*, the attitude of the satellite) varies $\pm 45°$. This would greatly simplify the orbit control problem and has other interesting advantages as well.

Scanning the Transactions appears on page 1111.

Why in the World in Europe?*

BRUCE B. BARROW†, SENIOR MEMBER, IRE

WHY in the world is the IRE organizing a symposium in Europe? This was a common reaction among those who, early in 1960, read the announcements of the International Symposium on Data Transmission, the first large international meeting to be sponsored by the IRE outside North America. This symposium manifested the continued development of the IRE as an international organization—it did not represent the *birth* of international IRE activity, though it might be seen in some respects as a coming of age.

How can the IRE, with slightly more than 90 per cent of its members living in the U.S.A., call itself an international organization? The question may be answered by asking another. What other adjective can describe a professional society that numbers more than 6500 members outside the U.S.A., that has local sections holding meetings in every continent except Australia, and that draws members from a dozen countries to a meeting in Holland? The Delft Symposium, which was appropriately organized in cooperation with two Dutch professional societies, attracted more than 500 participants. Ninety-five per cent of them, including more than a hundred IRE members, came from Europe. Already the IRE Professional Group on Information Theory has announced its intention of holding a symposium in Europe in 1962, and it is clear that within a few years the IRE will be regularly working, either as sponsor or as a supporting society, in a full program of European symposiums and conventions.

The IRE is American, and the IRE is international. With a nonpolitical organization the two are not mutually exclusive, and this is a point worth emphasizing in a world where political divisions are so sharp. Furthermore, both aspects of IRE activity are as old as the IRE itself.[1]

* Reprinted from IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-9, p. 3; March, 1961.

† SHAPE Air Defense Technical Center, The Hague, Netherlands.

[1] For a thorough discussion of IRE history, see L. E. Whittemore, "The Institute of Radio Engineers—forty-five years of service," PROC. IRE, vol. 45, pp. 597–635; May, 1957.

The international aspect of IRE activity is not new. The founders of the Institute specifically chose a name and constitution without national preference, and sections were operating in Canada and South America twenty years ago. What *is* new is the expansion in international activity, which follows not from a change in IRE policy but from the growth in IRE membership in all parts of world.

In the leading countries of Europe there are well established professional societies working effectively already. Here the IRE, with profit to everyone concerned, can help to obtain speakers and organize joint meetings, and can provide liaison for such large international meetings as the Delft Symposium. But these activities require a local IRE organization, *i.e.*, a section. Even in the European countries with the most vigorous and effective societies, which by no coincidence also happen to be the countries with the largest number of IRE members, there is therefore an appropriate place for IRE section activity. In such countries, IRE sections are not established as independent organizations to compete with local societies; they may, on the contrary, even be officially affiliated with the local societies.

The development of the IRE in Europe comes at a time when the European nations are busily forming ever-closer ties economically, industrially, and politically. Nevertheless, although the need for better exchange of technical information within Europe is now consciously felt, it is a fact that each of the national European technical societies is handicapped by history in building a professional society for a United Europe.

The IRE can build just such a society. The various IRE sections will continue to operate nearly autonomously, free to serve wherever needed and to adapt to local conditions as appropriate. They will, in the not distant future, be joined together in the structure of an IRE *region* (as the Canadian sections were joined long ago). The regional committees, automatically both international and European, will not only help to coordinate professional society activity, but will provide a force to attack the special problems facing the new Europe. This is the challenge facing the international IRE.

Frequency Allocations for Space Communications*

A Report of the
JOINT TECHNICAL ADVISORY COMMITTEE
IRE-EIA

Preface

THIS report has been prepared to examine the technical aspects of space communications with special attention to the use of satellites for commercial trunking purposes. The problems of frequency allocation and of the influence of system design on frequency allocation and frequency utilization are examined.

The Ad Hoc Subcommittee 60.2 of the Joint Technical Advisory Committee of the Institute of Radio Engineers and the Electronic Industries Association was formed to examine the complicated problems of frequency allocations for space communications. To assist the subcommittee in compiling the accomplishments, plans, and technical opinions of industry, universities, and laboratories, the services of the Boulder Laboratories of the National Bureau of Standards and of Stanford Research Institute were obtained. The contributions and findings of both these organizations are contained in this report. From these findings and other documents and reports available, the JTAC has compiled recommendations and conclusions concerning the allocation of the radio-frequency spectrum for space communications; these can be found in Section II of this report.

I. Introduction

In approaching the study a review was undertaken of the information developed for the Federal Communications Commission inquiry into the allocation of frequencies in the bands above 890 Mc, because the inquiry included introductory aspects of the problems posed by the advent of communication satellites. This inquiry—FCC Docket 11866—extended from 1956 to 1960 and concluded that adequate frequencies above 890 Mc are available to take care of the present and reasonably foreseeable future needs of both the common carriers and the private users of point-to-point communication systems. No determinations were made with respect to allocation of frequencies for space communications.

It has now developed that inquiry into the allocation of frequency bands for space communications should be undertaken in order to permit the timely development of plans for experimentation and for operational use of this new medium. The FCC has, therefore, announced that inquiry into the allocation of frequency bands for space communications will be undertaken in the spring of 1961. FCC Docket 13522 was established for this purpose. Statements and data sought by the FCC for consideration in this inquiry are to be responsive to the matters that follow.[1]

1. The nature and extent of experiments which have been conducted or are now in progress involving utilization of frequencies for space communications of various kinds.
2. What segments of the radio spectrum are required for the various space communications functions to be performed, *e.g.*, earth-space-earth, space-space, telemetry, tracking, guidance, command, etc.? How much spectrum space is required for each such function? At what point in time will access to these bands be required for each such function? Does each function require the same degree of protection from interference?
3. Under what conditions can frequency bands for space communications be shared with other services:

    a) in the case of passive relays?
    b) in the case of active relays?
    c) in the case of other space systems and functions?

Replies to this question should recognize that the emissions of the sharing services may be incompatible and should specify (1) the type of sharing service, *i.e.*, fixed, mobile, radio navigation, etc., (2) transmitter powers, (3) antenna gains and orientations, (4) geographical separations, (5) the minimum angle above the horizon to which the earth terminal antennas will be oriented, (6) orbital characteristics of the satellite system being considered, *i.e.*, altitude, degree of randomness, orbital plane, with or without attitude control, with or without station-keeping, etc., (7) the area to be served by the space system, if other than world-wide, and (8) any other pertinent information.

4. Do non-Government entities plan to launch either active or passive communication satellites?
5. If non-Government entities have no plans for launching active communication satellites, by what means would access to active relay satellites be achieved for non-Government communications?
6. Should there be separate or shared frequency allocations for Government and non-Government space communications?
7. Will the receiving sites for space communications systems be generally distributed throughout the United States or limited to specific areas?
8. The purposes to be served by space communications which are not met by other communications systems.
9. Assuming, at least initially, (1) that existing surface communications must continue to function, and (2) that geographical separation is the key to successful sharing of frequency bands; it appears that earth terminals should be located in sparsely settled areas, away from concentrations of communication installations. Therefore, should the Commission, on the basis of criteria developed pursuant to the new issue three, give consideration to amending its Rules at an early date to establish protected geographical areas to be held in reserve for the installation of future earth terminals for civil communication systems via space relays? If such a concept were adopted it might be advisable to prohibit, for example, the use of certain frequency bands between 1,215 Mc and 10,000 Mc within "X" miles of a given site for all uses other than space communication. Comments are requested on geographical areas which might be appropriate for such a protected reserve status and the frequency limits between which it would be applied.

Recognizing the technical and policy problems that concern us nationally and internationally in the utilization of space for communications and to assist in the

* Received by the IRE, April 19, 1961. Abridgement of a JTAC report of the same title, dated March, 1961.

[1] From FCC Notice of Inquiry Docket 13522, as released May 20, 1960; revised and amended by Supplement to Notice of Inquiry Docket 13522, released December 28, 1960.

consideration and resolution of these problems, the JTAC of the IRE and the EIA formed Ad Hoc Subcommittee 60.2 charged with responsibility for investigation in these areas of interest. The subcommittee was asked to review the subject and to recommend the nature of the contribution that JTAC should make to forthcoming technical inquiries.

Information developed for the FCC during earlier hearings has been reviewed together with details of the various systems proposed for the commercial exploitation of satellite communications. A summarization of these systems is contained in this report.

With the cooperation of members of the FCC, a series of technical questions was developed as a basis for study and survey of industry plans, interests, and opinion. These questions were addressed to approximately forty of the nation's leading corporations and universities who were thought to be engaged in or qualified in fields directly contributing to the development of space communications. The questions presented were:

1) What is the feasibility of frequency sharing with other radio services? There is a difference of opinion concerning this vital question, and the answer thereto will have a fundamental bearing upon the ultimate implementation of any satellite system, since the allocation of the necessary spectrum space for a world-wide system of space communications may impair the ability of this country and others to provide for essential terrestrial radio services. While there have been theoretical papers on the ability of space and terrestrial services to share the same portion of the spectrum, we need some program of concentrated experimentation in this field to give more definite answers.
2) What is the estimate of total government and non-Government requirements for circuits and terminal facilities for the near future (five years) and for the more distant future (1980)?
3) What are the estimated requirements of other countries for circuits and terminals in a satellite system?
4) How much spectrum space is necessary to satisfy the estimated requirements, and where in the spectrum should it be located (1–10 kMc, 1–16 kMc, or 1–20 kMc)?
5) What system of modulation is best suited for satellite systems, passive and active? Is the presently advocated system of broad-band feedback FM circuits the best system, or can some other narrower-band system be used with comparable results and more efficient use of the spectrum? This is a very important question, and should receive the attention of competent engineering groups such as this.
6) What is the estimate of knowledgeable engineers concerning the present and future power potentials of transmitters aboard active satellites? The answer to this question, of course, will have a direct bearing upon the very fundamental question of whether space and terrestrial services can live side by side, since it relates to the amount of signal that may be available at the earth terminal receiver location.
7) Probably one of the most fundamental questions relates to the ultimate type or types of satellites to be used in a system: 1) passive, 2) low-level random-orbit, or 3) 24-hour synchronous-orbit. What are the technical factors involved in each system, both on the ground and on the satellite?

The following questions were also presented to determine the extent to which information has been developed that will contribute to further study of the subject and space communications:

1) Do you have scientific reports or system studies reports which can be made available to JTAC for review? If so, please list their titles and authors.
2) Is your organization actively participating in plans for the exploration of commercial satellite communications? If so, please comment.
3) Will your organization be a user of the channels provided by satellite communications other than normal overseas telephone service? Will you utilize the channels for new services of any kind?
4) Reliability is a serious problem for an active satellite. Do you have studies and experimental programs which will help reduce the lifetime problem?

Significant responses were received by letters, reports, or personal interviews from approximately a dozen organizations and individuals in sufficient time to be included in this report. In some instances, these responses have been integrated into the technical discussion in this report. In other instances, the responses are set apart and identified as relating to a specific question. The sources of specific responses are not identified and the responses are not always verbatim. Time has not permitted rechecking the edited responses with the original source for accuracy and emphasis. It should be emphasized that these responses represent only individual views, not the views of JTAC.

The information presented in this report is preliminary in nature and limited by time and funds available and by restrictions imposed by corporate proprietary interests and military security. Recommendations for continued investigation and experimentation are included in the report.

II. Findings and Recommendations of the JTAC Regarding Frequency Allocations for Space Communications

1) Prior to reaching final technical conclusions, more extensive theoretical and experimental research is needed. Accordingly, the tentative nature of the findings which follow should be recognized.

2) Subject to important restrictions, sharing of spectrum space between fixed microwave point-to-point sys-

tems and satellite systems is possible. Among the restrictions that may be required are the following:

a) Separation between terminal locations of satellite systems and of surface microwave systems—Distances from 100 to 150 miles should suffice and, under ideal conditions, distances less than 100 miles may give adequate protection.
b) Control of antenna radiation patterns—Side- and backlobe suppression may be necessary for both surface and satellite systems.
c) Avoidance of terminal locations inadequately protected by terrain features.
d) Limitations on effective radiated power, including harmonics, of both satellite and surface systems.

It seems certain that frequency bands allocated to satellite communication systems must be different, and adequately separated from, those used by high-power ground radars, tropospheric scatter, and mobile communication services.

3) Both government and industry agree that United States and world needs make urgent the attainment of increased common-carrier communication capacity. Even now capital investments on an increasing scale are being made to implement common-carrier communication systems. Both of these considerations indicate the need for prompt action to attain the knowledge necessary adequately to reserve frequencies and geographic locations for common-carrier satellite purposes.

4) Allocation of spectrum space for space communications is a world-wide problem. An Extraordinary ITU Conference in 1963 has been proposed for the purpose of considering allocations for space communications. This Conference is needed; its convening at that time should be made final.

5) A review of plans and programs of organizations interested in space communication applications indicates the existence of diverse opinions on the methods of accomplishing common-carrier satellite relays. An intensive effort to obtain one coordinated view on system design which would permit the establishment of a United States policy and program is most desirable prior to international negotiations. To this end, initial testing of experimental common-carrier relays and the planning of operational commercial systems, coupled with intensive theoretical investigations of the unresolved technical problems involved, should be undertaken as soon as possible.

6) Three major classes of satellite relays have been proposed for commercial purposes: random-orbit low-altitude repeaters, 24-hour synchronous repeaters, and intermediate-altitude repeaters in orbits with accurately controlled periods (6 or 8 hours). Comprehensive system comparisons have not yet been made on which to base firm comparative conclusions in terms of cost, service, and spectrum use. However, there are no known technical features that will cause any one of the classes of systems to be eliminated on the basis of unfeasibility.

7) No one feature of radio propagation or the system-design problem is so predominant as to make any portion of the 2000- to 10,000-Mc band unfeasible for wide-bandwidth common-carrier purposes. Therefore, the decision as to which portion of this frequency band is to be allocated for common-carrier use can be separated from the decision as to which type of system (high-altitude, low-altitude, etc.) is to be employed. The availability of transmitting tubes and other existing equipment for certain specific frequencies does make it desirable to employ these frequencies for initial experiments; however, there is no technical reason why adequate devices cannot be developed for any portion of the 2000- to 10,000-Mc spectrum.

8) Frequencies in the range of 100 to 2000 Mc can be used for narrower-band, common-carrier purposes; however, time variations of polarization rotation and multipath introduce problems of signal degradation and of restriction of transmission bandwidth. These frequencies (100 to 2000 Mc), despite bandwidth limitations, are optimum for direct broadcast types of satellite systems where antennas of low directivity are to be employed at satellite and ground terminals.

9) For the immediate future, frequencies below 10,000 Mc must be made available for satellite common-carrier communications systems. An examination of the region above 10,000 Mc shows promise for use by satellite common-carrier systems, but adequate research and development will first be necessary. These higher frequencies can be usable where terminal locations do not receive high-intensity rainfall or when high standards of circuit reliability are not essential.

It should be possible to develop a system using the region of 2000 to 10,000 Mc for maximum-reliability channels backed up with additional channels above 10,000 Mc for traffic that can tolerate the occasional rainfall outage. Frequencies above 10,000 Mc may also be utilized with space-diversity earth terminals. Terminal separations should be such that heavy rainfall will not be expected simultaneously at two or more of the earth terminals.

10) An optimum modulation system for common-carrier purposes has not been determined. Many interrelated and unresolved system factors are involved, such as:

a) The number and location of earth terminals.
b) The number of communication channels.
c) The available transmitter power.
d) The receiver sensitivity.

The parts of the frequency spectrum above 2000 Mc to be used for point-to-point trunk service should be allocated in broad bands, on the order of hundreds of megacycles.

Broad blocks of frequency space are to be preferred technically to the same amount of spectrum in several small bands, since a continuous band facilitates the possible eventual use of more efficient modulation. Optimum modulation techniques with interference rejection properties applied to both satellite and surface systems should facilitate the sharing of frequencies.

11) Industry has the technical capability of accomplishing the task of designing, launching, and operating common-carrier satellite relays. Many companies have under way technical plans for providing the communication aspects of a common-carrier satellite relay system for global communication.

12) The pioneering space-research achievements of our Government agencies, both civil and military, generated the capability for early development of common-carrier satellite communication systems.

Means should be actively sought so that the pertinent technical information derived from these government satellite programs can serve to stimulate and expedite the most efficient development of commercial communication systems.

III. Recommendations for Experiments and Investigations Needed to Provide Technical Data for Satellite Communication Relays

A. General

While the progress and plans of various organizations were being reviewed, it became apparent that some technical areas had not been explored in the detail that would be desired by a group assigned with final system-design responsibility. In this section, recommendations for further research are presented and discussed. Due to the limited time available for the completion of this report, the recommendations discussed are not intended to cover all possible areas that require research. Those discussed are prominent areas which require examination before extensive application of satellites for common-carrier communications can be made.

Following is an itemization of the programs included in this section.

1) Compile an expanded spectrum usage chart.
2) Conduct a spectrum usage measurements program.
3) Investigate the feasibility of the use of frequencies above 10,000 Mc.
4) Determine the field-strength levels of distant interference sources.
5) Study the design of surface terminal antennas to minimize interference.
6) Study and measure effects of surface site screening.
7) Use computers to simulate the complex ground-interference situation.
8) Study and measure the harmonic radiation of other UHF high-power transmitters which might cause interference.

B. Discussion

1) *Expanded Spectrum Usage Chart:* Only limited knowledge is available concerning the use of the UHF spectrum space. If there must be sharing of frequencies between satellites and ground services, it is necessary to know the current extent and planned expansion of the ground systems on a world-wide basis. This information should be compiled from records of the Federal Communications Commission, International Telecommunications Union, the operators of microwave systems, foreign government frequency boards, and publications of other domestic and foreign government agencies. The technical parameters such as frequency, bandwidth, power outputs, antenna patterns, receiver characteristics, modulation, and location of transmitter and receiver terminals must be assembled in a form suitable for the computation of spectrum and space usage of existing systems.

It is recommended that such information be compiled for those specific bands that might be employed for satellite communications and be published in a volume available to interested parties.

2) *Spectrum Usage Measurements Program:* While the spectrum usage chart recommended in 1) will be valuable in the over-all study of spectrum availability, one improperly assigned or incorrectly operated transmitting station could cause field strengths capable of interfering with satellite receiver reception. The presence of unusual or unknown interference can be checked prior to the actual test of a complex communication relay by the use of a small low-altitude receiving experiment.

It is recommended that consideration be given to conducting a test using a low-orbit satellite with simple receiving equipment scanning those bands that might be useful for satellite relays. Alternatively, a satellite-tracking station having a large dish (capable of high directivity in the 1000- to 3000-Mc bands) and a *smaller precision dish* (capable of high directivity in the 3000- to 10,000-Mc bands) may be modified to provide continuous frequency coverage from 1000 to 10,000 Mc. Then this directive tracking station may be used to monitor the stray signals impinging on one or more Echo-type balloon satellites. Still other methods of measuring the interference may be proposed. The received field strength and frequency information obtained can be transmitted to existing ground terminals via standard telemetering equipment. Such a direct measurement of the electromagnetic environment would permit the extrapolation of measured results to higher orbits, thus providing key data for systems designers.

3) *Use of Frequencies Above 10,000 Mc:* An important concern is determination of the highest frequency that may sensibly and economically be used. At present, the frequency region of approximately 2000 to 6000 Mc appears to be most favored, with serious interest extending up to approximately 10,000 Mc.

Extremely-low-noise antenna-pre-amplifier combinations operated at frequencies higher than 5000 Mc will at times experience a noticeable degradation in sensitivity arising from other than gaseous attenuation in the earth's lower atmosphere. This attenuation, caused by concentrations of solid particles (rain, snow, fog, and cloud) in the receiving-antenna main beam, will increase the effective receiver temperature above the value set by antenna-pre-amplifier noise plus O_2 and H_2O vapor sky noise.

If we assume that a satellite communication circuit will be designed to have a satisfactory performance for 99.99 per cent of the time, many low- and mid-latitude receiving-terminal locations must be designed to expect a receiving effective noise temperature of some 300°K, since this temperature—associated with particle attenuation—may be expected to obtain for at least 0.1 to 0.01 per cent of the time. In terms of receiving sensitivity alone, then, there is no preference for any frequency region above, roughly, 3000 to 5000 Mc.

The fundamental limitation to the extension of satellite communication techniques to frequencies very much higher than 5000 Mc appears rather to be that of increased path loss caused by radiowave attenuation in rainfall. This effect will become important at much lower frequencies than that of increased gaseous attenuation for mid- and lower-altitude terminal location; the latter is not expected to increase greatly until frequencies above approximately 15,000 Mc are employed.

To a first approximation, we may expect a surface small-area-rainfall rate of 5 mm per hour to occur 0.1 per cent of the time. If we make the further crude assumptions that 1) ray paths of interest do not come closer than 10° to the local horizon, 2) heavy precipitation does not extend to beyond 10 miles in altitude, and 3) at any instant, this rainfall intensity does not occur in the beam for any more than one quarter of the total path length, total attenuation figures of some 3 db at 10,000 Mc, 8 db at 15,000 Mc, and 20 db at 20,000 Mc are indicated. To the extent that these assumptions prove to be valid, the penalty at 10,000 Mc does not appear too grave, and even that at 15,000 Mc is not prohibitive. In any event, these higher frequencies could be considered useful for terminal locations where high-intensity rainfall is not expected to occur or, independent of location, when a somewhat restricted circuit performance is acceptable.

From all of this, an important conclusion is evident: We must have a comprehensive and quantitive knowledge of the values of path attenuation expected to be equaled or exceeded over paths traversing the earth's lower atmosphere as a function of the percentage of a year's time that such losses occur, and for various terminal locations and low (1°–10°) slant angles. With this knowledge, sensible engineering judgments can be made as to the value of these higher-frequency regions.

It is to be noted that in the case of a passive Echo-type repeater, (setting aside the problems of accurate antenna pointing and sphere design) the higher frequencies will provide a lower free-space transmission loss (proportional to f^2) between fixed-aperture antennas. Inclusion of this factor in the transmission equations tends to eliminate any over-all frequency dependence for the entire 1000- to 15,000-Mc region where a very high circuit reliability performance is mandatory.

An experimental program is recommended using a low-altitude satellite to receive two or three ground transmissions in the 10- to 30-kMc region measuring signal strength to establish signal degradations from weather. This can be measured by transmission from ground transmitters to satellite receivers, signal-strength measurements being retransmitted to the ground via standard telemetering—a relatively simple experiment.

4) *Interference Field-Strength Levels:* To a very good first approximation for frequencies below 6000 Mc, free-space path-loss conditions may be expected to obtain over line-of-sight paths for slant angles greater than a very few degrees above the local horizon. We may, therefore, have a high degree of confidence in predicting interference levels between satellite repeaters and surface stations. Essentially, all that is needed to make such calculations is a knowledge of the frequency and antenna characteristics.

This knowledge is not generally sufficient to calculate interference levels between surface terminals. Since very high values of effective radiated power and extremely sensitive receivers are expected to be employed at satellite circuit surface terminals, we must be concerned with fields propagated between various surface terminals that are well beyond the geometrical line-of-sight of each other. Consequently, we should have an accurate and comprehensive knowledge of the propagation of radiowaves over great distances beyond the geometrical horizon at frequencies between, roughly, 1000 to 20,000 Mc.

In general, such knowledge is lacking. As a result of tropospheric-scatter propagation studies, it is probably reasonable to expect that useful estimations can be made to distances of some 500 miles at frequencies below, roughly, 500 Mc and to some 200 to 300 miles at frequencies below, roughly, 3000 Mc. At frequencies above this, useful data are meager, and none are known to exist above 10 kMc. Because the investigations that have been made have tended to emphasize measurement of the lowest values of field strength (for reliable communications-circuit-design purposes), the highest (1 to 0.01 per cent) fields were often ignored or not measured accurately.

Before generally accurate estimations of SHF interference between surface terminals can be made, therefore, data of particular pertinence to the interference question must be obtained throughout this frequency region. Perhaps concentrating first on frequencies below 10,000 Mc, long-term measurements should be made of the fields expected to be equaled or exceeded 1 to 0.01 per cent of the time over various path lengths out to 100 or 200 miles in various meteorological locations. Particular attention should be paid to such path charactersitics as surface roughness, horizon angles, and the prevalence of super-standard index profiles, elevated layers, and ducts.

It would appear that the most serious interference potential involves the satellite surface terminals rather than the active high-altitude space repeater. To the extent, then, that the effective radiated power of the surface transmitter and the sensitivity of the surface receiver can be reduced, so will their potential for interfering with others and of being interfered with by others.

It is to be noted, in this regard, that an active satellite receiver effective temperature looking earthward as generally discussed today is in the neighborhood of 3000°K. An active satellite antenna, looking earthward, will "see" the radio temperature of the earth (approximately 300°K) surrounded by a narrow halo of gaseous absorption (approximately 10°K to 100°K) and the cold of space (approximately 0°K).

For antenna half-power beamwidths $\leqq 50°$, approximately, satellites at 5000 miles and lower will see essentially only the earth, and the receiver sensitivity will be fundamentally limited to approximately 300°K; greater altitudes or beamwidths or both will allow a greater sensitivity, since then the earth will not completely fill the antenna beam.

There appears to be, therefore, the possibility of increasing the active satellite-receiver sensitivity by a factor of 10, perhaps by somewhat more. To date, this probably has not seemed worthwhile in view of the additional complexity this would imply for receiving equipment in the satellite and the fact that technical means are well within hand of generating and directing adequate values of microwave effective radiated power (ERP) toward the satellite. Eventually, this factor of 10 might be important from the standpoint of reducing ground transmitter power to reduce the radius of possible surface-to-surface interference. In current designs, attempts are now being made to keep the surface receiver noise level very low, because of the small amount of ERP expected from the satellite and the very long line-of-sight path lengths inherently involved in satellite communications. The low value of satellite ERP results from great reluctance to incorporate directive antennas in a satellite—with the consequent problems of dynamic long-term antenna orientation—and the need to keep the microwave power level low (approximately 1 to 10 watts) in order both to achieve very long cathode life and not to require heavy satellite power supplies which would increase the launch cost.

As more powerful rockets become available, and as more experience is gained in placing many nearly identical satellite packages into similar orbits, it appears reasonable to expect that the cost per pound of placing satellites into orbit will decrease. This would encourage the use of greater ERP from the satellite, either in terms of average microwave power output, or in terms of more sophisticated antennas, or both. Perhaps some 5 to 10 times the presently planned output power might be achieved by placing some 5 to 10 lower-power final amplifiers in parallel, thereby not only increasing the power output, but conceivably increasing transmitter long-term reliability.

While caution must be exercised in putting forth these particular suggestions, the goals of increased ERP and receiver sensitivity in high-altitude active satellites appear to be worthwhile areas for study—study prompted particularly by the possibility of reducing the interference potential of the surface equipments by perhaps a factor of 10 to 100.

5) *Surface Terminal Antenna Design:* It appears that the surface transmitting and receiving antenna of a satellite repeater circuit, in order to minimize atmospheric influences on radiowave propagation and receiver sensitivity, will probably not be used at angles lower than 5° above the local horizon—perhaps not much lower than 10°. For most antenna aperture sizes and wavelengths now being seriously considered, main radiation pattern free-space, half-power beamwidths of less than $\frac{1}{2}°$ will be employed—more likely, the beamwidths will approximate $\frac{1}{4}°$ or less. Therefore, the center of radiation will be, even during the time of greatest depression angle, at least 10 beamwidths above the local horizon. Even the second-order sidelobes of a circular (parabolic) antenna will be at least 2 to 3 beamwidths, and perhaps as much as 10 beamwidths, above this local horizon.

Under these conditions, the energy directed toward the horizon from such a high-gain antenna will, for the most part, be radiated from the backlobes of the antenna.

At the present time, the general backlobe level for a sensible antenna design should be at least somewhat below isotropic. A careful antenna design can reduce the average backlobe to 10 db below isotropic; careful study might result in a design 20 db or more below isotropic. In the latter case, a 1-kw transmitter would radiate only 10 watts in the horizontal plane, *i.e.*, an amount of power comparable to that presently used on line-of-sight microwave circuits.

From symmetry considerations, it is just as important that the receiving antenna present an extremely low effective absorption area in the direction of the radio horizon, *i.e.*, in the direction of potentially interfering sources. Here then, is a fruitful area for further research: the marked reduction of backlobe radiation from very-high-gain antennas.

6) *Surface Site Screening:* Inasmuch as the surface antennas are expected to be directed upward a minimum of some 5° to 10° above the local horizon, it becomes possible to consider screening transmissions from (or toward) the antenna from any azimuth sector—even in the direction of the main beam—for angles close to the local horizon.

An appropriately designed and located diffraction "screen" near the surface terminal location could conceivably attenuate low-angle signals by several tens of decibels without affecting the antennas' mainlobe performance in any significant manner. Such a technique has heretofore been used to screen a line-of-sight microwave circuit receiver from the surface reflected radiation component and thus to avoid the fading caused by such a multipath signal. It would appear that the technique might be logically extended to gain a further marked advantage over interfering signals propagated between surface terminals at angles at or very near to the horizon.

7) *Computer Simulation of Ground Interference:* It is evident that complex interference-producing situations would arise with the juxtaposition of a ground-microwave complex and satellite terminals with high-power transmitters and sensitive receivers. Also, satellite systems, except the 24-hour type, require moving antennas, further complicating the determination of interference levels.

Large electronic computers have been used with considerable success to compute the unwanted interference levels in large communication networks. Simulation of the ground-microwave complex and the satellite terminals on a large digital computer for the purpose of determining interference levels is feasible. Such a program would permit the alteration of system parameters including locations, power output, antenna patterns, receiver sensitivity, etc., and the immediate determination of interference levels for each suggested alteration.

8) *Harmonic Radiation:* It has been pointed out that there is a significant amount of radiation of higher harmonics from present-day radars. The increased use of the microwave region by satellite systems, utilizing, in most cases, low signal levels at the ground receiver, implies that restrictions should be placed on this undesirable harmonic radiation.

Studies of harmonic radiation from the many systems using these bands will be required, so that attainable standards for new equipment and the modification of existing systems can be determined.

Acknowledgment

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A Low-Noise Microwave Quadrupole Amplifier*

A. ASHKIN†

Summary—**Experimental results on a low-noise microwave amplifier operating at 4137 Mc are described. At a gain of 19 db the measured double-channel noise figure is 0.79 db, corresponding to an effective temperature of 58°K ± 10°K. The effective beam temperature is 25°K ± 10°K. The noise bandwidth between points 3 db down from the optimum is 32 Mc. The tube operates at 10 volts and 27 μa and requires about 150 mw of pump power. The dynamic range of the amplifier is about 100 db.**

Introduction

The quadrupole amplifier is a fast cyclotron wave parametric amplifier.[1] Of the electron beam parametric amplifiers proposed,[1-3] it has thus far been the most successful. Recently Adler, Hrbek and Wade, in comments on a letter by Lea-Wilson, have reported a double-channel noise figure of slightly better than 1.0 db, with a beam temperature of 35°K for a quadrupole amplifier operating at 425 and 780 Mc.[4] This was an improvement over earlier results[1] and involved use of a virtual cathode in the gun region. Previously reported noise figure results for a quadrupole amplifier operating at microwave frequencies had been considerably higher.[5,6] This report describes a recently designed version of the microwave quadrupole amplifier which has given a noise figure of 0.79 db, or an effective temperature of 58°K ± 10°K with a beam temperature of 25°K ± 10°K. These new results were achieved without the use of a virtual cathode, and without any current interception on the gun anodes. A brief description of this new version of the microwave tube and some experimental results follows.

Description of the Tube

The experiments were performed on the basic quadupole tube, since the prime objective was to see how low a noise figure could be achieved at microwave frequencies. Such a basic structure consists of two identical two-pole Cuccia coupler cavities and a quadrupole pump cavity. Figs. 1 and 2 show a photograph of the tube and

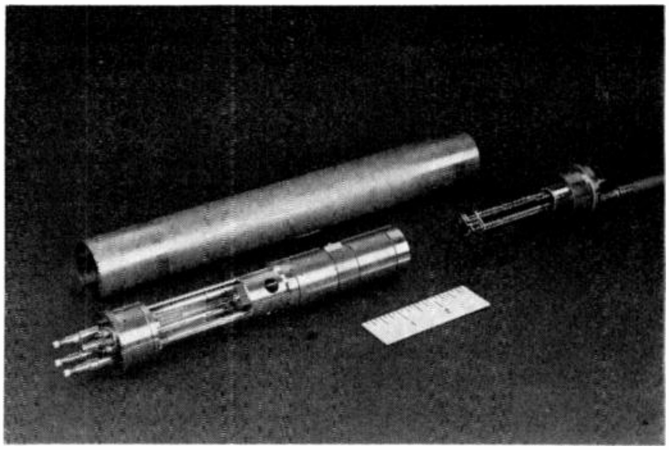

Fig. 1.

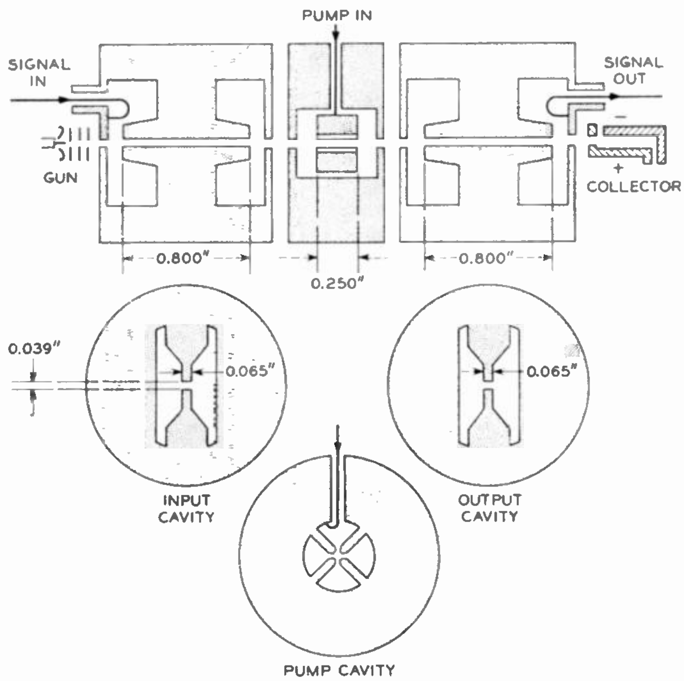

Fig. 2.

a sketch of its components. The tube differs from previous microwave quadrupole amplifiers[5,6] in the design of its signal and pump cavities, and also in several other features to be described.

The signal cavities have wider and longer vane faces (plates). The increase in width ensures that the beam will be in a region of uniform transverse field even when expanded by pumping.[5,7] The increased length reduces

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[1] R. Adler, G. Hrbek, and G. Wade, "The quadrupole amplifier, a low-noise parametric device," Proc. IRE, vol. 47, pp. 1713–1723; October, 1959.

[2] T. J. Bridges, "A parametric electron beam amplifier," Proc. IRE (Correspondence), vol. 46, pp. 494–495; February, 1958.

[3] W. H. Louisell and C. F. Quate, "Parametric amplification of space-charge waves," Proc. IRE, vol. 46, pp. 707–716; April, 1958.

[4] C. P. Lea-Wilson, R. Adler, G. Hrbek, and G. Wade, "Some possible causes of noise in Adler tubes," Proc. IRE (Correspondence), vol. 48, pp. 255–256; February, 1960.

[5] T. J. Bridges and A. Ashkin, "A microwave Adler tube," Proc. IRE (Correspondence), vol. 48, pp. 361–363; March, 1960.

[6] A. Ashkin, "A microwave Adler tube," *Proc. Internatl. Cong. on Microwave Tubes*, Munich, Germany; June 1960. (To be published.)

[7] R. Adler, A. Ashkin, and E. I. Gordon, "Excitation and amplification of cyclotron waves and thermal orbits in the presence of space charge," *J. Appl. Phys.*, vol. 32, pp. 672–675; April, 1961.

the coupling to synchronous wave noise by reducing the rate at which signal is applied or removed from the cyclotron wave.[8]

The spacing of the poles in the quadrupole pump cavity was increased, and the poles were reshaped to more nearly approximate a hyperbolic surface. The cavity was constructed using a hobbing technique which results in a high degree of symmetry and uniformity. All these changes assure increased purity of the hyperbolic fields in the region of beam interaction.[6] The pump cavity was also doubled in length to prevent the introduction of synchronous wave noise in the pumping process.[9]

Considerable care was taken with gun and cavity alignment. This feature is much more important in this type of tube than in traveling-wave tubes, since the tube naturally provides a built-in mechanism for amplifying any accidental beam spiraling. A sealed-off tube construction technique with all metal and ceramic materials was used, permitting high bakeout temperatures, and resulting in a cleaner tube with better ultimate vacuum. This manifested itself in improved cathode emission with lower cathode temperature. A precision solenoid with less than two gauss variation over the tube length greatly assisted in the proper testing of this tube. Tables I and II list some of the pertinent cavity dimensions and characteristics.

TABLE I

INPUT AND OUTPUT SIGNAL CAVITIES

1. Resonant Frequency	4137 Mc
2. Unloaded Q	2200
3. Plate Length	0.800 inch
4. Plate Width	0.065 inch
5. Plate Spacing	0.039 inch
6. Cavity Diameter	0.800 inch
7. Cavity Width	0.380 inch
8. Cavity Length	1.000 inch

TABLE II

PUMP CAVITY

1. Resonant Frequency	8274 Mc
2. Unloaded Q	1650
3. Plate Length	0.250 inch
4. Plate Spacing (along a diameter)	0.062 inch
5. Cavity Diameter	0.450 inch
6. Cavity Length	0.500 inch

A confined flow gun was used consisting of a 0.015-inch cathode, a beam forming electrode, and three anodes for controlling the potential profile between cathode and input signal cavity. The collector was similar to Adler, Hrbek, and Wade's, and consisted of two plates as shown in Figs. 1 and 2. A potential difference was applied between them which deflected the beam and provided an effective trap against returning secondaries.

[8] R. Kompfner and J. W. Kluver, private communication.
[9] J. W. Kluver, private communication.

EXPERIMENTAL RESULTS AND DISCUSSION

Fig. 3 shows a plot of the transmission loss through the tube as a function of the frequency with the pump off. The minimum loss of 1.9 db is made up of approximately 1.1-db loss from the two coaxial lines, and about 0.4-db loss in each of the signal cavities. The data were taken with the ratio of beam voltage to beam current arranged to give a match with the cavity and with the cyclotron frequency adjusted to equal the signal frequency. Aside from the circuit losses, the tube behaves as a perfect isolator.

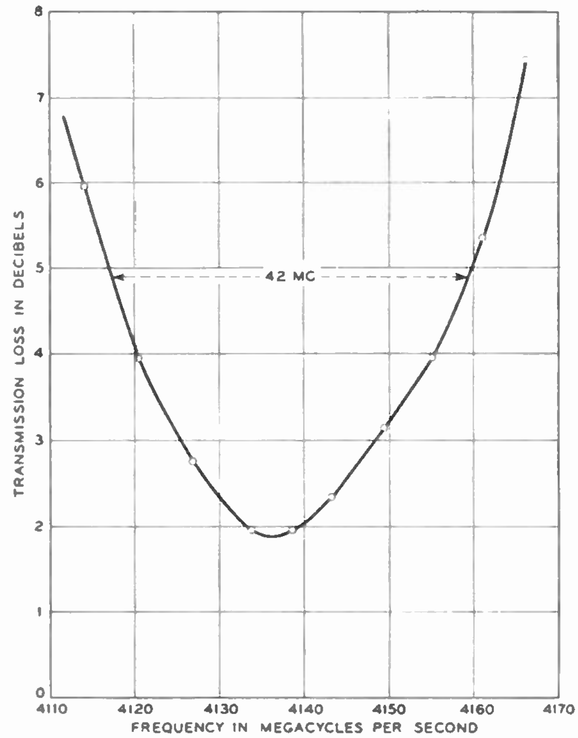

Fig. 3.

Turning on pump power gives signal gain as shown in Fig. 4. The signal gain in decibels theoretically should be linear when plotted against the square root of the pump power at high power, and should lie 6 db below a line of the same slope passing through the origin.[1,5] This is borne out quite well in practice.

The dynamic range of the amplifier can be determined by measuring the gain vs the input signal level for a fixed pump power (see Fig. 5). For the pump power corresponding to optimum noise figure, the gain is constant at about 19 db until the input power is approximately -18 dbm. At higher input power, beam interception occurs and the gain decreases. The maximum output power before beam interception is 1.4 mw. The thermal noise level in the 2-Mc receiver band used is -118 dbm at best noise figure. Thus the tube has a dynamic range of about 100 db.

Fig. 6 shows the measured variation in noise figure with frequency. The optimum noise figure of 0.79 db

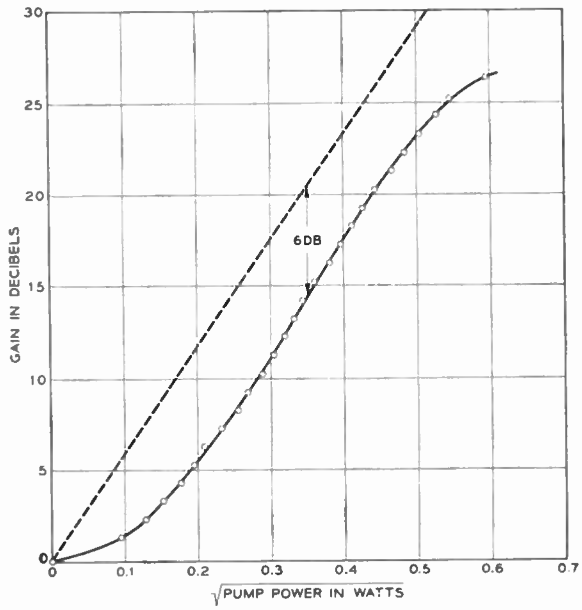

Fig. 4.

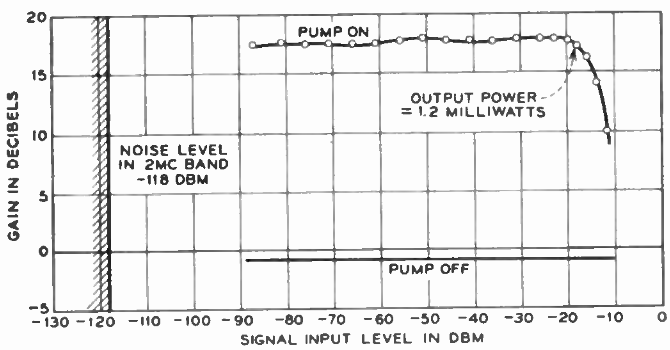

Fig. 5.

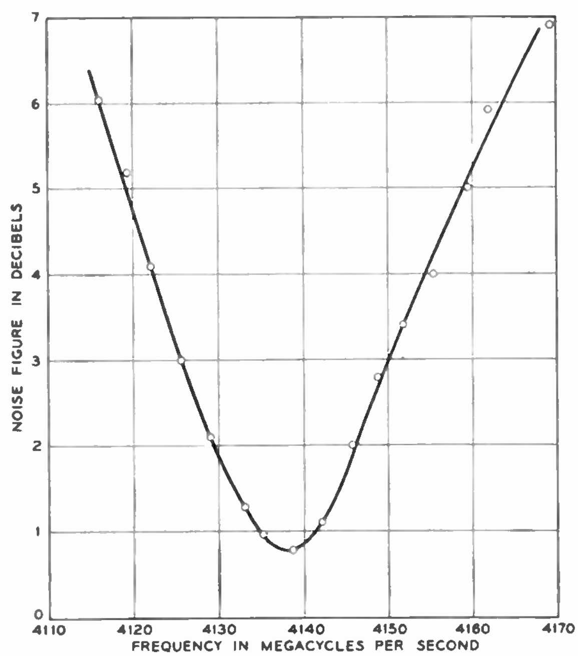

Fig. 6.

corresponds to an effective temperature of 58°K ± 10°K. The noise figure readings shown here include the input cavity loss, but not the input line loss. If the input cavity loss is subtracted, one obtains the effective temperature of the beam itself, which is 25°K ± 10°K. The effective temperature of the beam really lumps together all the residual uncompensated noise sources that remain on the beam after the input circuit loss is subtracted. Comparing Fig. 6 with Fig. 3, it is seen that the noise bandwidth is less than the signal bandwidth. This is to be expected since as one departs from the resonance frequency, not only does the transmission loss go up, but the noise stripping is no longer entirely effective. Under the conditions giving optimum noise figure, there is no current interception on any of the gun anodes. Also, the noise figure is found to depend only very slightly on the potential profile of the gun.

Fig. 7 shows the variation of noise figure with beam voltage for constant gain. The beam current and pump power are adjusted at each beam voltage to give the proper gain and optimum noise figure as shown in Figs. 8 and 9. The beam current variation is necessary in order to maintain constant beam impedance for proper noise stripping and minimum signal loss. The variation in pump power simply compensates for the variable electron drift time in the pump cavity. The observed straight line relationship between pump power and beam voltage for constant gain is to be expected since the gain is proportional to the ratio of the quadrupole field to the beam velocity. It is seen that above about 10 volts there is little improvement of the noise figure with beam voltage. The increase in the noise figure with decreasing beam voltage is probably associated with the increased percentage velocity spread of the beam at lower voltage. A velocity spread results in different gains for electrons with different axial velocities; this destroys the noise balance between different parts of the beam as established in the input coupler. Differences in axial velocity in conjunction with variations in the magnetic field between signal cavities can also introduce noise[10] which would increase at lower beam voltages.

With this tube, one can get a measure of a small part of the fast cyclotron wave noise spectrum by measuring the noise stripped off the beam by the input cavity. If the noise spectrum were smooth, one would expect the observed noise from the input cavity to start at room temperature, rise smoothly to a maximum corresponding to the true beam temperature at the center band frequency, and then fall smoothly again to room temperature. Fig. 10 shows the measured noise variation for the set of gun potentials indicated. An unexpected noise peak is seen superimposed on an otherwise smooth curve. Small changes in the gun electrode potentials drastically affect the noise peak. Changing anode 2 or 3 a few volts reduces the peak markedly. A change in the

[10] E. I. Gordon, private communication.

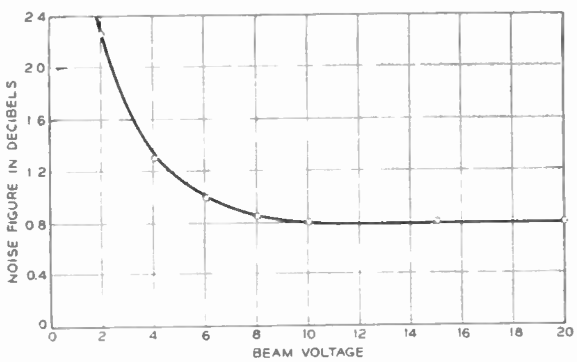

Fig. 7.

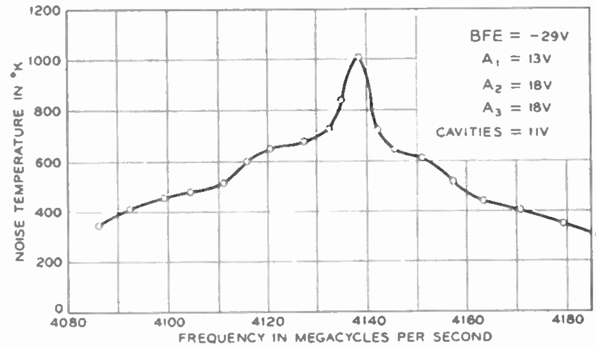

Fig. 10.

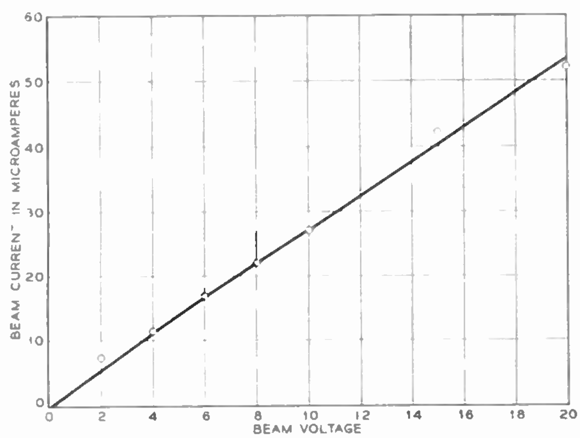

Fig. 8.

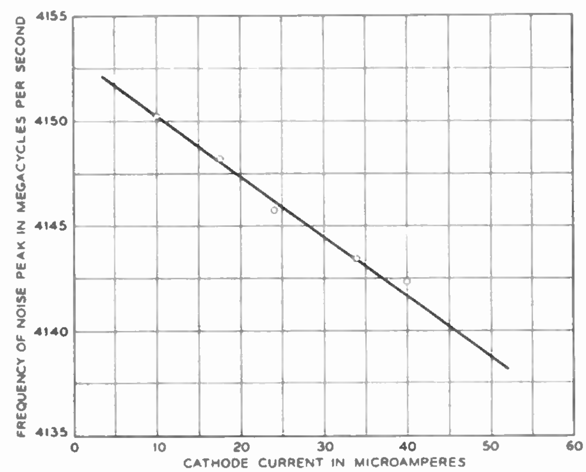

Fig. 11.

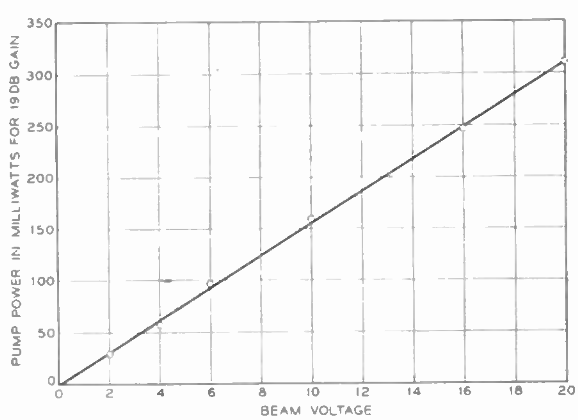

Fig. 9.

cyclotron frequency of a given number of megacycles causes a corresponding change in the position of the peak.

Although the details are not all clear, it seems that this noise peak is related to a phenomenon observed by Adler and Hrbek[11] in their low-frequency tube. They observed a signal transmission anomaly when studying transmission through their tube in the absence of pump power when the quadrupole was run at a potential higher than the neighboring couplers. They found absorption and gain, which they ascribed to slow electrons, and which were somehow scattered into the quadrupole section and trapped in the potential well. Their interaction was shifted from the cyclotron frequency by an amount which was presumably the space charge reduction in the rotational frequency of the trapped electron cloud. As evidence for a similar trapping effect in the gun region of this tube, Fig. 11 shows the variation in the frequency of the noise peak as the cathode current is varied by changing the heater power with all gun potentials fixed. According to the trapping supposition, electrons are trapped in the gun region due to the potential profile, and rotate at different frequencies depending on the beam space charge. They interact with the passing beam and serve to increase its fast cyclotron wave noise content at their rotational frequency. The slope of Fig. 11 is close to a straight line, which is the type of variation of rotational frequency with beam current one would expect on the assumption that the density of the space charge is uniform and proportional to current. The variations in the size of the noise peak with small changes in anode voltages may be due to variations in the noise content of the trapped cloud caused by changes in the transit time of electrons in the trapped region.

[11] R. Adler and G. Hrbek, private communication.

If one accepts the noise peak as due to induced noise, then the basic temperature of the fast cyclotron wave corresponds to the midband temperature in the absence of the peak, which is about 750°K. This is somewhat lower than cathode temperature, and possibly implies a little beam expansion.

The slight changes of the noise figure observed as the potential profile of the gun was varied are undoubtedly due to variations in the size and position of the noise peak on the incident cyclotron wave noise spectrum. The fact that fairly substantial variations in incident noise affect the measured noise figure so little is a tribute to the effectiveness of the input cavity noise stripping.

With the input cavity matched at room temperature, one can look at the noise being stripped off the output cavity. Assuming that no additional beam noise is being added, one expects to find room temperature again as the output noise temperature. The measurement yielded room temperature plus about 15°K. Thus, 15°K is the beam temperature at zero gain. Since the beam temperature at full gain is 25°K ± 10°K, it is seen that there may be some additional noise contributed to the beam in the pumping process. This point can be established by a precise measurement of the noise figure vs gain. Such measurements were attempted, but the accuracy at low values of gain was inadequate.

Conclusions

On the basis of the above measurements, it is concluded that thus far there is no evidence of any degradation of noise figure of quadrupole amplifiers with frequency up to 4137 Mc. The effective beam temperature of 25°K ± 10°K is comparable to that measured by Adler, Hrbek and Wade at 425 Mc and 780 Mc, and was achieved without use of a virtual cathode or beam shaving in the gun. The tube itself is likely to prove a useful tool for investigating the residual sources of noise in fast cyclotron wave amplifiers. On the basis of the present information, there is no evidence that the noise figure cannot be further improved.

Acknowledgment

The author wishes to express his appreciation to R. Kompfner and C. C. Cutler for their active interest and stimulation during the course of this work.

Thanks are also due to A. R. Strnad, who did all the mechanical design work in this tube, and thus contributed very tangibly to the success of these experiments.

Discussions with E. I. Gordon and J. W. Kluver were much appreciated.

Finally, the author wishes to thank J. J. Wiegand, who assisted so ably in every aspect of the experimental work

Correction

Edward E. David, Jr., author of "Digital Simulation in Research on Human Communication," which appeared on pages 319–329 of the January, 1961, Special Computer Issue of these Proceedings, has called to the attention of the Editor a confusion in the order of figures.

The photograph indicated as Fig. 2 should be Fig. 8(b). The photograph indicated as Fig. 8(a) should be Fig. 2. The photograph indicated as Fig. 8(b) should be Fig. 8(a). The figures are reproduced below as they should have appeared.

Fig. 2—Picture consisting of 100×100 points after passing through translator.

(a) (b)

Fig. 8—Comparison of picture codings. (a) 3 bits/picture point, differential quantizing. (b) 3 bits/picture point, pulse-code modulation.

Tapered Distributed RC Lines for Phase-Shift Oscillators*

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Summary—**Tapered transmission lines of distributed resistance and capacitance are useful in electronic oscillators because they permit a reduction in the required levels of impedance and gain. The analysis of such lines is also of value as a guide to estimating the performance of multisection lumped networks. An interesting feature of such lines is that the attenuation constant and characteristic impedance is different for forward and backward waves, and the voltage is shifted in phase and does not double in magnitude at the reflection from an open circuit. A set of normalized curves describing the behavior of such lines is presented as the principal result of this work.**

I. Introduction

Analyses by Sherr,[1] Sulzer,[2] and others show that in RC phase-shift networks for vacuum tube oscillators, the loss is greatly reduced by tapering the impedance and by increasing the number of sections from three to four or more. Extrapolating from these results, it is forseen that even lower losses could be secured by going to five, six, or even more sections, especially if impedance taper were employed. Consideration of the Nyquist diagram shows that no possibility of spurious oscillation exists unless the number of sections exceeds six because the limiting phase shift in each section is 90°. Moreover, because the voltage loss factor in a uniform three-section network producing a 180° phase shift is 29, it follows that the loss of a nine-section network producing a 540° shift is of the order of $29^3 = 24{,}389$. Results derived below show that spurious oscillation presents no real problem regardless of the number of sections.

Unfortunately, the analysis of multiple-section lumped networks is exceedingly complicated and tedious. Therefore, some alternative procedure is desirable. It is found that useful inferences may be drawn from the limiting case in which the number of sections becomes infinite and the network becomes a smoothly tapered transmission line. The properties of uniform transmission lines are well known, and the characteristics of tapered loss-free lines have been published.[3] However, a reasonably careful search failed to disclose any analysis applicable to the tapered RC transmission line. The following sections are devoted to such an analysis and its application to phase-shift oscillators.

* Received by the IRE, January 3, 1961.

† Electromagnetic Technology Corp., Palo Alto, Calif.

[1] S. Sherr, "Generalized equations for RC phase-shift oscillators," Proc. IRE, vol. 42, pp. 1169–1172; July, 1954.

[2] P. G. Sulzer, "The tapered phase shift oscillator," Proc. IRE, vol. 36, pp. 1302–1305; October, 1948.

[3] A. W. Gent and P. J. Wallis, "Impedance matching by tapered transmission lines," *J. IEE*, vol. 93, pt. IIIA, pp. 559–563; 1946.

II. Analysis of Infinite Line

To simplify the analysis, it is assumed that the transmission line is infinitely long and has an exponential variation of the parameters corresponding to

$$r = Re^{\pm 2kx} \quad \text{and} \quad c = Ce^{\mp 2kx}, \tag{1}$$

where r and c are respectively the resistance and capacitance per unit length, and the $\pm$ sign of k permits taper in either direction. Referring to Fig. 1, the differential

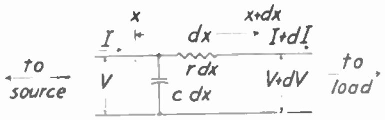

Fig. 1—Notation used.

equations corresponding to steady-state sinusoidal excitation are

$$dV/dx = -rI \quad \text{and} \quad dI/dx = -j\omega cV. \tag{2}$$

They combine to yield

$$\begin{aligned} d^2V/dx^2 &= -r(dI/dx) - I(dr/dx) \\ &= j\omega rcV \pm 2k(dV/dx). \end{aligned} \tag{3}$$

Assuming an exponential solution of the form

$$V = V_0 e^{\gamma x}, \tag{4}$$

and noting that $RC = rc$, one has by substitution

$$\gamma^2 \pm 2k\gamma - j\omega RC = 0. \tag{5}$$

Thus, the propagation constant is

$$\gamma \equiv \alpha + j\beta = \pm k + \sqrt{k^2 + j\omega RC}. \tag{6}$$

The physical situation is such that for small values of the taper parameter k, and for large values of ω, the attenuation constant α must be positive. Therefore, only the positive value of the radical has physical significance. However, as previously noted, k may be either positive or negative.

The analysis is simplified if we normalize the frequency variable to

$$m = \sqrt{\omega RC / 2k^2}. \tag{7}$$

Substitutions in (6) yield

$$\gamma = \alpha + j\beta = \pm k + \sqrt{k^2 + 2jk^2m^2}. \tag{8}$$

Transposing and squaring gives

$$\alpha^2 \mp 2k\alpha + k^2 - \beta^2 + j2\alpha\beta \mp j2k\beta = k^2 + j2k^2m^2. \tag{9}$$

Separating real and imaginary parts yields two equations:

$$\alpha^2 \mp 2k\alpha = \beta^2, \tag{10}$$

and

$$\alpha\beta \mp k\beta = k^2m^2. \tag{11}$$

It is relatively easy to evaluate m for assigned values of k, α, and β, but the converse process is much more difficult. Therefore, it is advantageous to prepare a plot relating α and β to k and m (see Fig. 2).

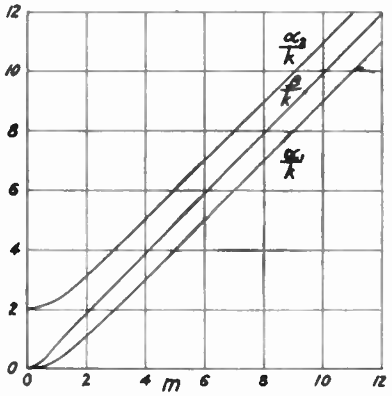

Fig. 2—Attenuation and phase shift.

It is found that the most natural representation involves the ratios α/k and β/k. Thus, the situation is comparable to that in loss-free tapered lines where the taper per wavelength is more significant than the taper per meter.

It is seen that the value of the phase constant is independent of the direction of propagation, but the attenuation constant is double valued, the lower attenuation corresponding to propagation in the direction of increasing impedance. Moreover, the ratio of phase shift to attenuation varies with m and increases without limit as m approaches zero.

The characteristic impedance, Z_0, of a transmission line is ordinarily of interest. However, the same information is provided by its reciprocal, the characteristic admittance, Y_0, which is used in the present case for simplicity. Substitution of (4) in (2) yields

$$Y_0 = G_0 + jB_0 = I/V = \gamma/r = \alpha/r + j\beta/r. \tag{12}$$

Therefore, the curves of Fig. 2 could also represent the characteristic admittance by addition of an appropriate ordinate scale.

It is seen that all the curves of Fig. 2 become straight and parallel for large values of m. This fact suggests that the data would be more useful if the ordinate were divided by the abscissa. The resulting design curves, presented in Fig. 3 (next page), are the principal result of the foregoing analysis. Emphasis is placed upon α_1, rather than α_2, because it is of greater practical interest and because α_2 is easily and accurately evaluated by the relationship

$$\alpha_2/k = 2 + \alpha_1/k.$$

III. Total Reflection

Assume a wave is propagated to the right in a finite line section which is open-circuited at the point $x=l$. The simple voltage doubling which is characteristic of smooth lines does not occur, and it is necessary to return to fundamental considerations. In terms of component voltages and currents, one has

$$I_1 + I_2 = 0 = V_1Y_{01} + V_2Y_{02}, \tag{13}$$

which with (12) gives

$$V_2/V_1 = Y_{01}/Y_{02} = \gamma_1/\gamma_2. \tag{14}$$

The total open-circuited voltage is related to V_1 by

$$V_t/V_1 = (V_1 + V_2)/V_1 = 1 + (\alpha_1 + j\beta)/(\alpha_2 + j\beta). \tag{15}$$

However, $\alpha_2 = \alpha_1 + 2k$. Therefore, we may convert this expression to

$$V_t/V_1 = 2(k + \alpha_1 + j\beta)/(2k + \alpha_1 + j\beta). \tag{16}$$

The magnitude and phase angle of this function are plotted in Fig. 3. By an odd coincidence, the magnitude of V_t/V_1 is closely equal to $1 + \alpha_1/km$.

The total resistance and capacitance of a finite line section are of interest. Assuming the negative sign for the exponential, one has for the total capacitance

$$C_t = \int_0^l c\,dx = C\int_0^l e^{-2kx}dx = (C/2k)(1 - e^{-2kl}). \tag{17}$$

Similarly, using the positive sign, the total resistance is

$$R_t = \int_0^l r\,dx = R\int_0^l e^{2kx}dx = (R/2k)(e^{2kl} - 1). \tag{18}$$

The product, which will be compared with that of lumped circuits, is

$$R_tC_t = (RC/4k^2)(e^{2kl} - 2 + e^{-2kl}), \tag{19}$$

and it is seen to be independent of the direction of taper, as indicated by the sign of k.

Introduction of a taper parameter,

$$p = e^{2kl}, \tag{20}$$

and substitution in (7), yields the useful relationship

$$R_tC_t = (m^2/2\omega)(p - 2 + 1/p). \tag{21}$$

IV. Application to an Oscillator

Let us assume that a line of length l is connected at its left end where $x=0$ to the plate of a triode having a dynamic plate resistance r_p and transconductance g_m. The problem is to determine line parameters which will result in stable oscillation at some specified angular frequency, ω. It is desirable to minimize the value of transconductance required for oscillation. However, the situation does not lend itself to a formal differentiation. Accordingly, the treatment is limited to the demonstration that the results obtained are more favorable than previous ones.

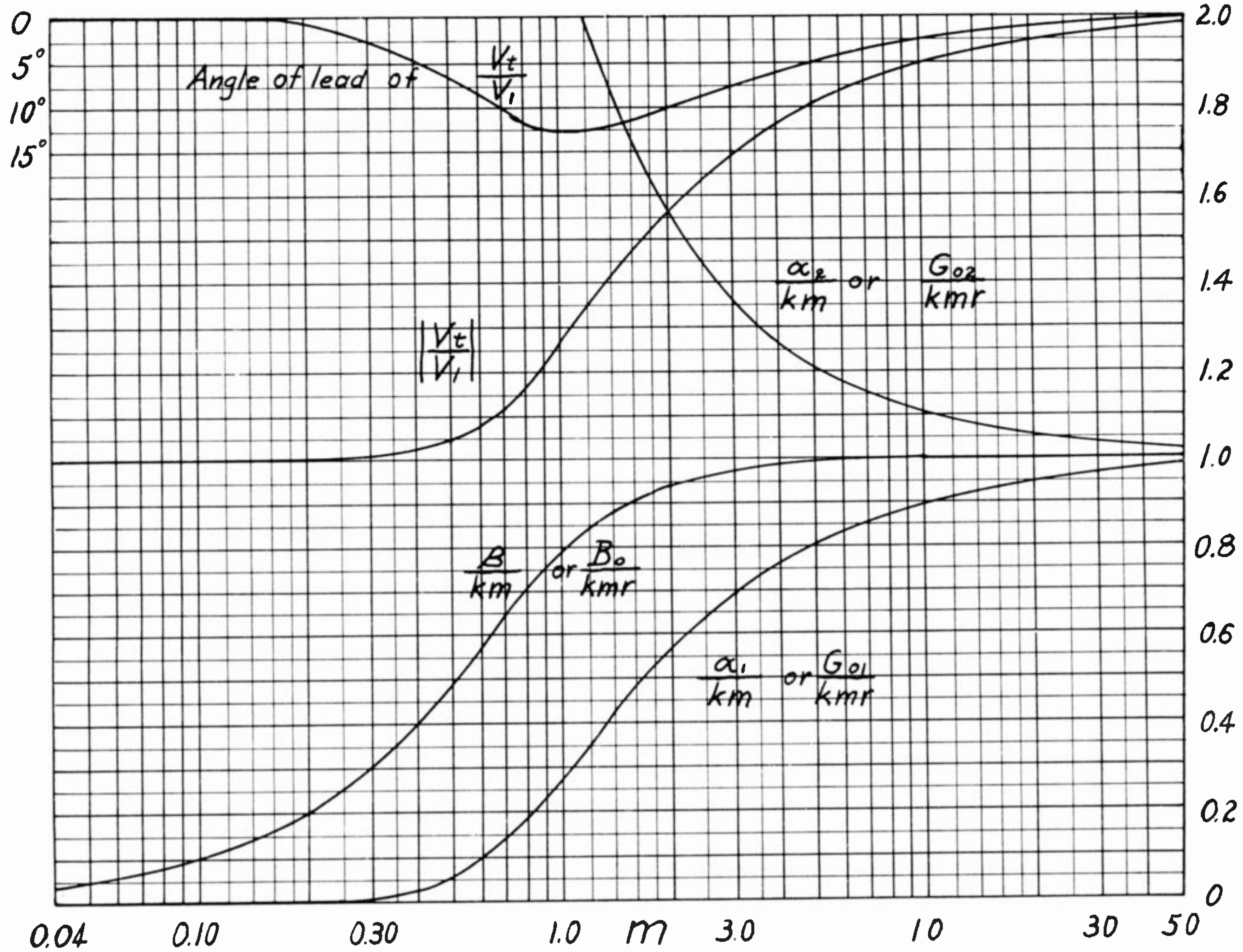

Fig. 3—Normalized parameters of a tapered line.

The basic conditions for oscillation (Barkhausen's criteria) are that the total loop phase shift is an integral multiple of 360°, and the total voltage gain is unity. Since a triode has an inherent phase shift of 180°, it follows that the line must also produce a phase shift of 180° or π radians. Because terminal phase shifts are relatively small, it follows that βl is approximately equal to π. From Fig. 3 it is seen that very small ratios of α to β are obtained by the use of small values of m. However, very small values of m are inappropriate because they lead to excessive values of impedance taper. To explore this situation, we introduce the approximation

$$\beta l \doteq 3 \tag{22}$$

which, with (20), yields the relationship

$$\beta/k \doteq 6/\ln p. \tag{23}$$

This expression is useful because it fixes the value of β/k, and hence of m, and thereby all the other parameters which correspond to a particular value of the overall impedance taper, p.

In very short sections of line, the input impedance is naturally sensitive to the impedance attached to the far end. However, in lines suitable for oscillators, the attenuation is high enough so that the input impedance is nearly independent of the termination. Attention is directed to Fig. 2, which shows that in all cases $\alpha_1+\alpha_2>2\beta$. Since $\beta l\doteq 3$ is a basic condition for oscillation, it follows that the corresponding two-way attenuation is of the order of six nepers, corresponding to a voltage ratio near 300.

V. Numerical Example

The application of the foregoing relationships will now be illustrated by carrying out the design of an oscillator using a triode having a plate resistance of 10^4 ohms. For a typical triode, the impedance ratio factor p may have a value as large as one hundred. Substitution of this value in (23) yields $\beta/k=1.30$. Reference to Fig. 2 fixes the value of the frequency variable at $m=1.41$.

An appropriate value for the line input admittance is $G_{01}=5\times10^{-5}$ (fifty micromhos). Referring to Fig. 3 at $m=1.41$, we have $B_{01}=(0.883/0.424)\cdot(5\times10^{-5})=108\times10^{-6}$. The total admittance at the plate terminal is the sum $(1/r_p+G_{01}+jB_{01})=(100+50+j108)\times10^{-6}=185\times10^{-6}$ at an angle 35.7° (lagging).

From Fig. 3, the phase angle at the grid end of the line is 12.1° leading. Thus, the total phase shift required in the line is

$$(180 + 12.1 - 35.7) = 154.4°, \text{ or } 2.70 \text{ radians.} \quad (24)$$

Therefore, the line length must be such that

$$\beta l = 2.70 \text{ and } \alpha l = (0.424/0.883) \times 2.70 = 1.30 \text{ nepers.} \quad (25)$$

Thus, the voltage loss within the line is

$$e^{1.30} = 3.68. \quad (26)$$

However, this loss is partially compensated by the voltage step-up at the open-circuit, which for $m = 1.41$ has the value 1.42, so that the net voltage loss of the total line is only $3.68/1.42 = 2.59$.

A voltage V_t at the grid terminal produces at the plate a driving current equal to $g_m V_t$ and a plate voltage having magnitude equal to $g_m V_t/Y_t$. The loop-gain condition requires $2.59 = g_m/(185 \times 10^{-6})$ or $g_m = 480$ micromhos. To guarantee oscillation in the presence of unavoidable degradation of tube parameters, a larger transconductance such as 1000 micromhos is desirable.

Use of the corrected value of phase shift gives, for the final impedance taper,

$$kl = 2.08 \quad \text{and} \quad p = 65. \quad (27)$$

Introducing these values in (13) gives

$$R_t C_t = 15.75 RC/k^2 = 63/\omega. \quad (28)$$

The value of the $R_t C_t$ product is abnormally large because of the relatively large value of impedance taper chosen. However, the required value of voltage amplification is exceptionally low.

The impedance presented to the grid is obtained by the use of the corrected taper factor, p. The susceptance value is 108×10^6 divided by 65 or 1.66 micromhos. The conductance is enhanced by the fact that propagation is in the reverse direction. Using Fig. 2 to obtain the ratio G_{02}/G_{01}, we have as the total conductance

$$(50 \times 10^6 \times 2.6)\ (0.6 \times 65) \text{ or } 3.3 \text{ micromhos.} \quad (29)$$

This value corresponds to a very reasonable impedance level of about one-fourth megohm.

Submillimeter Wave Radiometry*

M. W. LONG†, SENIOR MEMBER, IRE, AND W. K. RIVERS, JR.†

Summary—**The sensitivities of square-law detectors and superheterodyne receivers presently available at the shorter millimeter wavelengths are reviewed and extrapolated to the submillimeter region. It is concluded that, for applications which permit wide reception bandwidths, significantly more sensitivity will result from use of a direct-detection system consisting of a chopper, square-law detector and narrow-band amplifier than will be attainab e with superheterodynes in the foreseeable future. The expected sensitivity of a direct-detection system at wavelengths of the order of 1 mm is compared with that of the most sensitive centimeter radiometers.**

I. Introduction

Microwave researchers are interested in extending the spectral region in which they can make observations to include wavelengths which are presently unused. Such an extension is limited by weakness of generators and lack of sufficiently sensitive detectors. Sophisticated radiometry receivers such as the superheterodyne, traveling-wave tube or maser-amplifier types have not yet been realized for the wavelength region between 1 mm and 0.1 mm. A direct-detection system consisting of a chopper, square-law detector, and audio amplifier, however, is sufficiently sensitive in the submillimeter region to be of use in radiometry. Such a system is practical because wide bandwidths are available at the high operating frequencies of the submillimeter wave region. The direct-detection system is less complex than the superheterodyne or the tuned RF receiver; consequently, it is more easily transportable in satellites and other vehicles. The very narrow radiation patterns available from submillimeter antennas result in high antenna temperatures for radio sources which subtend small angles, such as planets; therefore, less receiver sensitivity is required for a given aperture size at the shorter wavelengths.

II. Direct-Detection Radiometer

Noise power from a matched microwave load and its temperature can be related by the approximation[1]

$$P = kTB,$$

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[1] This approximation is not usually valid for infrared; it causes only a small error in calculations for the submillimeter region providing low temperatures are not involved (see either Nicoll[2] or Harris[4]). For 0.1-mm wavelength, the maximum error in ΔP calculated [Eq. (1)] for switching between 300°K and any other temperature is approximately 1 db.

where k is Boltzmann's constant (1.38×10^{-23} joules per degree Kelvin), T is the temperature of the load in degrees Kelvin, and P is the power in watts available in the bandwidth B in cycles per second. If the input to a detector is switched between two loads at different temperatures, as is conventional in radiometry,[2-4] the change in available power is

$$\Delta P = 1.38 \times 10^{-23}\Delta TB, \tag{1}$$

where ΔT is the difference in the temperatures of the two loads. If the minimum detectable change in power, $\Delta P_{\min}$, is known for a given detector-amplifier combination, the minimum detectable temperature change, $\Delta T_{\min}$, can be determined from

$$\Delta T_{\min} = \frac{\Delta P_{\min}}{1.38 \times 10^{-23}B}. \tag{2}$$

In (2), B represents an equivalent bandwidth over which the detector can be assumed to be matched. Because the detector is a square-law device, $\Delta P_{\min}$ varies as

$$\Delta P_{\min} = K\sqrt{B_v}, \tag{3}$$

where K is defined as the minimum detectable change in power referred to a 1-cps video bandwidth, and B_v is the video bandwidth. Expected values of K are discussed in Section III. When combined, (2) and (3) result in

$$\Delta T_{\min} = \frac{7.3 \times 10^{22}K\sqrt{B_v}}{B}. \tag{4}$$

Data on commercially available centimeter-wave detector mounts suggest that B may be larger than one half the operating frequency. At submillimeter wavelengths the equivalent bandwidth will probably be limited due to matching difficulties, so it will be assumed that

$$B = f/3. \tag{5}$$

The result of substituting (5) into (4) is

$$\Delta T_{\min} = \frac{2.2 \times 10^{23}K\sqrt{B_v}}{f}. \tag{6}$$

For $K = 10^{-10}$ watts cps$^{-1/2}$, $B_v = 1$ cps and a wavelength of 1 mm ($f = 3\times 10^{11}$ cps),

$$\Delta T_{\min} = 73° \text{ K}.$$

Although the sensitivities of square-law detectors are poor, sensitivities available with direct-detection systems are practical for submillimeter wavelengths because of the very wide bandwidths available. For example, several investigators[5-9] have used the Golay cell as the detector for such a system to measure the equivalent black-body radiation of the sun and the moon, and measurements have been made to obtain a better understanding of total attenuation through the atmosphere.

III. Square-Law Detectors

Kaufman[10] gives a good general description of detectors for the region between microwaves and infrared. The crystal diode, the barretter, and the Golay cell are practical devices for a direct-detection system. Little quantitative sensitivity data are available for this region because good measurement techniques have not as yet been developed.

The crystal diode has a very short time constant that gives it an advantage over the barretter and Golay cell, which are heat detectors. As early as 1953, Gordy[11] and others at Duke University used a crystal diode in a video system to investigate absorption by the molecule OCS at 0.77 mm, but insufficient data were given to compute sensitivity. Richardson and Riley[12] have measured the 3-mm sensitivity of silicon wafers extracted from commercial crystal cartridges. For the best crystals, they suggest a sensitivity of 5×10^{-12} watt for a cycle of amplifier bandwidth and claim that the sensitivity has not suffered greatly by working in the millimeter region.

The Golay[13] cell is an instrument which detects pressure changes in a small volume of gas that is heated by incident radiation. This detector is reported to have constant sensitivity throughout the infrared region and well into the millimeter region. For example, the Eppley Laboratory, Inc., reports[14] a sensitivity for a commercially available Golay cell-amplifier combination of 6×10^{-11} watt for a time constant of 4 seconds for wavelengths as long as 4 mm. By assuming that the relation-

[2] G. R. Nicoll, "The measurement of thermal and similar radiations at millimetre wavelengths," *Proc. IEE*, vol. 104, pp. 519–527; September, 1957.

[3] F. D. Drake and H. I. Ewen, "A broad-band microwave source comparison radiometer for advanced research in radio astronomy," Proc. IRE, vol. 46, pp. 51–60; January, 1958.

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[5] W. M. Sinton, "Observations of solar and lunar radiation at 1.5 millimeters," *J. Opt. Soc. Am.*, vol. 45, pp. 975–979; November, 1955.

[6] W. M. Sinton, "Observation of a lunar eclipse at 1.5 mm." *Astrophys. J.*, vol. 123, pp. 325–330; March, 1956.

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[8] H. H. Theissing and P. J. Caplan, "Atmospheric attenuation of solar millimeter wave radiation," *J. Appl. Phys.*, vol. 27, pp. 538–543; May, 1956.

[9] H. A. Gebbie, "Detection of submillimeter solar radiation," *Phys. Rev.*, vol. 107, pp. 1194–1195; August, 1957.

[10] I. Kaufman, "The band between microwave and infrared regions," Proc. IRE, vol. 47, pp. 381–396; March, 1959.

[11] W. Gordy, "Millimeter and submillimeter waves in physics," *Proc. of the Symp. on Millimeter Waves*, New York, March 31, 1959. Polytechnic Press of the Polytechnic Institute of Brooklyn, New York, N. Y.; 1960.

[12] J. M. Richardson and R. B. Riley, "Performance of three-millimeter harmonic generators and crystal detectors," IRE Trans. on Microwave Theory and Techniques, vol. MTT-5, pp. 131–135, April, 1957.

[13] M. J. E. Golay, "Bridges across the infrared-radio gap," Proc. IRE, vol. 40, pp. 1161–1165; October, 1952.

[14] "Golay infra-red detector," The Eppley Laboratory, Inc., Bull. No. 10, revised April, 1959; private communications with Mr. A. R. Dennett of The Eppley Laboratory, Inc.

ship[15] between the filter time constant, τ, of a phase sensitive detector and noise bandwidth is $B_v = 1/\pi\tau$, the sensitivity of this Golay cell-amplifier combination corresponds to about 2×10^{-10} watt per cycle per second of amplifier bandwidth. The cell itself has a 15-msec response time.

Bolometers have been used for some time to detect weak infrared signals,[16] and Rohrbaugh[17] has reported the use of a fine-wire bolometer (barretter) at 1.4 mm. A minimum detectable X-band signal of 4×10^{-10} watt per cycle per second of amplifier bandwidth has been reported.[18] These measurements were made with a Narda-type 610B barretter and with a modulation rate of 1000 cps. More recent measurements, at the Georgia Institute of Technology, with a Narda 610B barretter evacuated to a pressure of about 20 microns resulted in a sensitivity of 5×10^{-11} watt per cycle of amplifier bandwidth for a modulation rate of 30 cps. Improved barretter performance may possibly be attainable by operating at the low temperatures[19] of superconducting transitions. Sensitivities comparable to our X-band results are reported for infrared bolometers; hence, it seems plausible that bolometers which are at least as sensitive as $10^{-10}\sqrt{B_v}$ watt can be used to span the region between microwaves and infrared. Our only measurements made at shorter wavelengths to this date have been at 4 mm with a PRD 634 mounted in an evacuated PRD 632 holder. Minimum detectable power was not measured because our main interest was in the minimum detectable temperature. Measurements using a noise source[20] indicate a minimum detectable temperature, $\Delta T_{\min}$, of 130°K for a 30-cps chopping frequency and an amplifier bandwidth of 1 cps. The use of this value for $\Delta T_{\min}$ in (6) would indicate a minimum detectable power, K, of 4×10^{-11} watt.

IV. The Dicke Radiometer

The sensitivity of a superheterodyne employing the Dicke RF switching scheme may be approximated[21] as

$$\Delta T_{\min} = \sqrt{2}T_e \sqrt{\frac{B_v}{B}}.$$

where T_e is the equivalent receiver noise temperature, B is the IF bandwidth, and B_v is the post-detection bandwidth.

As far as is known by the authors, 2 mm is the shortest wavelength for which superheterodyne sensitivity has been published. Some years ago Johnson reported[22] a sensitivity in the 2 to 3 mm region corresponding to an equivalent temperature of about 10^6°K, but it is likely that better sensitivities have been obtained since then. Available local oscillator power is low, and therefore submillimeter superheterodyne performance in the near future is expected to be very poor. Even with sufficient local oscillator power for optimum sensitivity, sensitivities available from crystal mixers developed for the submillimeter region are expected[23] to be poor compared with those for the centimeter region. For speculating on differences in superheterodyne and direct-detection performance, it is assumed here that the 10^6°K equivalent temperature that Johnson obtained several years ago in the 2- to 3-mm region can now be obtained for wavelengths less than 1 mm. Then for an IF bandwidth of 10 Mc and a video bandwidth of 1 cps, the minimum detectable temperature would be

$$\Delta T_{\min} = 450°\mathrm{K}.$$

A comparison of this result with that calculated for the 1-mm direct-detection system indicates a somewhat better sensitivity for the simpler system. If one were willing to increase the IF bandwidth to the order of 1 kMc by adding traveling-wave amplifiers, at the expense of more complexity and a higher noise figure, the sensitivity of the superheterodyne could be improved so that it would approximate that of the direct-detection system at this wavelength.

V. Comparisons Between Submillimeter and Centimeter Radiometers

The radiometer having the smallest minimum detectable temperature for a specified integration time employs a superheterodyne with a maser preamplifier for operation at 3 cm. For this instrument, an rms fluctuation level of about 0.04°K has been attained[21] by using an integration time constant of 5 seconds. Within the authors' knowledge, the most sensitive microwave radiometer was developed by Drake and Ewen.[3] This radiometer operates at 8 kMc; and, with an integration time constant of 320 seconds, it can detect an antenna temperature of 0.01°K. Drake and Ewen minimized deleterious effects of gain fluctuations by a new technique that results in some degradation of receiver-noise temperature. The effects of gain fluctuations on the sensitivity of a direct-detection system are also small because equivalent receiver-noise temperatures are large.

In many microwave radiometry applications, the solid angle subtended by the source is smaller than the antenna beam. For these applications, the temperature

[15] R. L. Cosgriff, "A study of Detectors and Amplifiers Used in Antenna Instrumentation," Antenna Lab., Ohio State University (Res. Foundation), Columbus, Tech. Rept. 487-5; December, 1953.

[16] R. A. Smith, F. E. Jones, and R. P. Chasmar, "The Detection of Infrared Radiation," Oxford University Press, London, Eng., sects. 3.3.3, 3.4, 7.2; 1957.

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[18] M. W. Long, "Detectors for microwave spectrometers," *Rev. Sci. Instr.*, vol. 31, pp. 1286–1289; December, 1960.

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[20] Noise Source Model GNW-X18, for which the noise temperature was given as 14,500°K ± 1800°K in private communications with R. White, Roger White Electron Devices, Inc.

[21] J. A. Giordmaine, *et al.*, "A maser amplifier for radio astronomy at X-band," Proc. IRE, vol. 47, pp. 1062–1069; June, 1959.

[22] C. M. Johnson, "Superheterodyne receiver for the 100 to 150 kMc region," IRE Trans. on Microwave Theory and Techniques, vol. MTT-2, pp. 27–32; September, 1954.

[23] C. T. McCoy, "Present and future capabilities of microwave crystal receivers," Proc. IRE, vol. 46, pp. 61–66; January, 1958.

which is measured by the receiver is proportional to antenna gain, which, for a fixed aperture size, increases as the square of the operating frequency. It is instructive to compare the potential of a direct-detection radiometer with the high performance X-band radiometers. A 1-mm direct-detection radiometer operating with a 5-second or a 320-second time constant could detect antenna temperatures of 18°K or 2.3°K, respectively. However, for a given aperture size, antenna gain will be 30 db higher at 1 mm than at X band; thus, the direct-detection system could, in the absence of atmospheric attenuation, detect source temperatures which are less than those detectable with the best microwave radiometers presently available. The authors believe that submillimeter-wave antennas can be developed which would be sufficiently large to justify this comparison in spite of the stringent requirements on surface accuracies and dimensional stability.

The above examples, which neglect atmospheric attenuation, indicate greater system sensitivity for 1-mm direct-detection systems than for highly sensitive X-band systems. These examples are valid only for those thermal sources, such as planets, subtending angles smaller than the submillimeter antenna beamwidth; however, some day submillimeter antennas may be developed which have beamwidths smaller than the angles subtended by some of the planets.

VI. Atmospheric Attenuation

Effects of atmospheric absorption on the capabilities of ground-based submillimeter-wave radiometers are appreciable. For example, the total attenuation[24] along the zenith at a wavelength of 1 mm is expected, depending on humidity, to lie between 3 db and 12 db. Some advantage may be gained by operating at 1.3 mm, where there seems to be a relative minimum of atmospheric absorption; results of previous research[8] indicate that the total attenuation for humid conditions should not exceed 5 db in this window.

There seems to be little submillimeter spectrum available where atmospheric attenuation is sufficiently low to allow detection of thermal sources outside the atmosphere by a ground-based radiometer. Iaroslavskii and Stanevich report[25] that humid air causes attenuations greater than 100 db/km for most of the frequency interval between 3×10^{11} cps (1 mm) and 3×10^{12} cps (0.1 mm). Results of several spectroscopic investigations[9,25,26] indicate a relative minimum in the atmospheric absorption at 0.35 mm (860 kMc). Iaroslavskii and Stanevich are the only investigators who have attempted to give quantitative data on the absorption. However, the attenuation they obtained in the laboratory over a 7.5-meter path was small, and consequently their measurements were subject to large fractional errors in regions of low absorption. Thus, little can be surmised about the numerical value of total atmospheric attenuation for the relative minimum. The authors have planned experiments to make total atmospheric measurements with a ground-based direct-detection radiometer by using the sun as the source for 0.35-mm waves.

Strong[27] has described a balloon-borne telescope to reduce effects of attenuation. In a window, the total atmospheric attenuation along the zenith is approximately equal to the attenuation per kilometer caused by the atmosphere near sea level; therefore, it seems that a substantial reduction in attenuation would result by operating from a mountain top, as was done by Gebbie.[9]

VII. Discussion

The shorter millimeter wavelengths and the submillimeter wavelengths lie in a virtually unexplored region of the electromagnetic spectrum. Difficulties associated with these wavelengths include high atmospheric attenuation, increased waveguide losses, and increased aperture-phase errors because of dimensional errors. These wavelengths offer the potentiality of developing very narrow-beam, high-gain antennas of reasonable size.

The direct-detection radiometer, which at present provides more sensitivity for submillimeter waves than does the more complex superheterodyne, has been used to study atmospheric attenuation and determine equivalent black-body temperatures of the sun and the moon.

The very narrow beamwidths available from submillimeter antennas result in high antenna temperatures for radio sources which subtend small angles, such as planets; therefore, less receiver sensitivity is required for a given aperture size at the shorter wavelengths. Because of the large antenna gains available, existing submillimeter technology permits the capability of detecting source temperatures comparable to those detectable with the most sensitive centimeter systems. However, care must be used to insure that maximum advantage is taken of available windows.

Further developments on frequency multipliers, millimeter-wave klystrons and magnetrons, parametric amplifiers, and the sophisticated new devices, such as masers and tunnel diodes, may make the submillimeter region attractive for communications systems. In the meanwhile, valuable information on atmospheric attenuation, temperatures of thermal sources, absorptivity of materials, and RF component design can be obtained from measurements with direct-detection submillimeter radiometers.

[24] E. S. Rosenblum, "Atmospheric absorption of 10–400 kMcps radiation: summary and bibliography to 1961," *Microwave J.*, vol. 4, pp. 91–96; March, 1961.

[25] N. G. Iaroslavskii and A. E. Stanevich, "Rotational spectrum of water vapor and the absorption of humid air in the 40-2500μ wavelength region," *Optics and Spectroscopy*, vol. 6, pp. 521–522; June, 1959. (Translated by Am. Opt. Soc.)

[26] H. Happ, W. Eckhardt, L. Genzel, G. Speiling, and R. Weber, "The crystal detector as a receiver of thermal radiation in the wavelength region of 100–1000μ," *Z. Naturforsch.*, vol. 12-A, pp. 522–524; June, 1957.

[27] J. Strong, "Interferometry for the far infrared," *J. Opt. Soc. Am.*, vol. 47, pp. 354–357; May, 1957.

Results of a Long-Range Clock Synchronization Experiment*

F. H. REDER†, ASSOCIATE MEMBER, IRE, M. R. WINKLER†, MEMBER, IRE AND C. BICKART†

Summary—**The USASRDL is conducting a feasibility study of World Wide Synchronization of Atomic Clocks (WOSAC) with an error of less than 5 μsec (1-μsec target). Original synchronization is established by flying an atomic clock from the master to the slave clocks. Synchronization is maintained by phase tracking a VLF signal controlled by the master clock. This paper gives the results of various phases of the project. Influences of the propagation medium allow VLF frequency transfer over the Atlantic with an accuracy of $2p:10^{11}$ for a 24 hour averaging period. A synchronization experiment over 1000 miles gave an accuracy of 0.2 μsec. Further improvements by a factor of 5 of LF frequency transfer over 300 miles distances was accomplished by phase control of the radiated signal by an Atomichron.**

The U. S. Army Signal Research and Development Laboratory at Fort Monmouth, N. J., is undertaking a feasibility study of world wide synchronization of atomic clocks (Project WOSAC), in close cooperation with Prof. J. A. Pierce of Harvard University, the Rome Air Development Center of USAF, the U. S. Navy Electronics Laboratory, and the British Post Office. Target is the demonstration of clock synchronization to 1 μsec and better over distances of several 1000 km. This is to be accomplished first, by synchronizing an atomic clock mounted in an airplane with the master clock, second, by synchronizing all slave clocks with the flying clock, and third, by maintaining synchronization of the slave clocks through phase tracking of VLF transmissions controlled by the master clock. Details of WOSAC are described elsewhere.[1]

The project falls into various phases. Between last August and October, a phase-monitoring station for the VLF station GBR (16 kc) was established temporarily at Banbury, England, and the phase of the GBR signal was measured simultaneously with identical equipment at Banbury and Cambridge, Mass., as shown in Fig. 1. This way, influences of the propagation medium could be separated from perturbing influences originating in the transmitter facilities. It was shown that VLF frequencies could be transferred over several thousand km's with a standard deviation of $2p:10^{11}$ for 24-hour averaging periods.[2] It is to be noted that the GBR propagation path to Cambridge, Mass., constitutes the worst condition as a result of its east/west direction because the diurnal phase variations during sunset and sunrise render substantial portions of the reception time useless for this purpose.

The results of the first flight tests of atomic clocks of the Cs-beam type (Atomichron), carried out in November, 1959, were reported recently.[3]

It is the purpose of this paper, to report briefly the results of further flight tests and of our preliminary high precision synchronization test over a distance of 1600 km, and to present new data on frequency transfer via LF and VLF propagation from transmitters directly controlled by Atomichrons.

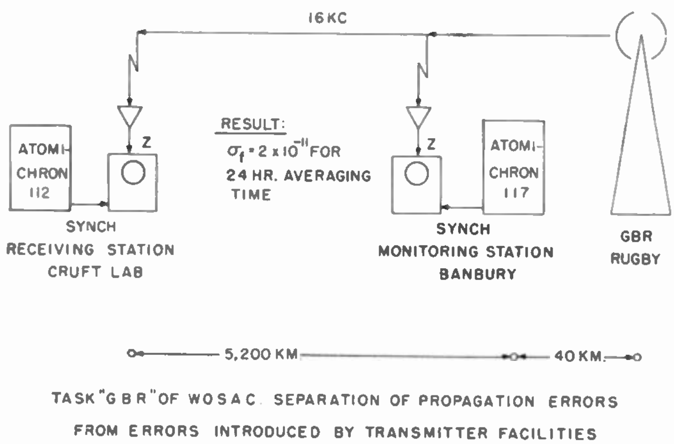


Fig. 1—Principle set-up for measuring the influences of the propagation medium on the phase of the 16-kc GBR transmission signal.

A $1\frac{1}{2}$-hour round trip around Fort Monmouth, similar to those conducted last November,[3] was carried out in good flying weather on February 9, 1960. Comparisons of the plane clock, 119, with the ground clocks, 118 and 115, over a 10-km long microwave link before and after the flight showed that the plane clock advanced by only 60 nanosec, which corresponds to an average frequency offset during the flight of $1.1p:10^{11}$. Thereafter, two more round trips around Fort Monmouth were carried out in rather turbulent weather. As a result, the Atomichron in the plane fell out of lock once in each case, and

* Received by the IRE, August 15, 1960.
† USASRDL, Fort Monmouth, N. J.

[1] F. H. Reder and G. M. Winkler, "World-wide clock synchronization," IRE Trans. on Military Electronics, vol. MIL-4, pp. 366–376; April–July, 1960.

[2] J. A. Pierce, G. M. Winkler, and L. Corke, "The GBR experiment," *Nature*, vol. 187, pp. 914–916; September 10, 1960.

[3] F. H. Reder and G. M. Winkler, "Preliminary flight tests of an atomic clock," *Nature*, vol. 186, pp. 592–593; May 21, 1960.

each time it took a few minutes to lock it again. During the 2-hour flight on February 26, the plane clock lost 20.8 μsec while it was out of lock for about 10 minutes. During the $1\frac{1}{2}$-hour flight on March 1, it gained 6.0 μsec while it was out of lock for about 4 minutes. These two flights gave three important results.

1) The Atomichron, if properly aligned and mounted parallel to the fuselage axis,[3] operates much better than expected, even in moderately turbulent flying weather. The roll of the airplane is the most frequently encountered motion. Its effect is diminished by the special mounting of the Atomichron. Pitch would be dangerous, but it is not experienced as often as yaw which has, therefore, the most perturbing influences (coriolis forces) during passage through severe turbulences. Accelerations caused by temporary falls in air pockets have no ill effect.

2) Usually when a yaw put the Atomichron out of lock, the standard ended up locked to one of the side peaks of the Ramsey-type resonance pattern, which corresponds to a frequency offset of 250 cycles at 9.2 kMc or to a frequency deviation of $2.7p:10^8$.

3) The decade dividers used in the experiment revealed a relatively high sensitivity to power source instability. Changes in the line voltage level caused occasional slips in the final stages of the divider with the result that the counted time differences displayed then wrong figures for the 1/10 second numbers.

In February, 1960, atomic clocks were set up at the Air Force Missile Test Center, Patrick Air Force Base, Fla. (#134), at the U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J. (#118), and in a C-47 airplane (#119).[3] Each clock consisted of an Atomichron and a pulse divider, with a 1 pulse per second output signal[3] in addition to the 5-Mc and 100-kc signals from the Atomichrons. On February 15, clock #119 was compared with clock #134 over a 1-km long coaxial cable. Observation of the beat note at 5 Mc for a period of 2 hours gave the necessary information for adjusting the rate of plane clock #119 to that of the ground clock #134. The adjustment was accomplished by changing the magnetic C-field of #119. Then, the 1/pps pulses from the ground and plane clocks were used to trigger the start and stop gate, respectively, of a time interval meter, giving the offset, T_1 μsec, between the 2 clocks. At noon the plane took off from Florida and arrived at 1900 at Fort Monmouth, where the ground clock was adjusted in rate and measured in phase in the same way as described above, with the only difference that a 10-km long microwave link was used instead of a coaxial cable. The measurement gave the offset T_2 μsec. Overnight, the plane clock #119 was turned off.

On the following morning, the plane clock was remeasured in terms of the Fort Monmouth clock #118, giving an offset of T_3 μsec. At noon the plane returned to Florida, arriving at 1900 hours. From the values $134\text{-}119 = T_1$, $119\text{-}118 = T_2$, and $119'\text{-}118 = T_3$, one could calculate the value of $134\text{-}119' = T_4 = T_1 + T_2 - T_3$, which was to be expected for the final comparison at Florida, if it was assumed that all clocks performed properly. Table I displays the various measurements and the final result. The clock error between #119 and #134 after a 32-hour interval was only 0.1 ± 0.1 μsec, which corresponds to an average offset in rate of only $1.7p:10^{12}$. It must be noted that the flight weather was excellent, but that otherwise the C-47 airplane is probably the most unsuitable plane for transporting an operating Atomichron because of its unusually large wing surface. We, therefore, no longer have any doubt that we can accomplish a world-wide synchronization of atomic clocks to within a 1-μsec error, provided that fast planes are used and turbulence is only moderate.

During March, 1960, the Signal Corps Atomichrons #117 and #116 were installed on Oahu, Hawaii, and Forestport, N. Y., respectively, for a direct atomic frequency control of the high-power VLF Navy OMEGA transmitters,[4] located at these places. 12.2-kc signals from both stations were measured by Prof. J. A. Pierce at Cambridge, Mass., in terms of Atomichron #112. Table II gives some preliminary results. The relative fractional frequency offsets represent the slopes of the

TABLE I

RESULTS OF THE CLOCK SYNCHRONIZATION TEST BETWEEN FORT MONMOUTH, N. J., AND PATRICK AIR FORCE BASE, FLA., ON FEBRUARY 15–16, 1960

Date	EST	Clocks	Measured T_i μsec	Number of counts	σ_T μsec	Predicted value of T_4 μsec
Feb. 15	11:00	134–119	542173.5^{0}	30	.04	
15	19:00	119–118	778900.5^{3}	261	.04	
16	12:00	$119'$–118	741661.7^{0}	39	.02	
16	19:00	134–$119'$	579412.2^{4}	58	.06	579412.3^{3}

TABLE II*

ATOMICHRON-CONTROLLED VLF TRANSMISSIONS FROM HAWAII (117) AND FORESTPORT (116), RECEIVED AND MEASURED BY PIERCE AT CAMBRIDGE (112), MASS.

Date 1960	Atomichrons	Duration, hours	$\frac{\Delta f^*}{f} 10^{11}$	$\sigma\phi^*$ μ second	$\sigma f \times 10^{11}$
March 3	117–112	5	− 9.	1.4	11.
21/22	117–112	24	− 6.	1.7	3.
24	117–112	6	− 3.	3.	20.
24	116–112	5	−16.	2.5	20.
25	117–112	6	−13.	3.2	21.
25	116–112	6	−27.	3.6	24.

* Data supplied by Prof. Pierce.

[4] C. J. Casselman, D. P. Heritage, and M. L. Tibbals, "VHF propagation measurements for the radux-omega navigation system," PROC. IRE, vol. 47, pp. 829–839; May, 1959.

best fitting straight lines through the recorded phase curves.[5] σ_ϕ is the standard deviation of the phase curve in microseconds, and σ_f is the standard deviation of $\Delta f/f$, calculated from σ_ϕ by dividing it through the total observation time and multiplying it with $\sqrt{2}$ to take into account the uncertainty of both end points of the phase curve. It is believed that the data for σ_ϕ and σ_f can be improved soon by introducing servo control for the phase of the radiated signals, as discussed later. It should also be noted that the available transmission periods were in general too short for a good measurement of frequency. It is significant that the results of the 24-hour transmission almost approached the accuracy of $2p:10^{11}$ achieved during the GBR experiment. The relatively poor quality of the signal from the much closer Forestport station may, at least partly, be caused by strong interference of the first sky wave with the ground wave in the Cambridge area.

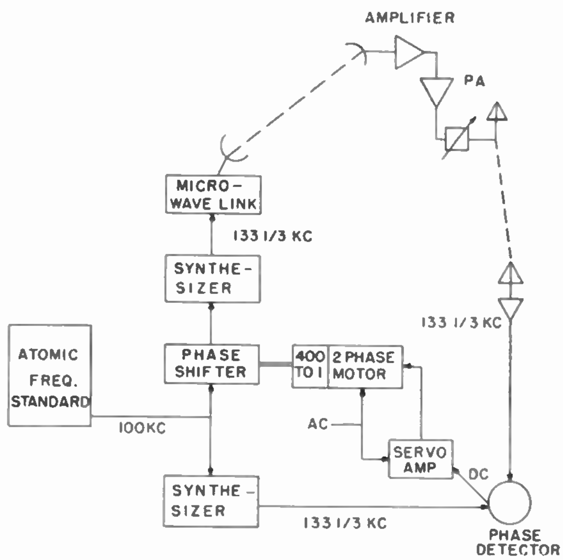


Fig. 2—Block diagram of the A5XA transmitter facilities and its frequency and phase control circuit.

Since October, 1959, the Signal Corps has been operating an experimental LF station, A5XA ($133\frac{1}{3}$ KC), at Earle, N. J., in order to study direct Atomichron control of LF transmission signals and means of reducing phase variations introduced by the transmitter system. The $133\frac{1}{3}$-kc signal is derived from the 100-kc output signal of the Atomichron, and is transferred to the power amplifier at the transmitter site over a 3-km long microwave link. Since April 18, the phase of the radiated signal has been servocontrolled, as shown by the block diagram of Fig. 2. A flat-top antenna with 5 wires, each 100 meters long, and an effective height of 15 meters was used during October and November, 1959. The efficiency of the tuning unit and antenna was about 2.2 per cent, the radiated power 28 watts. In December, a 56-meter high base-insulated antenna tower was substituted for the flat-top antenna, giving a 5-db gain in radiated power for the same input power of 1250 watts. During January, the tower was fitted with a capacitive top load (about 800 $\mu\mu$f) consisting of 6 radial and 1 peripheral wire, which gave an additional antenna gain of 4 db. It was now also possible to raise the input power from 1250 to 5000 watts, so that the total gain over the flat-top assembly was about 15 db, giving a final over-all efficiency of the transmitter facilities of about 18 per cent (900 watts radiated). It was observed that full-power operation since the introduction of the antenna top load introduced considerable slow variations of the transmitter impedance because of heat dissipation in the tuning unit. There were also short-time variations during windy weather because of the unavoidable sag of the peripheral wire. It is the purpose of the above mentioned servo control to counteract the resulting phase variations. Table III, Figs. 3 and 4 (page 1032) give the results of comparing Atomichrons, located at Fort Monmouth, with Atomichron #112 at Cambridge over the $133\frac{1}{3}$-kc transmission path. A significant improvement of the standard deviations since the introduction of the servo control is apparent despite some antenna breakdown problems and a deteriorating beam tube of Atomichron #115 (Fig. 3), both of which happened to enter the picture just about the same time as the servo was put into operation. The results shown in Fig. 4 were achieved with the better Atomichrons #111 and #119. During August, 1960, new models NC2001 of atomic beam standards will be employed on both ends to see if further improvements will be possible by using improved frequency standards for controlling the transmitter and receiver.

[5] J. A. Pierce, "Intercontinental frequency comparison by very low-frequency radio transmission," Cruft Lab., Harvard Univ., Cambridge, Mass., Tech. Rept. No. 220; 1957.

TABLE III*

RESULTS OF COMPARISONS OF ATOMICHRONS LOCATED AT FORT MONMOUTH WITH ATOMICHRON 112 AT CAMBRIDGE, MASS., VIA 133⅓-KC TRANSMISSION SIGNALS

Date	Antenna	Atomichrons	Time, hours	Average interval, minutes	$\frac{\Delta f^*}{f} \times 10^{11}$	$\sigma\phi^*$ μsec	$\sigma f \times 10^{11}$
1959							
Oct. 22	Flat top	8–112	6	60	−1.4		
28	"	8–112	6	60	−3.4		
29	"	8–112	6	60	−5.2		
30	"	8–112	2	10	−0.6	0.06	1.2
30	"	119–112	4		−10.8	0.11	1.1
Nov. 4	"	8–112	2	10	−3.6	0.07	1.4
4	"	119–112	2	10		0.06	1.2
4	"	8–112	3	30	−3.4	0.05	0.7
4/5	"	8–112	14	60	+22.6	0.26	0.7
5	"	8–112	5	60	−2.0	0.18	1.4
5	"	8–112	2	10	−4.7	0.06	1.2
5	"	115–112	2	10	−3.0	0.04	0.8
5	"	115–112	2	10	−0.1	0.03	0.6
12	"	115–112	4½	30	−6.0	0.07	0.6
13	"	115–112	6	30	−5.5	0.05	0.3
Dec. 10	Tower	115–112			−7.0		
1960							
Feb. 5	Tower with top load	115–112			−11.0		
8	"	115–112			−14.0		
8	"	115–112			−17.0		
11	"	115–112			−13.0		
15	"	115–112			−13.0		
16	"	115–112			−15.0		
17	"	115–112			−17.0		
18	"	115–112	6	30	−16.0		
19	"	115–112	6	30	−14.0		
Mar. 10	"	115–112	8	30	−3.4	0.17	0.8
14	"	115–112	6	30	−11.8	0.22	1.4
17	"	115–112	4½	30	−1.2	0.16	1.3
21	"	115–112	6	30	−10.0	0.08	0.5
24	"	115–112	3	20	−25.0	0.06	0.7
28	"	115–112	6	30	−2833.0‡		
31	"	115–112	6	30	−2.8	0.16	1.1
Apr. 18	Tower with top load, and servo control	123–112	2½	10	−8.7	0.06	0.9
21	"	115–112	3	10	−4.5	0.06	0.8
25	"	115–112	3	30	−30.5†	0.11	1.5
25	"	115–112	2	30	−48.0†	0.20	3.9
28	"	115–112	2	10	−22.0	0.70	1.4
May 12	"	111–112	6	10	+21.0	0.05	0.52
16	"	111–112	7	30	+19.5	0.1	0.56
19	"	111–112	8	30	+21.1	0.06	0.27
23	"	119–112	4	20	−10.6	0.07	0.65
June 2	"	119–112	8	30	+11.	0.08	0.4
6	"	119–201	3⅓	20	+15.	0.05	0.58
6	"	112–119	4½	30	+ 4.5	0.05	0.45
6/7§	"	112–119	6½	30	+14.	0.71	4.3
7	"	112–119	7½	30	+ 8.3	0.08	0.41
7	"	201–119	8	30	+ 0.11	0.06	0.31
7/8§	"	201–119	6	30	+ 0.37	1.03	6.8
8	"	201–119	12½	30	+ 1.0	0.16	0.5

* Data supplied by Prof. Pierce.

† Beam tube of 115 had very low signal-to-background ratio, and failure of the room air conditioner caused frequency drift. Measurements were therefore divided over 2 intervals.

‡ Atomichron 115 was accidentally locked at lower side peak.

§ Night transmission only.

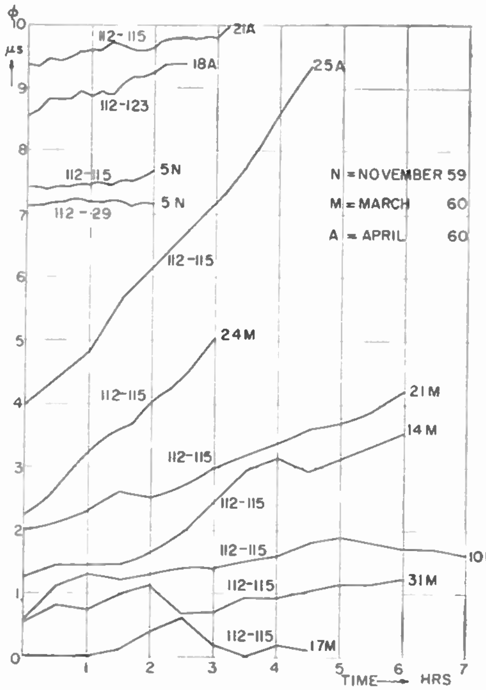


Fig. 3—Some typical phase recordings for A5XA transmission signals.

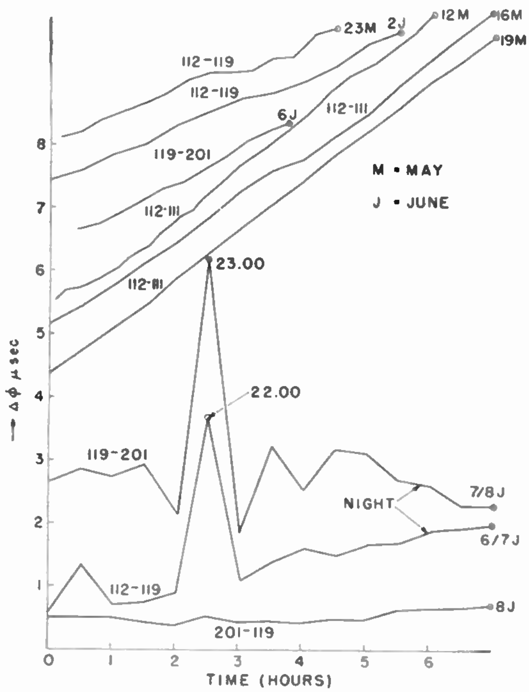


Fig. 4—Phase recordings for A5XA achieved during May and June, 1960.

TABLE IV

Results of Comparison of Atomichrons in the Signal Corps Laboratory

Date	Atomichrons	Time, hours	Average interval, minutes	$\frac{\Delta f}{f} \times 10^{11}$	$\sigma\phi$ μsec	$\sigma f \times 10^{11}$
1959						
Nov. 8/9	201-8	5	30	4.6	0.005	0.04
Nov. 8/9	119-8	13	60	7.9	0.07	0.2
Nov. 8/9	201-8	21	60	5.0	0.05	0.1
Nov. 8/9	201-8	5	60	4.5	0.009	0.1
Nov. 8/9	201-8	5	60	5.4	0.04	0.3
1960						
Feb. 12/20	118-115	195	900	1.0	0.3	0.1
Feb. 12/20	123-118	180	900	1.8	1.9	0.4
Feb. 12/20	123-118	84	900	0.3*	0.5	0.2
Feb. 12/20	123-118	75	900	3.9*	0.3	0.2
Apr. 29	119-8	5½	30	0.4	0.2	0.2
Apr. 29	111-119	5½	30	21.3	0.02	0.14
Apr. 29	115-119	3	30	0.75†	0.006	0.08
Apr. 29	115-119	2½	30	−59.†	0.5	8.
Apr. 29	118-119	5½	30	9.4	0.02	0.12
Apr. 29	123-119	5½	30	37.2	0.23	1.7
May 3	115-119	6	30	4.9	0.01	0.1
25	201-202‡	8	30	1.77	0.0034	0.017

* Split-up over two intervals because 123 developed a small frequency shift after 84 hours, caused by changing room temperature.
† Split-up over two intervals because 115 drifted because of defective air conditioner.
‡ New Atomichrons, type NC2001.

Table IV gives the results of comparisons of several Atomichrons in the Signal Corps Laboratory in order to show that there is some good hope to approach with LF or VLF transmissions the precision achieved with present standards under laboratory conditions.

Further experiments on long-range synchronization, and additional propagation studies were performed during the summer of 1960.

ACKNOWLEDGMENT

The authors acknowledge Prof. J. A. Pierce's enthusiastic cooperation, and appreciate his permission to include his measurement results in this publication. The authors also appreciate the excellent cooperation of the Missile Test Center at the Patrick Air Force Base, Fla.

High-Frequency Power in Tunnel Diodes*

G. DERMIT†, MEMBER, IRE

***Summary*—An expression for high-frequency tunnel-diode power is developed with a simple dissipative load and for signals confined within the nearly linear range of the diode negative resistance. The existence of maximum power output for a given frequency of operation, and for an optimum peak current-to-capacitance ratio is demonstrated. The power and the inductance appear in the equations as a product, and consequently, the higher-power output diodes require the lower inductances.**

The results are plotted for germanium as well as for gallium arsenide in the form of charts with the ratio of peak current-to-capacity and the operating frequency as coordinates. These charts present two families of curves which cross each other, one referring to constant products of power times inductance and the other referring to constant products of capacity times inductance. The presented data cover a wide operational range of tunnel diodes. For a diode oscillating at 10 kMc and having a resistive cutoff frequency of 30 kMc, the maximum power-inductance product is 0.0224 (mw)(mμh) for a voltage swing of 0.15 volt in the nearly linear range.

I. Introduction

Majority carrier tunneling in Esaki diodes takes place at extremely high speeds compared to minority carrier injection across ordinary junctions. Tunnel diodes that utilize this mechanism have very high-frequency capabilities, and promise useful applications in high-speed switching, and as high-frequency oscillators and amplifiers.

One approach to high-frequency tunnel diodes has been to build a point contact junction directly into the cavity. GaAs diodes fabricated in this way have been reported to oscillate in third- or fourth-order modes[1] at as high as 40 kMc. These "pulsed-junction" diodes have the advantage of giving extremely small junction areas directly without requiring subsequent reduction by etching.

Another approach has been to construct very small, individually enclosed, "alloyed-junction" units. Dominant mode oscillations up to 10.8 kMc have been reported[2] with germanium tunnel diodes of this type. Their advantage lies in that they allow fairly accurate reproduction of parameters and stable encapsulation.

The purpose of this paper is to show the development of a general expression for the maximum power output of tunnel diodes operating within the nearly linear range of their negative resistance characteristic. This linear range is shown to be quite large in comparison to the entire negative resistance region in tunnel diodes. The analysis, and the formulation of a general expression for power is carried out for conditions of steady sinusoidal oscillation.[3,4]

II. Steady Sinusoidal Oscillation

A dc current-voltage characteristic of a typical tunnel diode in the forward direction is shown[5] in Fig. 1. It consists of three distinct parts: the peak region, the valley region, and the forward exponential rise. The positive and negative resistance regions on either side of the peak are due to majority carrier tunneling. The valley region is variously regarded as being due to carrier tunneling to intermediate states followed by recombination, and the final exponential rise is due to forward injection of minority carriers. The last two mechanisms of conduction are slow in comparison to normal tunneling. Thus, the usefulness of the valley and forward injection regions at high frequencies, 10 kMc and higher, is questionable. In this paper the emphasis is on operation in a region of the characteristic dominated by normal band-to-band tunneling only.

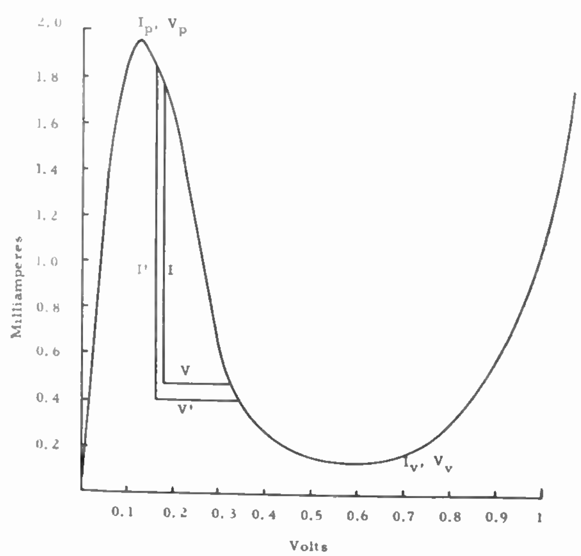


Fig. 1—Typical GaAs tunnel-diode current-voltage characteristic with forward bias. V_p = peak voltage; V_V, valley voltage; V and L, voltage and current swings over linear range of negative resistance, respectively.

* Received by the IRE, December 28, 1960; revised manuscript received, February 24, 1961. The contents of this paper appeared in a Tech. Memo. (TM 366.1), October 7, 1960, except for some qualitative considerations on nonlinear oscillation and a specific calculation of maximum power at 10 kMc. A treatment of inductance and frequency limitation of cylindrical cavities with tunnel diodes was included in the Memo. but has been omitted in this paper.

† Genl. Telephone and Electronics Labs., Inc., Bayside, N. Y.

[1] R. Trambarulo and C. A. Burrus, "Esaki diode oscillators from 3 to 40 kMc," Proc. IRE, vol. 48, pp. 1776–1777; October, 1960.

[2] G. Dermit, H. Lockwood, and W. Hauer, "10.8 kMc germanium tunnel diode," Proc. IRE, vol. 49, pp. 519–520; February, 1961.

[3] M. E. Hines, "High-frequency negative-resistance circuit principles for Esaki diode applications," *Bell Sys. Tech. J.*, vol. 39, pp. 477–513; May, 1960.

[4] R. P. Murray, "Biasing Methods for Tunnel Diodes," *Electronics*, vol. 33, pp. 82–83; June, 1960.

[5] N. Holonyak and I. A. Lesk, "Gallium arsenide tunnel diodes," Proc. IRE, vol. 48, pp. 1405–1409; August, 1960.

For a first-order characterization of the device we restrict ourselves to the linear portion of the negative resistance region which is described by the voltage and current swings, V and I. The linear portion we are defining as the region in which the average value V/I does not exceed the minimum value of the negative resistance by more than 15 per cent. In Section IV-B it is shown that this region accounts for a large fraction of the total power available in tunnel diodes. The quantities V and I are related by the equations

$$V = m_v(V_v - V_p)$$
$$I = m_i(I_p - I_v), \tag{1}$$

where V_v, V_p and I_v, I_p are the valley and peak values of the voltages and currents, respectively. The symbols m_v and m_i are constant fractions specifying the linear portion of the over-all negative resistance characteristic. For most diodes the difference between the peak and valley currents, $I_p - I_v$, is close to I, and since we mainly consider diodes with appreciable peak-to-valley current ratios, I_p/I_v, we can take $I = I_p$. Diodes with small m_i and small current ratios are considered in Appendix I. The voltage swing is an inherent property of the junction. It is related to the energy-gap and Fermi levels of the semiconductor, and does not depend on the area of the junction. This is in contrast to the peak current, which is directly proportional to the area of the junction and depends very much upon the semiconductor material. The quantity V does depend on doping, but this dependence is very small, because the peak voltages of the diode change only a small amount with doping.

A. Conditions for Steady Sinusoidal Oscillations

The circuit diagram in Fig. 2 depicts a lumped parameter network for a tunnel diode, where L is the inductance of the diode, R_s the series resistance, R_L the external load, $-R$ the negative resistance of the junction, and C its capacitance. In order to start oscillations at the angular frequency ω, the diode is biased in the negative resistance region with a low-impedance source which in turn is bypassed with a large capacitor. The equivalent series impedance of the circuit is given by:

$$Z = R_L + R_s - \frac{R}{1 + R^2\omega^2C^2} - j\frac{R^2\omega C}{1 + R^2\omega^2C^2} + j\omega L. \tag{2}$$

When self oscillations occur, the current I must be finite. Since the equation

$$IZ = 0 \tag{3}$$

must hold at all times, steady sinusoidal oscillations not exceeding the linear range occur only when the impedance Z is zero.[4] Setting both the real and imaginary parts of Z equal to zero, and solving for the corresponding frequencies, leads to

$$\omega_R = \frac{1}{CR}\sqrt{\frac{R}{R_L + R_s} - 1}, \tag{4}$$

in the case of the real part, and to the inductive cutoff frequency

$$\omega_L = \sqrt{\frac{1}{CL} - \frac{1}{R^2C^2}} \tag{5}$$

in the case of the reactive part. It should be noted that ω_R increases with the decreasing time constant RC, whereas ω_L approaches zero for a given inductance L, when RC approaches $\sqrt{CL}$. For steady sinusoidal oscillations at a frequency ω, both conditions (4) and (5) have to be fulfilled simultaneously, which means $\omega = \omega_R = \omega_L$. A more general discussion of the oscillation conditions is given in Appendix II.

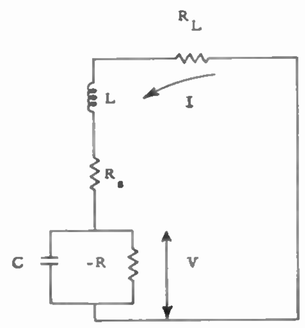


Fig. 2—Equivalent network for tunnel diode. The junction parameters are C, $-R$, and R_s. The inductance of the cavity and the load are L and R_L.

B. Power Output

For the purposes of the present treatment the oscillations have been confined to the linear range. This mode of operation is of primary interest in amplifiers and high-frequency oscillators. It is also of interest in low-frequency oscillators, especially since it turns out that the available power in the linear range is a substantial percentage of the total power of a tunnel-diode oscillator which is not restricted to the linear range (see Section IV-B).

Since the useful energy is dissipated in the load R_L, the available power output is given by

$$P = I^2R_L = I^2\left(\frac{R}{1 + R^2\omega^2C^2} - R_s\right), \tag{6}$$

where the second relation follows from (4). Referring to Fig. 2, we can write the relationship between the rms current I and the linear voltage swing V as

$$I = \frac{V}{2\sqrt{2}}\left(-\frac{1}{R} + j\omega C\right), \tag{7}$$

which gives

$$|I|^2 = \frac{V^2}{8}\,\frac{1 + R^2\omega^2C^2}{R^2}. \tag{8}$$

Inserting this into (6) we have

$$P = \frac{V^2}{8}\,\frac{1 + R^2\omega^2C^2}{R^2}\left(\frac{R}{1 + R^2\omega^2C^2} - R_s\right). \tag{9}$$

This equation must be subjected to condition (5), before it can describe the power for sinusoidal oscillations. However, we can discuss the concept of resistive cutoff frequency and a few experimental results without introducing (5). At low enough frequency, when $\omega \ll 1/RC$, the power output in (9) approaches the dc power

$$P_{\mathrm{dc}} = \frac{V^2}{8}\,\frac{R - R_s}{R^2}. \tag{10}$$

The frequency at which P drops to $1/n$ of the low-frequency power is

$$\omega^2 = \frac{1}{R^2C^2}\,\frac{R - R_s}{R_s}\,\frac{n-1}{n}. \tag{11}$$

From this formula we can calculate the resistive cutoff frequency ω_{R_s} at which the power drops to zero. This occurs at $n = \infty$, *i.e.*,

$$\omega_{R_s}{}^2 = \frac{1}{R^2C^2}\,\frac{R - R_s}{R_s}. \tag{12}$$

Note that this is simply (4) for $R_L = 0$ or zero power. Since the concept of resistive cutoff frequency is contained as a special case within condition (4), we have been able to arrive at a correct conclusion despite the fact that (9) is not sufficiently restricted to include, as yet, (5).

For most diodes the series resistance R_s is much smaller than the negative resistance R, which reduces the resistive cutoff frequency to

$$\omega_{R_s} = \frac{1}{C}\,\frac{1}{\sqrt{RR_s}}. \tag{13}$$

and the power output in (9) to

$$P = \frac{V^2}{8}\left(\frac{1}{R} - R_s\omega^2C^2\right), \tag{14}$$

or

$$P = \frac{V^2}{8R}\left(1 - \frac{\omega^2}{\omega_{R_s}{}^2}\right) \tag{15}$$

The limitations imposed by the assumption $R \gg R_s$, and a more general expression for power avoiding this assumption are given by (31) and (30) in Appendix III, respectively.

This expression for power is based on a strictly resistive load R_L. According to the oscillation condition (5), capacitance, inductance and negative resistance are enough to determine the frequency of oscillation. Experimentally, it is observed that the frequency can be changed slightly by moving the bias point to higher resistance values in the negative resistance region. For diodes oscillating close to their cutoff frequencies, oscillations may cease altogether with a small change in bias, whereas, for oscillations far below cutoff, the bias can be shifted well into the valley region without stopping the oscillations.[2] For a typical tunnel diode with $L = 0.25$ mμh, $C = 0.784$ μμf, $R_s = 6$ ohms, and $R = 55$ ohms, we observed oscillations up to 10.8 kMc, which is in good agreement with 10.75 kMc as calculated from (5). Experimentally, the frequency for such a diode can be shifted about 100 Mc by changing the bias point.

Although our analysis applies strictly to resistive loads, the extension to complex loads is straightforward. In either case, however, for a given diode operating at a given frequency, the load must be fixed to a value determined by the oscillation conditions (4) and (5). If the load exceeds this value, oscillations cease. If the load is slightly smaller than this value, the voltage swing extends into the nonlinear range to a certain degree and causes a stabilization due to a corresponding change in R. For much smaller loads the oscillations become highly nonlinear, provided the frequency of oscillation is not too high.

Eq. (15) as it stands includes only the effects of oscillation condition (4). According to (3), both oscillation conditions (4) and (5) must be satisfied simultaneously in order for the diode to carry out sinusoidal self-oscillations. Consequently, we have to introduce into (15) the functional dependence given by (5). We could substitute either for ω or R; since for our purposes we are interested in obtaining a power expression having an explicit dependence on the time constant of the junction $RC = CV/I$, we substitute for R. RC is essentially one of the most significant parameters in the design of high-frequency tunnel diodes. Since V is fixed, let us introduce the peak current-to-capacity ratio $I_p/C = k$ which can be measured easier than R in a tunnel diode. We then can substitute in (15)

$$R = \frac{V}{I} = \frac{V}{I_p} = \frac{V}{kC}. \tag{16}$$

which results in

$$P = \frac{VkC}{8}\left(1 - \frac{\omega^2}{\omega_{R_s}{}^2}\right). \tag{17}$$

A combination of (5), (16), and (17) finally yields

$$P = \frac{V}{8L}\,\frac{k}{\omega^2 + k^2/V^2}\,(1 - \omega^2/\omega_{R_s}{}^2). \tag{18}$$

This is the equation for the output power of a tunnel diode operating at a frequency ω, and within the linear region specified by V and I in Fig. 1, subject to condition $R \gg R_s$ or (31). The general power expression where this condition is removed is given by (32).

When the frequency of oscillation is close to the resistive cutoff frequency, the power approaches zero. On the other hand when $\omega \leq 1/3\omega_{R_s}$, the ratio $\omega^2/\omega_{R_s}^2$ is small compared to 1 and can be neglected. In this case it can easily be shown that at a given frequency of operation $\omega = \omega_0$, the power P reaches a maximum at an optimum value of $k = k_0$, given by

$$\omega_0 = k_0/V, \tag{19}$$

where k_0 corresponds to an optimum I_p/C ratio of the tunnel diode. Consequently,

$$P_{\max} = \frac{V^3}{16k_0L} = \frac{V^2}{16\omega_0L}. \tag{20}$$

In Appendix III, (19) and (20) are shown to hold even when ω is close to ω_{R_s}. Although (19) is shown to be completely general, (20) holds strictly when $R/2 \gg R_s$. The general expression for (20) is given by (33). From (19) the important conclusion that can be drawn is that at a given frequency ω_0 there exists an optimum value k_0 for which the tunnel diode yields a maximum power output given by (20). Furthermore, it is very important to note that the maximum power, which will be discussed further in Section IV, is strongly affected by V. This voltage-swing V depends mainly on the semiconductor material, its energy gap, and its Fermi levels. High-energy-gap semiconductors apparently give larger voltage swings, and consequently lead to higher powers. Besides its dependence on the degeneracy of the Fermi levels, V also is related to the tail-off of the density of states that takes place at the band edge. Quantitative relations for these effects have not been published. Based on experimental data we used $V = 0.1$, and 0.15 for Ge, and GaAs, respectively. It is interesting to note that when the effects of R_s are negligible, *i.e.*, if $\omega \leq \omega_{R_s}/3$, (15) does not contain ω explicitly, whereas (18) does. Of course, (15) is a more general expression and while it holds for every oscillating diode, the converse statement that any arbitrary value of R is sufficient for a diode to carry out sinusoidal oscillations is not correct.

So far, we have considered only the linear voltage range, *i.e.*, the range where the negative resistance is maximum. When the diode oscillates close to its cutoff frequency, or when R_L is sufficiently large [see (23)], the voltage is confined very closely within this range. However, if the diode is oscillating at low frequencies and considerably below its cutoff, V can swing over a wider range, extending well into the nonlinear regions. The oscillations under these conditions become quite complex and, depending on the diode, the apparent power which now includes strong harmonics and cross-modulation, can be about a factor two higher than the peak powers given before (see Section IV-B). It should be noted that in our analysis V is fixed; as ω approaches the cutoff frequency ω_{R_s}, the power goes to zero because R_L becomes zero (see Section IV-B), and not because of a decreasing voltage swing.

III. Tunnel-Diode Charts

In this section numerical computations based on (18) are presented. For this purpose $\omega_{R_s}^2$ in (18) can be replaced by

$$\omega_{R_s}^2 = k/CR_sV, \tag{21}$$

which follows from (13) and (16).

The product CR_s should be considered further. To a first order of approximation, this product is independent of the junction size, because C is proportional to the area of the junction, and R_s is inversely proportional. We shall now show that the dependence of the product on various doping concentrations is also small. The resistivity range in germanium tunnel diodes is about 0.0008 to 0.0004 ohm-cm. This range would give a 2 to 1 variation in R_s. If the usual *p-n* junction theory holds for these degenerate levels, the capacitance is proportional to the square root of the impurity concentration. Due to the severe drop in mobility going from 0.0008 to 0.0004 ohm-cm, the impurity concentration must be increased by a factor of about 4. Thus, over the entire useful range of resistivities, the product CR_s would tend to remain independent of doping.

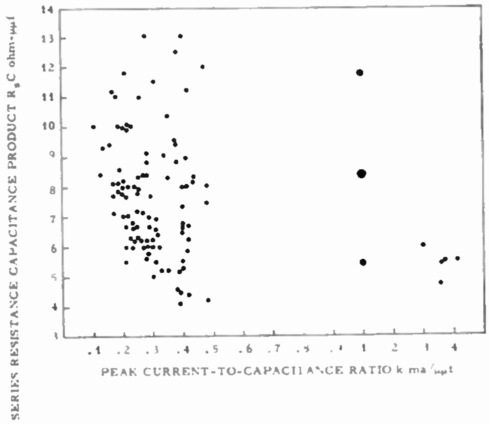


Fig. 3—Series resistance-capacitance product R_sC for various Ge diodes vs peak current-to-capacitance ratios k.

In Fig. 3 a distribution of the product R_sC for over a hundred experimental germanium diodes, covering a wide range of k's, is shown. More than 50 per cent of these diodes which were prepared by alloying and subsequent etching fell within the range of 5.5 to 8 ohm-$\mu\mu$f. Since we are interested in the higher k units for high frequency, we shall choose $R_sC = 6$ ohm-$\mu\mu$f. With point contact junctions this figure would be lowered by as much as an order of magnitude. If R_sC is different from 6 it will be shown in Section III-B that a correction, particularly at high frequencies, can be applied very easily. Further discussion of R_s is taken up in Section IV following the presentation of charts for Ge and GaAs diodes.

A. Germanium

Using (21) for the cutoff frequency in the power expression (18), and inserting for germanium $V=0.1$ volt and $CR_s=6$ ohm-$\mu\mu$f, we are left with the quantities P, L, k and ω. Since the power and the inductance occur as a product, we can plot (18) as a series of constant power-inductance curves with k as a function of ω. In Fig. 4(a)–4(c), next page, these plots are shown over a wide range of frequencies. These plots are accurate for k values that are about 3 ma/$\mu\mu$f or less. This limitation on k is due to condition (31), giving $R \geq 5R_s$, but is removed in (32). The power-inductance values for each curve have been indicated as power by assuming a fixed inductance of 0.25 mμh. The powers corresponding to any other inductance can be computed easily from the same plots. One striking feature of these curves is that the k values are double valued with respect to frequency. In general, the lower the frequency, the higher the power, provided one uses the proper k value, or time constant, for the junction. k is also proportional to the peak-current density of the diode. Thus, for a given frequency of operation there is an optimum peak-current density that one must use in order to obtain maximum power. If k is much higher than this optimum value, the power output drops. This behavior holds primarily for power diodes. If, for instance, a one kMc oscillator is desired, the permissible value for k is only 0.63, which corresponds to a peak-current density of approximately 6000 amp/cm^2. If a k value of 1.7 is used instead, the power at 1 kMc will drop from the original level of 0.358 mw to 0.25 mw.

Considering now the low-power curves of Fig. 4(a)–4(c), we observe that the zero-power curves are the cutoff curves of the diodes. For high-frequency, low-power diodes, we see that the frequency increases with k. The highest k values reported in Batdorf, *et al.*,[6] are for InSb, with $k=16$ ma/$\mu\mu$f. For Ge, the maximum values obtained so far are smaller, in the range of 10 ma/$\mu\mu$f. However, even before the frequency limitation due to the k value is reached, there are other more serious limitations related to the problem of the smallest junction size that can be achieved in practice.

We have already seen that the area of the junction is proportional to its capacitance. The behavior of capacitance can be plotted also on the tunnel-diode chart. To show this, we eliminate R in (5) through the substitution of (16). The resulting expression is

$$CL = \frac{1}{\omega^2 + k^2/V^2} \cdot \tag{22}$$

Constant capacitance-inductance product curves can now be plotted with k as a function of ω. These product curves are shown in Fig. 4(a) and 4(b) as capacitance curves, since they have been computed for a fixed inductance of 0.25 mμh. The capacitance values corresponding to any other inductance can be calculated from the same curves.

The characterization of a tunnel diode on these charts is now complete. Any point on the chart gives the frequency of oscillation, peak-current, capacitance and power output of the diode for an inductance of 0.25 mμh. In case of larger inductances, for instance, if the diode is placed in a larger cavity, the power and capacitance figures are reduced by the ratio of the small inductance to the large one. Since k as an intrinsic property of the junction remains unchanged, we move to the left on a horizontal line representing constant k; vice versa, a reduction in inductance results in a gain of the power output.

The constant CL curves, besides giving practical information about peak-currents, also establish the highest possible frequency of operation. Moving to higher frequencies on the 2.5 μw curve requires smaller and smaller capacitances, or smaller junction sizes. The smallest mechanically-stable junctions that can be fabricated by etching are estimated to be close to 0.2 mil. This corresponds approximately to capacitances as low as 0.7 $\mu\mu$f, for $k=3.5$. Referring now to Fig. 4(b) we see that for $k=3.5$ and $C=0.7$ $\mu\mu$f, we obtain on the 2.5 μw curve approximately a frequency of about 11 kMc. Diodes represented by this point on the chart have been constructed.[2]

B. Gallium Arsenide

The charts shown in Fig. 5(a) and 5(b), p. 1039, are equivalent plots for GaAs. The importance of a larger voltage swing, $V=0.15$ volt as compared to 0.1 volt for Ge, becomes clearly apparent in these curves. These plots are accurate for values of $k \leq 5$ ma/$\mu\mu$f which satisfy (31), giving $R \geq 5R_s$. This limitation is removed in (32). All other parameters for GaAs are the same as for Ge. The curves show that the maximum power according to (20) has now increased with V^2 for a fixed frequency. For instance, at 1 kMc the peak power is 0.84 mw, instead of 0.358 mw, the value for germanium.

For GaAs we have assumed $CR_s=6$ ohm-$\mu\mu$f as for Ge. Zinc-doped GaAs tunnel diodes actually have a resistivity of about 0.002 ohm-cm, and consequently lead to series resistances that are higher than those of Ge diodes. On the other hand, the capacitance per unit area of GaAs diodes is about one half that of Ge diodes, and therefore, the CR_s product does not change appreciably. In order to show clearly the effects of higher series resistance, we have included in Fig. 5(a) a set of dashed curves for a diode that has a series resistance an order of magnitude higher than 6, *i.e.*, $CR_s=60$ ohm-$\mu\mu$f. In this case, the dashed curves in Fig. 5(a) are accurate for $k \leq 0.5$ ma/$\mu\mu$f only. The effect on the 2.5-μw oscillation is obviously detrimental; the frequencies for a given k value have dropped by the square root of ten.

[6] R. Batdorf, G. Dacey, R. Wallace, and D. J. Walsh, "Esaki diode in InSb," *J. Appl. Phys.*, vol. 31, pp. 613–614; March, 1960.

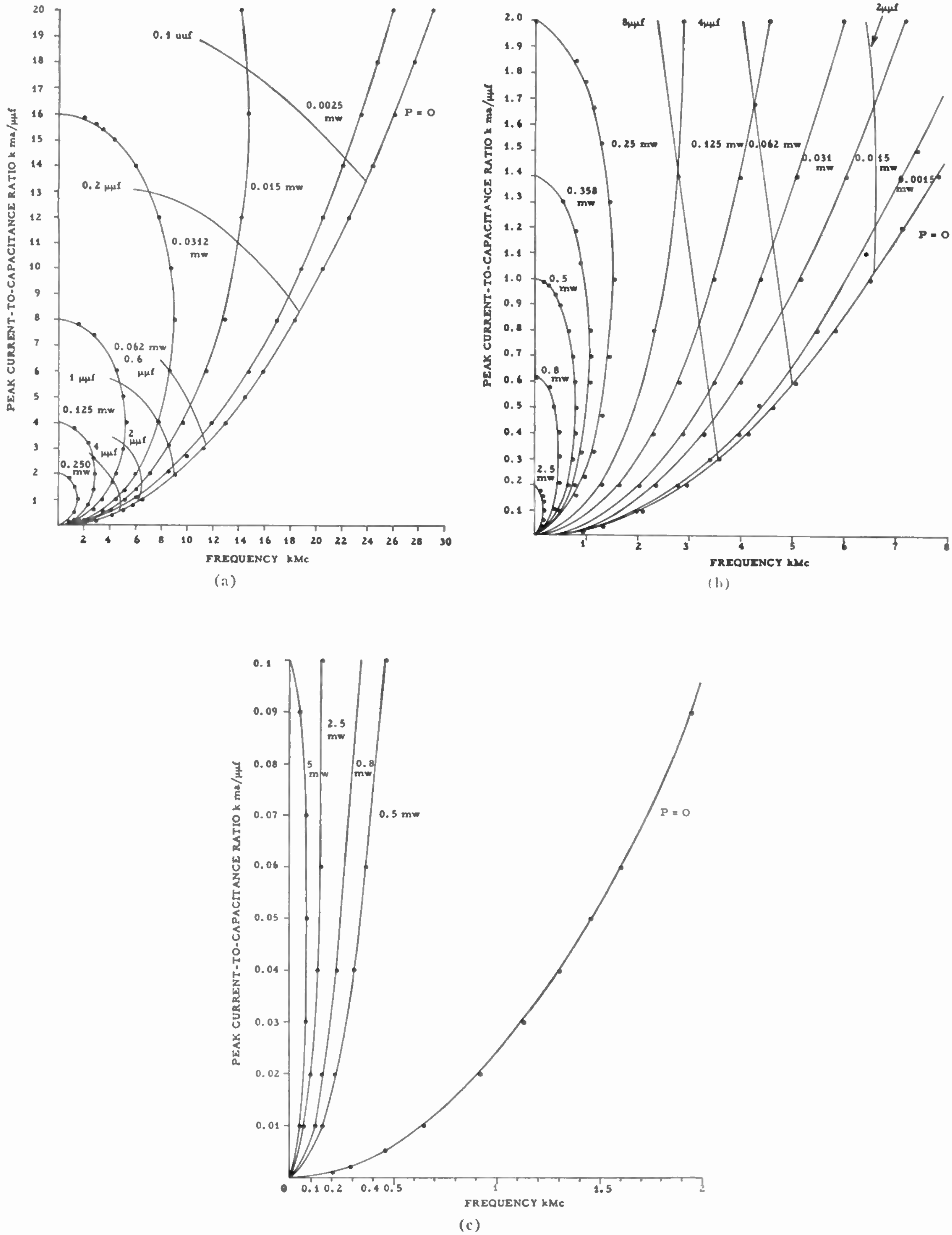
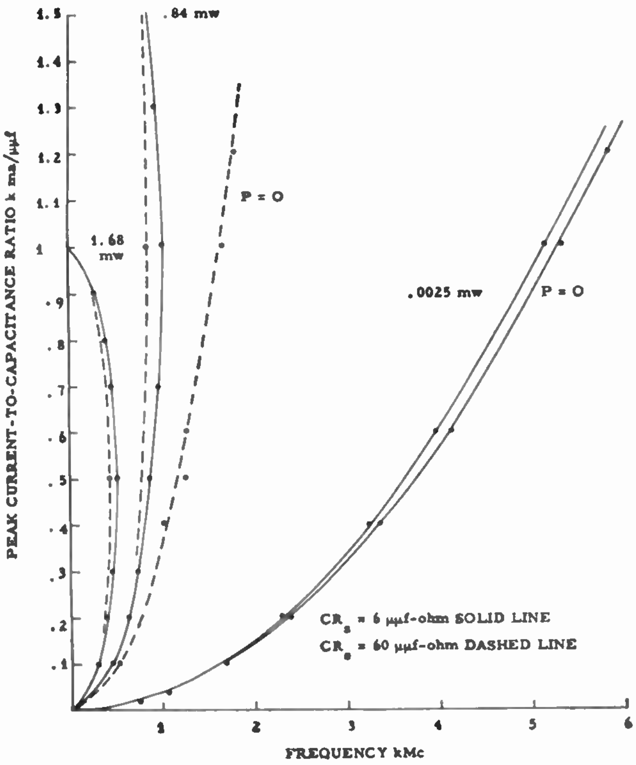
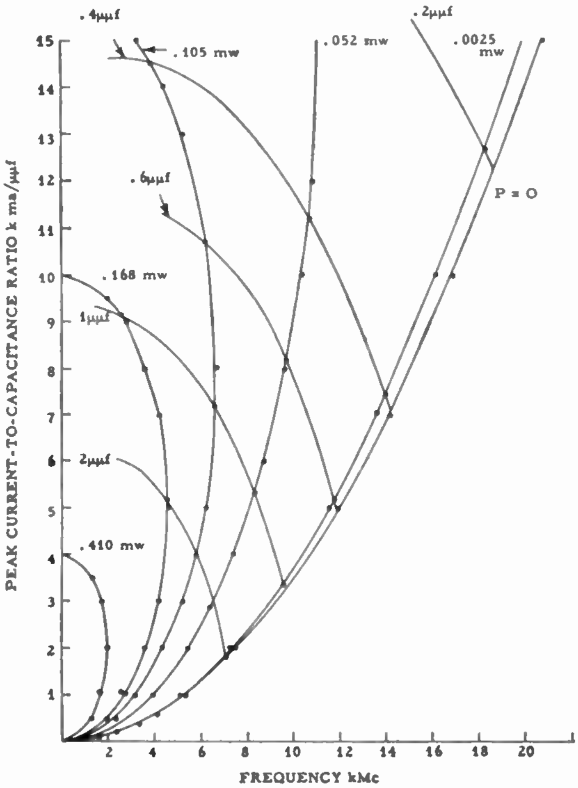


Fig. 4—(a), (b) and (c). Germanium constant power-inductance product, PL curves for $L=0.25$ mμh, $V=0.1$ volt, and $R_sC=6$ ohm-μμf. V is the voltage swing over the linear range of negative resistance; R_sC is the diode series resistance junction-capacitance product. Constant capacitance-inductance CL curves for $L=0.25$ mμh, and $V=0.1$ volt have been included in (a) and (b).



(a)



(b)

Fig. 5—(a) GaAs constant power-inductance product, PL curves for $L=0.25$ mµh, voltage swing $V=0.15$ volt, and diode series resistance-capacitance product $R_sC=6$ ohm-µµf. (b) Similar constant PL curves with constant capacitance-inductance CL curves for $L=0.25$ mµh and $V=0.15$ volt added.

However, at lower frequencies of operation the effect is negligible.

Besides offering higher power at low frequencies, GaAs diodes could be etched down to about one half the capacitance of Ge junctions because of their lower capacitance per unit area. This property extends their high-frequency range.

IV. Maximum Power Output

The conditions for steady sinusoidal oscillations are given by (4) and (5) with the additional requirement that $\omega_R=\omega_L$. The equality of the two frequencies leads to

$$\frac{R_L+R_s}{L}=\frac{1}{RC}, \tag{23}$$

where R_s depends on junction size and semiconductor resistivity. For constant resistivity, R_s decreases with increasing junction area, but RC remains constant. For large enough junctions R_s is small compared to R_L, and therefore R_L approaches L/RC, a value independent of junction size.

For very high-frequency tunnel diodes, (5) requires small capacitances which necessitate the use of very small junctions. Alloyed junctions can be etched down to very small diameters of about 0.1 to 0.2 mil, but etching results in relatively high R_s because of the formation of a thin column at the junction. In computing the charts of Section III-A and B, we used R_s values that were taken from ordinary heavily-etched type junctions with relatively high R_s ($R_sC=6$ ohm-µµf).

The problem of reducing R_s and, at the same time, maintaining a small junction area can be attacked in two ways. One way is to use pulsed-point contact junctions.[1] Here, R_s approaches the spreading resistance of the point which is much lower than the resistance of the etched column. The other way of reducing R_s is by the use of very thin semiconductor layers. In this case, small junctions require wafer thicknesses which are a fraction of the junction diameter, and quite difficult to achieve. These two ways of reducing R_s leave the parameters L, R, C unchanged, and from (5) it follows that the operating frequency remains unchanged. For small enough R_s, the load R_L approaches L/RC, as stated above.

It should be noted that an attempt to decrease R_s is of importance only if the condition $R_s \ll L/RC$ is not yet fulfilled. It is quite obvious that this condition is easier met if the time constant RC is lower or k is higher. On the other hand, the lower the inductance L, the lower R_s has to be.

A. Maximum Power in the Linear Range

When a tunnel diode oscillates close to its resistive cutoff frequency, (4) and (12) require R_L to approach zero and the power developed in the load also approaches zero. This means, from (23), that R_s approaches the value L/RC. If R_s is decreased, the cutoff frequency

ω_{R_s} goes up and consequently the power output at an operating frequency ω increases as seen by (18). The power loss due to R_s is less than 10 per cent of the total power when $\omega < \frac{1}{3}\omega_{R_s}$.

As pointed out in Section II-B, (18) reduces, in this case, to

$$P = \frac{V}{8L}\,\frac{k}{\omega^2 + k^2/V^2}\,. \tag{24}$$

The $P_{\max}$ and k_0 values are given by (20) and (19). If tunnel-diode charts, similar to the ones shown in Figs. 4 and 5, are plotted for low R_s according to (19), the $P_{\max}$ points fall on a straight line which goes through the origin. Furthermore, by combining (22) and (19), the diode capacity-inductance product for maximum output is

$$C_0L = \frac{1}{2\omega_0^2} = \frac{V^2}{2k_0^2}\,. \tag{25}$$

The load resistance for maximum power output is obtained by combining (23), (19) and (16). For $R_s \ll R_L$ we have

$$R_L = Lk_0/V = L\omega_0. \tag{26}$$

It is interesting to note that the maximum power output is obtained when the external impedance seen by the tunnel diode consists of a resistance and an inductive reactance which have the same value.

Using these results, we calculate a specific example of a tunnel diode operating at 10 kMc. R_s is assumed to be low enough to give a resistive cutoff of 30 kMc. For $V=0.15$ volt and with the aid of (19) and (20), we obtain the optimum peak current-to-capacitance ratio, $k_0=9.4$ ma/$\mu\mu$f and the maximum power-inductance product $P_{\max}\,L=0.0224$ mw-mμh. If the inductance is assumed as low as 0.025 mμh, the power output is 0.896 mw. The significance of a low inductance for achieving a useful power output is quite evident. According to (26), the cavity impedance seen by the junction for this case turns out to be $1.57+j1.57$ ohms. Such a low impedance can be realized with the diode in the thin, center section of a tapered cavity operating at higher-order modes.[1]

For lower-power applications of up to 100 μw at 10 kMc, a plain cylindrical cavity in a coaxial arrangement[2] gives an inductance as low as 0.25 mμh. This value is constant up to 18 kMc, and the simple equivalent circuit of Fig. 2 has been shown to hold under the following conditions:[7]

1) Inner radius of outer conductor of cavity <60 mils,
2) Radius of inner conductor >2 mils,
3) Thickness of cavity <57 mils,
4) Diameter of semiconductor junction <0.4 mil.

Practical, stable and self-contained devices have been built to meet these conditions. If these devices are operated in higher-order mode cavities instead of in the simple cylindrical cavity referred to previously, the inductance seen by the diode becomes smaller. Consequently, the junction size and, therefore, the capacity and peak current can be made larger to increase the power output.

B. Power in the Nonlinear Range

At low enough frequencies the oscillations extend into the nonlinear region. This will happen when the diode is operating much below its resistive cutoff frequency (12), and when the load is smaller than the R_L value determined by (23). The evaluation of the nonlinear case is a complex problem which is not treated here; however, some qualitative considerations should be noted.

When the load is very small or zero, the oscillations extend into the valley region, and partly into the positive resistance regions. The fundamental sinusoidal component of these nonlinear oscillations can still be described by an equivalent negative resistance R_e that takes the place of R in (23). R_e will be larger than R, but the exact functional relationship between R and R_e cannot be obtained by simple averaging. As the load is increased, the voltage swing of the nonlinear oscillations begins to contract, R_e decreases, and the power developed in the load begins to increase.

An approximate estimate of the maximum of nonlinear power compared to the maximum power in the linear range can be made by a simple computation. The total power available in the entire negative-resistance range at dc is compared with the power available in the linear range defined in Section I. For a typical GaAs diode shown in Fig. 1, the area under the negative-resistance region extending from the peak to the valley gives 261 μw. This is the total available power in the tunnel diode. The power in the triangle defined by V and I is 96 μw, or 37 per cent of the total available power.

In Fig. 1 the minimum value of R is 101.5 ohms and $V/I=114.5$ ohms for $V=0.15$ volt. Thus, the variation of R is 12.8 per cent. If we allow R to vary up to 47 per cent, we obtain $V'=0.18$ volt and the power in the triangle $V'I'$ is 50 per cent of the total available power. From these data one sees that the maximum linear power is a significant part of the total possible power output. Moreover, only part of the total power of nonlinear oscillators can be utilized in many cases due to the higher content of undesired frequencies. It should be noted that for $\omega \leq \omega_{R_s}/3$, a 15 per cent variation in R results in the same variation in P, according to (15), since $dR/R=-dP/P$ for constant V. Thus, the nonlinearity error in the maximum power estimates is also limited to 15 per cent in our calculations.

V. Conclusions

Equations for the power output of high-frequency tunnel diodes have been derived. The treatment is limited to the linear range of operation of a tunnel diode.

[7] G. Dermit, Tech. Memo. (TM 366.1); October 7, 1960. Unpublished.

The linear range is defined by the condition that the deviation of the average negative resistance V/I from the minimum negative resistance value R is small.

The linear range is of interest in the following three cases:

1) Amplifier applications where signal distortions are not acceptable. The equations of Section II and the charts can be applied directly for estimating the upper limit of the undistorted power level that can be handled by the tunnel diode within its frequency range.

2) Low-power oscillators where the emphasis is on achieving a high frequency of oscillation close to the resistive cutoff frequency of the diode. In this case, the oscillations are only possible within the linear region.

3) High-frequency oscillators where the diode oscillates at a frequency considerably below resistive cutoff. Here, both linear as well as nonlinear oscillations are possible. The conditions and power output are discussed in Section IV.

An important result is that in the derived equations the power output of a tunnel-diode oscillator and the series inductance appear as a product. Consequently, in all applications the role of inductance is a very dominant one. Lowering of the inductance means increasing the power, but low inductances also lead to very low impedances. This presents special problems of coupling the input and the output. For instance, the cavity impedance seen by the diode must be about 1.6 ohm in order to achieve about 1 mw output at a frequency of 10 kMc.

The linear voltage swing V has a drastic influence on power also. The maximum power-inductance product at a given frequency is proportional to the square of V. Another important conclusion is that tunnel-diode oscillators operating in the linear range give maximum power output only for an optimum value of the junction time constant or its k value. These maxima are further improved by reducing the series resistance of the junction to the point where the resistive cutoff frequency becomes approximately three times the operating frequency.

It has been shown that a significant part of the total possible power output of a high-frequency tunnel-diode oscillator can be realized in the linear region.

Appendix I

Extension to Small I_p/I_v Ratio Diodes

In Section II the linear portion of the current swing I was assumed equal to the peak-current I_p. Tunnel diodes with small peak-to-valley current ratios are useful in amplifier and oscillator applications. To include the effect of small ratios and the effect of nonlinearity through m_i in (1), we have only to modify (16) into

$$R = \frac{V}{kC(1 - 1/n)m_i}, \tag{27}$$

where $n = I_p/I_v$. Thus, all the k's in the text have to be multiplied by $(1-1/n)m_i$ whenever this factor is appreciably smaller than one.

Appendix II

Alternate Derivation of Conditions for Steady-State Oscillation

The results of (4), (5) and (23) can also be obtained by solving the circuit differential equation for the tunnel diode. This solution is quite general and includes all the possible states of operation, such as transients and blocking oscillations, as well as steady sinusoidal states. The complete solution of the differential equation is[3]

$$i = A_1 e^{p_1 t} + A_2 e^{p_2 t} + B, \tag{28}$$

where i is the total instantaneous current at time t; A_1, A_2 and B are constants depending on initial conditions; and p_1, p_2 are given by

$$p_{1,2} = \frac{1}{2}\left(\frac{1}{RC} - \frac{R_s}{L}\right) \pm j\sqrt{\frac{1}{LC}\left(1 - \frac{R_s}{R}\right) - \frac{1}{4}\left(\frac{R_s}{L} - \frac{1}{RC}\right)^2} \tag{29}$$

In (29), R_L has been lumped into R_s because R_s and R_L occur in series. When the two p's are equal, they are real, and a transient undergoes an exponential growth or decay; when they are complex, we have exponentially decaying or growing sinusoids; and when they are imaginary we have steady-state oscillation. For the imaginary case, (23) must hold. Using this relation, and eliminating R_s under the square root of (29), we arrive at the original oscillation condition given by (5).

The two methods clearly give the same results. The impedance given by (2) is just a short-cut to the specialized steady-state sinusoidal solution of the differential equation and excludes the more general transient states.

Appendix III

General Expression for Power

By removing the condition that $R \gg R_s$ in (13) and combining (9) and (12), (15) becomes

$$P = \frac{V^2}{8R^2}(R - R_s)\left(1 - \frac{\omega^2}{\omega_{R_s}{}^2}\right). \tag{30}$$

It is clear that when $R = R_s$, the power is zero; similarly, $\omega_{R_s} = 0$ from (12), and therefore, $\omega = 0$. Obviously (13), (15), and (18) fail when R approaches R_s, but these expressions are quite good when R is more than 5 times larger than R_s. The condition $R \gg R_s$ can be made general by multiplying both sides by C. Making use of (16), we have

$$\frac{V}{k} \gg R_s C. \tag{31}$$

This is the necessary condition for (15) and (18) to hold. In Section III, it was shown that R_sC is independent of the capacitance and negative resistance product RC of the junction. For severely-etched junctions, a practical value for R_sC is about 6 ohm-$\mu\mu$f, and for unetched junctions, such as large area junctions or small point-contact junctions, it is estimated to be about 0.6 ohm-$\mu\mu$f. Thus, for unetched-type junctions, (15) and (18) would hold up to very high values of k (*viz.*, up to 32 ma/$\mu\mu$f for Ge, see Section III-A and 50 ma/$\mu\mu$f for GaAs, see Section III-B).

By subjecting (30) to condition (5) and making use of (16), we obtain the general power expression for pure sinusoidal oscillations.

$$P = \frac{V}{8L}\left\{\frac{k}{\omega^2 + \dfrac{k^2}{V^2}} - (R_sC)V\right\}. \tag{32}$$

Eq. (32) can be plotted in the manner of the tunnel diode charts of Section III. Since R_sC and V are independent of the doping, we see immediately that, at a fixed frequency of operation ω_0, there is an optimum value of k given by $\omega_0 = k_0/V$ that leads to maximum power. This is in agreement with (19), which is thus quite general and rigorous. The maximum power is

$$P_{\max} = \frac{V^2}{8L}\left(\frac{V}{2k_0} - R_sC\right). \tag{33}$$

This becomes equal to (20) when $V/2k_0 \gg R_sC$, or when $R/2 \gg R_s$.

The power expression (32) is based upon three assumptions which are quite valid for sinusoidal oscillations. The first assumption is that tunnel diodes can be made to oscillate with a voltage amplitude extending beyond the linear range in Fig. 1. It is always possible to change R_L in (23) and thereby to increase the voltage swing from V to a value such as V', if R_s is low enough.

The second assumption is that R_sC is independent of the resistivity of the p- and n-type regions of the junction. This property, which was shown to be true in Section III, is quite basic in deriving the conclusions on the maximum power and optimum k. It implies that R_sC is independent of RC, despite the fact that R_s by itself depends on both the area and the resistivities.

The third assumption is that L can be varied without affecting the area and the other parameters of the junction. Since we are mainly considering coaxial cavities with the diode occupying a very thin section in the center conductor, it is possible to substitute greatly varying junctions in the cavity without changing L.

The frequency at which (32) becomes zero is

$$\omega^2 = \frac{k}{R_sCV} - \frac{k^2}{V^2}, \tag{34}$$

which is simply the resistive cutoff frequency (12), obtained by substituting for (16). Following the assumption of independent R_sC, we can now compute from (34) the maximum possible cutoff frequency of tunnel diodes.

$$\omega_m = 1/2R_sC, \tag{35}$$

which occurs for the optimum time constant

$$RC = 2R_sC. \tag{36}$$

Observe that for time constants smaller than (36) the cutoff frequency will decrease. If $R_sC = 0.6$ ohm-$\mu\mu$f, the maximum frequency of oscillation is 130 kMc!

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Design Theory of Optimum Negative-Resistance Amplifiers*

E. S. KUH†, MEMBER, IRE, AND J. D. PATTERSON†

Summary—In this paper we consider general amplifiers obtained by imbedding a linear active 1-port device in arbitrary 3-ports. The active device is assumed to have a representation of a negative conductance $-G_D$ in parallel with a parasitic capacitance C_D. We prove that for the lossless reciprocal imbedding, the transducer voltage gain is limited by

$$|S_{21}| < \tfrac{1}{2}(1 + e^{\pi G_D/\omega_0 C_D}),$$

where ω_0 is the angular bandwidth. For arbitrary passive imbedding,

$$|S_{21}| \leq e^{\pi G_D/\omega_0 C_D}.$$

Synthesis methods to approach or achieve the optimum are presented.

Other types of amplifier configurations are next considered. In each case, the optimum gain-bandwidth formula, synthesis procedure and some useful design curves are given.

I. Introduction

The recent development of new 1-port active devices has given amplifier designers a new challenge, *i.e.*, the limitation of the given device in terms of maximum gain-bandwidth product and synthesis procedures for an optimum amplifier. It is obvious that an ideal negative resistance can provide infinite gain-bandwidth. The limitation of practical active devices is due to parasitic elements. In the case of conventional amplifiers employing 2-port active devices, such as vacuum tubes, the maximum obtainable gain-bandwidth product was first derived by Bode.[1] Bode also gave synthesis methods for optimum passive networks which act as input, output or interstage coupling networks.

Tunnel diodes and degenerate parametric amplifiers have often been used as the active elements to design simple amplifiers.[2–5] However, no general theory has been found for the limitations of the devices and optimum synthesis procedures. In this paper, it is assumed that the 1-port active device has an equivalent representation of a negative conductance $-G_D$ in parallel with a capacitance C_D. We first consider the problem where the active device is imbedded in a 3-port lossless reciprocal network. The maximum obtainable gain-bandwidth and a synthesis procedure for an optimum amplifier are presented. We then consider other amplifier configurations. In each case the maximum gain-bandwidth and synthesis procedure are given.

A problem which is not considered in this paper is the reflection coefficients at the input and output of the amplifiers. It will be noted that except for the matched circulator and hybrid types, all amplifiers seem to have large reflection coefficients.

II. Lossless, Reciprocal Imbedding

The most general amplifier employing a 1-port active device is illustrated in Fig. 1, where G_S and G_L are source and load conductances. For obvious reasons, we wish to restrict the passive 3-port to be reciprocal and lossless. The network is then redrawn as shown in Fig. 2. N_a is a

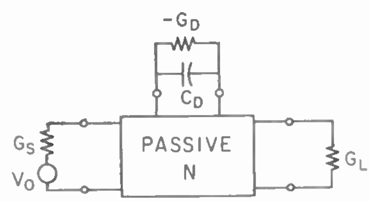

Fig. 1—Passive 3-port imbedding of an active 1-port.

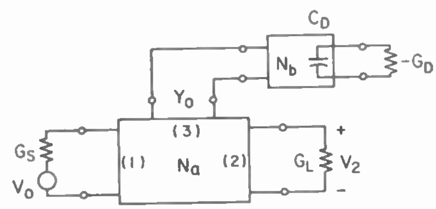

Fig. 2—Lossless, reciprocal imbedding of an active device.

lossless reciprocal 3-port described by its normalized scattering matrix with respect to a set of reference admittances: G_S, G_L, and Y_0, where Y_0 is arbitrary.[6] N_b is a lossless reciprocal 2-port with reference admittances, Y_0 and $-G_D$. The parasitic capacitance C_D is now included in N_b. With this representation, we can express the over-all transmission coefficient in terms of the

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scattering coefficients of N_a and N_b as follows:

$$S_{21} = \frac{2V_2}{V_0}\sqrt{\frac{G_L}{G_S}} = S_{21a} + \frac{S_{23a}S_{31a}S_{11b}}{1 - S_{33a}S_{11b}}. \tag{1}$$

Since the reference admittance Y_0 is not involved in the expression, we can arbitrarily choose $Y_0 = Y_{33}$, where Y_{33} is the input admittance of N_a at port (3) with terminations G_S and G_L. Thus,

$$S_{33a} = 0, \tag{2}$$

and (1) is simplified without losing any generality.

$$S_{21} = S_{21a} + S_{23a}S_{31a}S_{11b}. \tag{3}$$

The physical interpretation of (3) is clear. The overall transmission is equal to the sum of two parts. The first part S_{21a} represents the direct transmission of N_a from port (1) to port (2). The second part represents the product of three terms: the direct transmissions from port (1) to port (3) and port (3) to port (2), and the reflection at port (3) due to mismatch of N_b. Since S_{21a}, S_{31a}, and S_{23a} are the scattering coefficients of a lossless network with passive reference admittances, their magnitudes cannot be larger than unity. The only term which could contribute gain to this expression is S_{11b}. In the following we will consider separately N_a and N_b and derive the optimum gain-bandwidth formula, which can be approached when both N_a and N_b are optimized.

N_a is a lossless reciprocal 3-port described by its normalized scattering matrix S_a with reference admittances G_S, G_L, and Y_0. Without losing any generality we can assume that $Y_0 = Y_{33}$ is real at one frequency. This is explained in Fig. 3, where a tuned circuit is inserted between N_a and N_b. Clearly, at the resonant frequency, the complete network is not changed by this insertion, yet Y_{33} can now be made purely real.[7] With real reference admittances, S_a is unitary. In addition, since $S_{33a} = 0$, we have

$$|S_{31a}|^2 + |S_{32a}|^2 = 1, \tag{4}$$

$$|S_{11a}|^2 + |S_{21a}|^2 + |S_{31a}|^2 = 1, \tag{5}$$

and

$$S_{31a}\bar{S}_{11a} + S_{32a}\bar{S}_{12a} = 0, \tag{6}$$

where the bars ($^-$) designate the complex conjugate. Moreover, since N_a is reciprocal,

$$S_{23a} = S_{32a} \quad \text{and} \quad S_{21a} = S_{12a}. \tag{7}$$

Combining the above four equations and eliminating $|S_{11a}|$, we obtain the following useful result:

$$|S_{21a}| = |S_{23a}S_{31a}|. \tag{8}$$

Eq. (8) is now substituted in (3). The over-all transducer voltage gain is

$$|S_{21}| = |S_{23a}S_{31a}|\,|1 + S_{11b}e^{j\phi}|, \tag{9}$$

where ϕ is a phase angle. Clearly, we wish to maximize $|S_{23a}S_{31a}|$ to obtain the optimum $|S_{21}|$. From (4), it is seen that the maximum is realized if

$$|S_{23a}| = |S_{31a}| = \frac{1}{\sqrt{2}}. \tag{10}$$

and the scattering matrix for an optimum N_a with all real coefficients is found as

$$S_a = \begin{bmatrix} -\frac{1}{2} & \frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{2} & -\frac{1}{2} & \frac{1}{\sqrt{2}} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 \end{bmatrix}. \tag{11}$$

The complete network for $G_S = G_L = 1$ and $Y_0 = 2$ is shown in Fig. 4. From (3) and (11), we conclude that

$$S_{21} = \tfrac{1}{2}(1 + S_{11b}). \tag{12}$$

The problem of finding the maximum gain-bandwidth product for given C_D and $-G_D$ is reduced to the problem of finding the optimum 2-port N_b such that S_{11b} is the maximum for a specified bandwidth. For this, we refer to Fig. 5, where N_b is shown with terminating con-

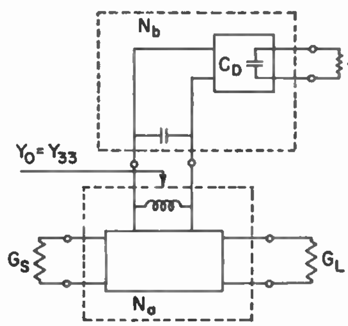

Fig. 3—Insertion of a tuned circuit between N_a and N_b.

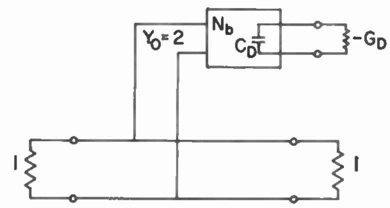

Fig. 4—An optimum network.

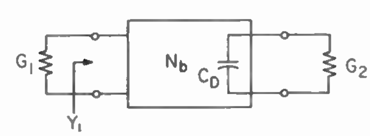

Fig. 5—Lossless 2-port.

[7] Note that the real part of Y_{33} cannot be zero if S_{23a} and S_{31a} are finite. A lossless balanced bridge will illustrate this particular case.

ductances G_1 and G_2. Let ρ_1 be the reflection coefficient at the input for an output termination $G_2=+G_D$ and ρ_1' be the reflection coefficient at the input for $G_2=-G_D$. It is shown in Appendix I that

$$|S_{11b}| = |\rho_1'| = \frac{1}{|\rho_1|}. \tag{13}$$

Thus, the problem of finding the maximum $|S_{11b}|$ over a given band is reduced to the problem of finding minimum $|\rho_1|$ if the output termination is a positive conductance G_D. This is the familiar lossless matching problem and the solution is well known for a GC load. From Bode,

$$\int_{\omega_1}^{\omega_2} \log \frac{1}{|\rho_1|}\, d\omega \le \pi \frac{G_D}{C_D} \tag{14}$$

or

$$\int_{\omega_1}^{\omega_2} \log |S_{11b}|\, d\omega \le \pi \frac{G_D}{C_D}. \tag{15}$$

For a specified bandwidth ω_0, the constant gain is limited by

$$|S_{11b}| \le e^{\pi G_D/\omega_0 C_D}. \tag{16}$$

Substituting in (12), we obtain the final result

$$|S_{21}| < \tfrac{1}{2}(1 + e^{\pi G_D/\omega_0 C_D}). \tag{17}$$

This expression gives the upper bound of the gain obtainable for a given bandwidth by imbedding the active 1-port in a lossless reciprocal 3-port. The treatment in Section IV will indicate that the upper bound cannot be attained over a band through ω_0. This is because S_{11b} of the resulting network has a phase even if infinite number of elements are allowed. However, for a high-gain amplifier the upper bound can essentially be approached, since the second term in the parenthesis of (17) is much larger than unity. Since the amplification of such an amplifier is obtained through reflection, we refer to this as the reflection-type amplifier. The synthesis procedure of the matching network and some design formulas are given in Section IV after we consider two other similar configurations in the next section.

III. The Circulator- and Hybrid-Type Amplifiers

In this section, we analyze two familiar types of amplifiers. The first employs a nonreciprocal passive 3-port.[2,5] The second one uses two active 1-ports.[4] The gain-bandwidth performance is better than the reflection type discussed in the previous section.

A. The Circulator-Type Amplifier

We refer again to Fig. 2, but remove the requirement of reciprocity on N_a. From (3) it is clear that in order to optimize S_{21}, the product $|S_{23a}S_{31a}|$ has to be maximized. An ideal circulator has the following normalized scattering matrix:

$$S = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ 1 & 0 & 0 \end{bmatrix}. \tag{18}$$

Thus, $S_{23a}S_{31a}=1$, which is the maximum that can be obtained from a passive network. Hence, the network as shown in Fig. 6 gives an optimum amplifier configuration. The transmission coefficient is equal to the reflection coefficient of N_b.

$$S_{21} = S_{11b}. \tag{19}$$

For a given ω_0,

$$|S_{21}| \le e^{\pi G_D/\omega_0 C_D}. \tag{20}$$

This is about 6 db better than the reflection-type amplifier. However, the circulator-type amplifier is a strictly one-way amplifier in contrast to the two-way amplifier of the reflection type.

B. The Hybrid-Type Amplifier

To avoid the use of nonreciprocal elements, to provide two-way amplification and to achieve the gain level of (20), we need to use two active 1-ports. This is shown in Fig. 7. N_a is a lossless hybrid which has a scattering matrix

$$S_a = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 & 0 & 1 & 1 \\ 0 & 0 & 1 & -1 \\ 1 & 1 & 0 & 0 \\ 1 & -1 & 0 & 0 \end{bmatrix}. \tag{21}$$

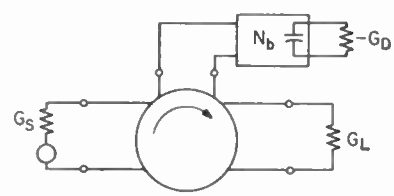

Fig. 6—The circulator-type amplifier.

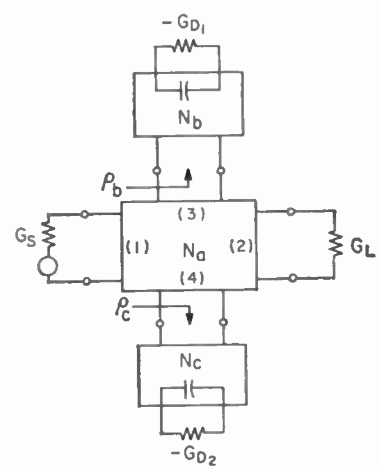

Fig. 7—The hybrid-type amplifier.

Networks N_b and N_c are lossless 2-ports; each is terminated by an active 1-port device. Let ρ_b and ρ_c be the reflection coefficients at the inputs of N_b and N_c as shown. The resultiug scattering matrix of the over-all 2-port is

$$S = \frac{1}{2}\begin{bmatrix} \rho_b + \rho_c & \rho_b - \rho_c \\ \rho_b - \rho_c & \rho_b + \rho_c \end{bmatrix}. \tag{22}$$

If we choose

$$\rho_c = -\rho_b, \tag{23}$$

then

$$S_{21} = S_{12} = \rho_b. \tag{24}$$

Thus, we have a two-way amplifier with exactly the same gain-bandwidth limitation as the circulator type. The requirement of (23) indicates that the input admittance of N_b must be the reciprocal of that of N_c. Since the two 1-port active devices have the same equivalent circuit, the dual network concept cannot be used to obtain the reciprocal admittance. To realize approximately the condition of (23), a broad-band 90° phase-shift network can be used in one of the two networks.

IV. Synthesis of Matching Networks

In order to achieve or approach the optimum gain-bandwidth, we need to design an optimum matching network. The matching problem is a familiar one in network theory and has been solved by Bode, Fano and others.[8] In our problem, since the load is a parallel GC, exact synthesis procedure can be carried out to accomplish either a maximally-flat magnitude or an equal ripple match. The following simple derivation for the general maximally-flat match is presented along with some useful design formulas.

We refer again to Fig. 5. Let

$$G_1 = 1 \quad \text{and} \quad G_2 = \frac{1-\epsilon^n}{1+\epsilon^n}, \tag{25}$$

where $\epsilon < 1$ is a positive design parameter. We denote t and ρ as the normalized transmission and reflection coefficients of the lossless 2-port with reference to G_1 and G_2. For a maximally flat match,

$$|t|^2 = \frac{1-\epsilon^{2n}}{1+\omega^{2n}} \tag{26}$$

and

$$|\rho|^2 = \frac{\epsilon^{2n}+\omega^{2n}}{1+\omega^{2n}}. \tag{27}$$

Since the lossless ladder-matching network is transparent at dc, $G_2 = (1-\epsilon^n/1+\epsilon^n)$ follows from (26) and (27).

The transducer voltage gain of the over-all amplifier is equal to $1/|\rho|$ for the circulator and hybrid types and is equal to approximately $1/2|\rho|$ for the reflection type. In this section, we designate $|S_{21}|$ as the gain of the circulator or hybrid type; thus, $S_{21} = 1/\rho$. From Eq. (27) the dc gain is

$$S_{21}(0) = \frac{1}{\epsilon^n}, \tag{28}$$

and the 3-db bandwidth is found at $S_{21} = 1/\sqrt{2}\epsilon^n$ or

$$\omega_1 = \frac{\epsilon}{(1-2\epsilon^{2n})^{1/2n}}. \tag{29}$$

The complete network can be obtained in the following way: First determine the reflection coefficient from (27):

$$\rho(s)\rho(-s)\big|_{s=j\omega} = |\rho(j\omega)|^2 = \frac{\omega^{2n}+\epsilon^{2n}}{1+\omega^{2n}}. \tag{30}$$

The poles of $\rho(s)$ are restricted to be in the left half plane for stability reasons. The zeros of $\rho(s)$ for maximum gain-bandwidth are restricted to be in the right-half plane.[9] Once $\rho(s)$ is determined, the input admittance $Y_1(s)$ can be found.

$$Y_1 = \frac{1-\rho}{1+\rho}. \tag{31}$$

A low-pass ladder can be developed based on Cauer's continued fraction expansion. For the specified G_2 in Eq. (25) the last shunt capacitance is given by[10]

$$C_n = G_2 \frac{2\sin(\pi/2n)}{1-\epsilon}. \tag{32}$$

From (29) and (32), the actual 3-db bandwidth for a given set of C_D and G_D can be found by frequency and admittance denormalization.

$$\omega_{3\text{-db}} = \frac{\omega_1 C_n}{C_D}\frac{G_D}{G_2} = \frac{\epsilon}{(1-\epsilon^{2n})^{1/2n}}\frac{2\sin(\pi/2n)}{1-\epsilon}\frac{G_D}{C_D}. \tag{33}$$

The asymptotic behavior of $\omega_{3\text{-db}}$ as $n\to\infty$ with a constant dc gain can be found by letting ϵ^n fixed and $\epsilon\to 1$:

$$\omega_{3\text{-db}} \to \frac{\pi}{\ln S_{21}(0)}\frac{G_D}{C_D}, \tag{34}$$

which is the optimum gain-bandwidth given by (20). The network for $n=3$ is shown in Fig. 8. For design convenience, the normalized angular bandwidth $\omega/(G_D/C_D)$

[8] R. M. Fano, "Theoretical limitations on the broadband matching of arbitrary impedances," *J. Franklin Inst.*, vol. 249, pp. 57–83, 139–154; January, February, 1950.

[9] From Bode,[1] the zeros of the reflection coefficient at the port where the capacitance appears should be in the left half plane. Thus, zeros of $\rho(s)$, which is the reflection coefficient at the input port, should be in the right half plane.

[10] L. Weinberg and P. Slepian, "Takahasi's results on Tchebycheff and Butterworth ladder networks," IRE Trans. on Circuit Theory, vol. CT-7, pp. 88–101; June, 1960.

is plotted vs the transducer gain in Fig. 9 for $n=1, 2, 3, 4$ and ∞. The gain for the reflection type is about 6 db less. It should be pointed out that exact element values of the ladder network for both the maximally-flat and the equal-ripple matching networks can also be calculated from formulas given by Weinberg and Slepian.[10]

V. Other Amplifier Configurations

In this section, we consider three more amplifier configurations. For each configuration, we derive the optimum gain-bandwidth and give the synthesis procedure.

A. The Direct-Connected-Type Amplifier

The simplest amplifier configuration is shown in Fig. 10, where Y_1 is a 1-port passive network. Chang uses a single inductance as the 1-port passive network in his tunnel-diode amplifier.[3] The transmission coefficient can be expressed by

$$S_{21} = 2\sqrt{G_S G_L}\, Z, \tag{35}$$

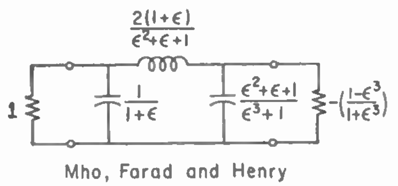

Fig. 8—Maximally-flat magnitude matching network.

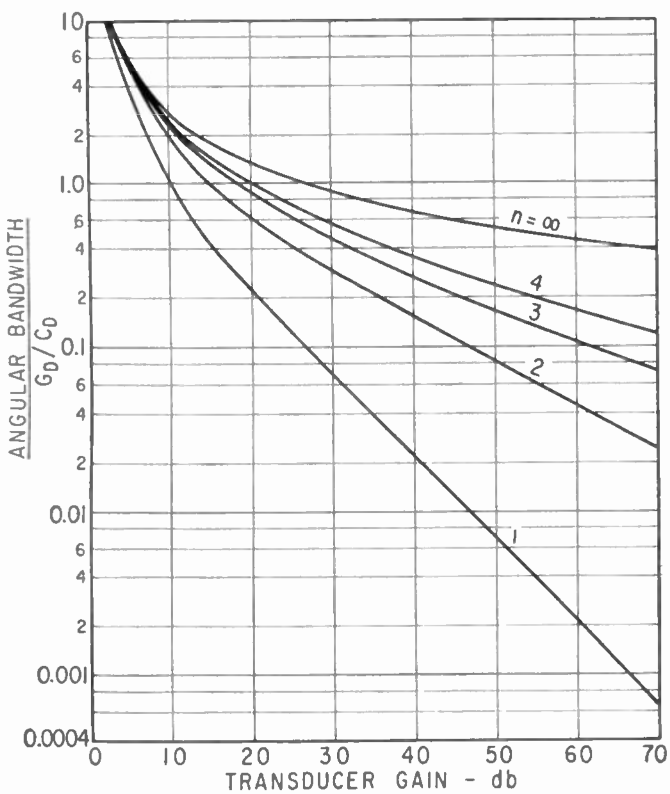

Fig. 9—Gain-bandwidth performance of the circulator-type and hybrid-type amplifiers.

where

$$1/Z = G_S + G_L - G_D + sC_D + Y_1. \tag{36}$$

The problem of optimum gain-bandwidth is similar to Bode's problem of the 2-terminal interstage. Using a similar approach, we find that for a given bandwidth ω_0, the maximum constant gain is given by

$$S_{21} = \frac{4\sqrt{G_S G_L}}{\omega_0 C_D} \frac{1}{\epsilon/\omega_0 C_D + \sqrt{1 + (\epsilon/\omega_0 C_D)^2}}, \tag{37}$$

where

$$\epsilon = G_S + G_L - G_D.$$

For $G_S = G_L = G_D/2$, $\epsilon = 0$, the maximum gain of the amplifier is given by

$$S_{21} = \frac{2G_D}{\omega_0 C_D}. \tag{38}$$

The required Y_1 is the same as that given by Bode, that is, the input admittance of a constant-K-type image filter. The circuit is shown in Fig. 11.

B. The Transmission-Type Amplifier

This configuration was first used by Sard.[5] As shown in Fig. 12, the load and the source are separated by a coupling network N, which contains the parasitic capacitance C_D as the final element. The other configuration, where the active 1-port is at the input end as shown in Fig. 13, can be treated identically.

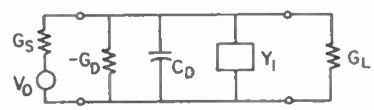

Fig. 10—The direct-connected-type amplifier.

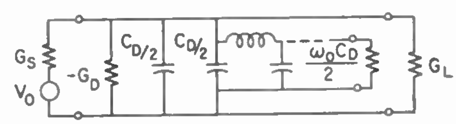

Fig. 11—The optimum direct-connected-type amplifier.

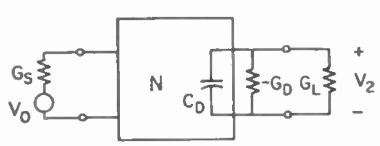

Fig. 12—The transmission-type amplifier.

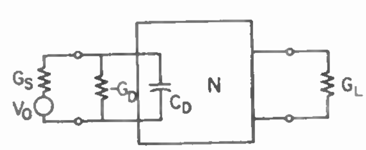

Fig. 13—The transmission-type amplifier.

It is assumed that $G_L - G_D < 0$; hence, the over-all output conductance in Fig. 12 is negative. As shown in Appendix II, the transmission coefficient with load conductance G_L can be written as

$$S_{21} = \sqrt{\frac{G_L}{G_D - G_L}} \, \frac{t}{\rho}, \tag{39}$$

where t and ρ are, respectively, the transmission and reflection coefficients of the same network N with source and load conductances, G_S and $G_D - G_L$. Since

$$\frac{1}{|\rho|} \leq e^{\pi(G_D - G_L)/\omega_0 C_D} \tag{40}$$

and

$$|t|^2 = 1 - |\rho|^2, \tag{41}$$

it is found that for a given ω_0,

$$|S_{21}| \leq \sqrt{\frac{G_L}{G_D - G_L}} \left(e^{2\pi(G_D - G_L)/\omega_0 C_D} - 1\right)^{1/2}. \tag{42}$$

It is seen that the gain limitation is a function of the load conductance. For a maximally-flat magnitude design, let

$$|t|^2 = \frac{1 - \epsilon^{2n}}{1 + \omega^{2n}} \tag{43}$$

and

$$|\rho|^2 = \frac{\epsilon^{2n} + \omega^{2n}}{1 + \omega^{2n}} \,. \tag{44}$$

The transducer power gain is

$$|S_{21}|^2 = \frac{G_L}{G_D - G_L} \, \frac{1 - \epsilon^{2n}}{\epsilon^{2n} + \omega^{2n}} \,. \tag{45}$$

For the normalized conductances

$$G_S = 1, \qquad G_D - G_L = \frac{1 + \epsilon^n}{1 + \epsilon^n} \,. \tag{46}$$

$$|S_{21}|^2 = G_L \frac{(1+\epsilon^n)^2}{\epsilon^{2n} + \omega^{2n}} \tag{47}$$

The dc gain and the 3-db bandwidth are

$$S_{21}(0) = \sqrt{G_L} \, \frac{1 + \epsilon^n}{\epsilon^n}; \tag{48}$$

$$\omega_{3\text{-db}} = \frac{2\epsilon \sin(\pi/2n)}{1 - \epsilon} \, \frac{G_D - G_L}{C_D} \,. \tag{49}$$

Thus, the gain-bandwidth is a function of G_L. Sard has shown that the optimum choice of G_L is given by

$$G_L = G_D \frac{1 - \epsilon^{2n}}{2n(1 - \epsilon)} \,. \tag{50}$$

The design curves for such a load are shown in Fig. 14.

C. The Cascade-Type Amplifier

This configuration is shown in Fig. 15, where the negative conductance is separated from the source and the load by two coupling networks N_a and N_b. As shown in Appendix III, the over-all transmission coefficient can be expressed in terms of the normalized scattering parameters of N_a and N_b:

$$S_{21} = \frac{2S_{21a}S_{21b}}{2 - G - G(S_{22a} + S_{11b}) - (2 + G)S_{22a}S_{11b}} \,. \tag{51}$$

If we arbitrarily set $G = 2$ and $S_{22a} = -S_{11b}$, (51) becomes

$$S_{21} = -\frac{S_{21a}S_{21b}}{2S_{22a}S_{11b}} \,. \tag{52}$$

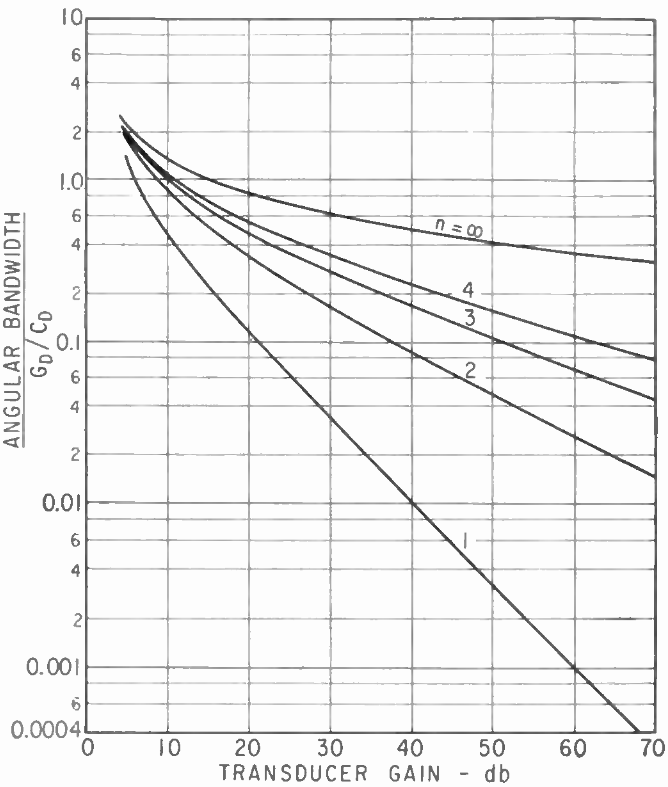

Fig. 14—Gain-bandwidth performance of the optimum transmission-type amplifier.

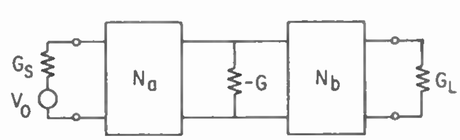

Fig. 15—The cascade-type amplifier.

The equation is similar to (39) of the previous case. The gain limitation is found to be

$$|S_{21}| \leq \tfrac{1}{2}(e^{\pi G_D/\omega_0 C_D} - 1), \tag{53}$$

which approaches the optimum case for the reciprocal lossless 3-port imbedding as given by (17). The requirement of $S_{22a} = -S_{11b}$ indicates that N_a is the dual of N_b. For a maximally-flat design,

$$|S_{21}| = \frac{1}{2}\,\frac{1-\epsilon^{2n}}{\epsilon^{2n}+\omega^{2n}}. \tag{54}$$

The dc gain and the 3-db bandwidth are

$$S_{21}(0) = \frac{1}{2}\,\frac{1-\epsilon^{2n}}{\epsilon^{2n}}; \tag{55}$$

$$\omega_{3-db} = (\sqrt{2}-1)^{1/2n}\,\frac{\epsilon}{1-\epsilon}\,\sin(\pi/2n)\,\frac{G_D}{C_D}. \tag{56}$$

As n approaches infinity while $S_{21}(0)$ is fixed, (56) approaches (53), the limiting case. The network for $n=2$ is shown in Fig. 16 below. The design curves are given in Fig. 17.

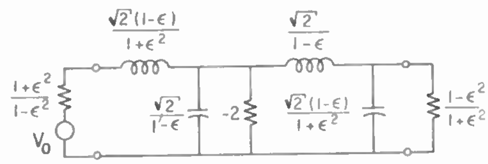

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Fig. 16—A maximally-flat cascade-type amplifier with $n=2$.

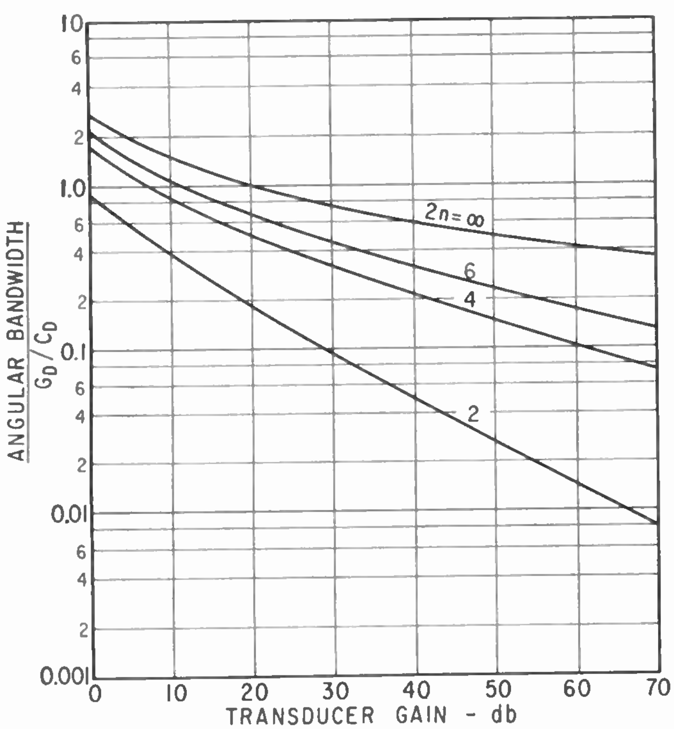

Fig. 17—Gain-bandwidth performance of the cascade-type amplifier.

Appendix I

With reference to Fig. 5, let Y_1 be the input admittance of N_b with $G_2 = +1$:

$$Y_1 = y_{11}\,\frac{1/z_{22}+G_2}{y_{22}+G_2} = \frac{m_1+n_1}{m_2+n_2}. \tag{57}$$

where y_{11} and y_{22} are the short-circuit driving-point admittances, z_{22} is the open-circuit driving-point impedance at port (2), and m's and n's are, respectively, the even and odd polynomials of the complex frequency s. The short-circuit admittances and the open-circuit impedance can be written in either of the following forms.[11]

$$\begin{cases} y_{11} = \dfrac{m_1}{n_2} \\ y_{22} = \dfrac{m_2}{n_2} \\ z_{22} = \dfrac{m_1}{n_1} \end{cases} \quad \text{or} \quad \begin{cases} y_{11} = \dfrac{n_1}{m_2} \\ y_{22} = \dfrac{n_2}{m_2} \\ z_{22} = \dfrac{n_1}{m_1} \end{cases}. \tag{58}$$

Let Y_1' be the input admittance of N_b with $G_2 = -1$, then

$$Y_1' = y_{11}\,\frac{1/z_{22}-G_2}{y_{22}-G_2} = \frac{m_1-n_1}{n_2-m_2}. \tag{59}$$

Thus,

$$\rho_1 = \frac{1-Y_1}{1+Y_1} = \frac{m_2-m_1+n_2-n_1}{m_2+m_1+n_2+n_1} \tag{60}$$

and

$$\rho_1' = \frac{1-Y_1'}{1+Y_1'} = \frac{-(m_2+m_1)+n_1+n_2}{m_1-m_2+n_2-n_1}. \tag{61}$$

Let $s = j\omega$:

$$|\rho_1'|^2 = \frac{1}{|\rho|^2}. \tag{62}$$

Appendix II

We refer to Appendix I and let t and t' be the transmission coefficients of N_b with $G_2 = +1$ and -1, respectively. In terms of the m's and n's,

$$t = \frac{2\sqrt{m_1m_2 - n_1n_2}}{m_1+m_2+n_1+n_2} \tag{63}$$

[11] E. A. Guillemin, "Synthesis of Passive Networks," John Wiley and Sons, Inc., New York, N. Y.; 1957.

and

$$t' = \frac{2\sqrt{n_1 n_2 - m_1 m_2}}{m_1 - m_2 - n_1 + n_2}. \tag{64}$$

Let $s = j\omega$:

$$t'(j\omega) = j\frac{t(j\omega)}{\rho(j\omega)}. \tag{65}$$

We therefore conclude from Fig. 12 that the transmission coefficient t' for the network with a load admittance $G_L - G_D$ can be expressed in terms of the transmission coefficient t and the reflection coefficient ρ with a load admittance $G_D - G_L$. Since the actual load is G_L,

$$\begin{aligned} S_{21} &= \frac{2V_2}{V_0}\sqrt{\frac{G_L}{G_S}} = \frac{2V_2}{V_0}\sqrt{\frac{G_L - G_D}{G_S}}\sqrt{\frac{G_L}{G_L - G_D}} \\ &= t'\sqrt{\frac{G_L}{G_L - G_D}} = \sqrt{\frac{G_L}{G_D - G_L}}\,\frac{t}{\rho}. \end{aligned} \tag{66}$$

Appendix III

Eq. (51) can be derived most easily if Fig. 15 is redrawn, as shown in Fig. 18. The Thevenin's equivalent

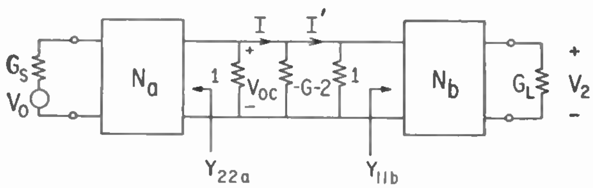

Fig. 18—The cascade-type amplifier.

circuit can be used to determine the current I.

$$I = \frac{V_{oc}}{Z_{eq} + Z_L}, \tag{67}$$

where

$$S_{21a} = \frac{2V_{0c}}{V_0}\,\frac{1}{\sqrt{G_S}} \tag{68}$$

$$Z_{eq} = \frac{1}{1 + Y_{22a}}, \tag{69}$$

$$Z_L = \frac{1}{-1 - G + Y_{11b}}. \tag{70}$$

Next I' can be related to I by

$$I' = \frac{1 + Y_{11b}}{-1 - G + Y_{11b}}\,I. \tag{71}$$

V_2 can then be determined from I':

$$S_{21b} = \frac{2V_2}{I'}\sqrt{G_L}. \tag{72}$$

Combining the above equations, we have

$$S_{21} = \frac{2V_2}{V_0}\sqrt{\frac{G_L}{G_S}} = \frac{S_{21a}S_{21b}(1 + Y_{22a})(1 + Y_{11b})}{2(-G + Y_{11b} + Y_{22a})}. \tag{73}$$

Since

$$\frac{2}{1 + Y_{11b}} = 1 + S_{11b}, \tag{74}$$

then

$$\frac{2}{1 + Y_{22a}} = 1 + S_{22a}. \tag{75}$$

We obtain

$$S_{21} = \frac{2S_{11a}S_{21b}}{2 - G - G(S_{22a} + S_{11b}) - (G + 2)S_{22a}S_{11b}}. \tag{76}$$

A General Steady-State Analysis of Power-Frequency Relations in Time-Varying Reactances*

R. G. SMART†

Summary—A time-varying reactance in an electrical network absorbs power at some frequencies and returns part of this power at other frequencies, the balance being delivered to the source of the reactance changes. This analysis is a study of these power transfers and establishes steady-stage energy relations which apply under very general conditions to varying reactances, including linear, nonlinear, static and electromechanical systems. These relations include the Manley-Rowe results as special cases and provide a useful description of the external characteristics of varying (and fixed) reactances. An "energy coefficient" is derived which specifies the behaviour of an individual frequency transformation. The modes in which an individual transformation can operate are described and the results related to the characteristics of a wide range of varying reactance devices. An example of the application of the theory is given in the Appendix.

This theoretical analysis was undertaken as a basis for digital computer studies of varying reactance systems, since it provides both an over-all description of general characteristics and a means of numerical analysis of particular devices. A fuller account of the work will be available in a later paper by the author [1].

List of Symbols

$e(t)$, $i(t)$, $L(t)$ = general voltage, current and inductance

L_n, q_n, θ_n = the peak amplitude, angular frequency and phase of a frequency component in the inductance

I_v, v, ϕ = the peak amplitude, angular frequency and phase of a frequency component in the current

d_n, d_n^* = the complex amplitude of an inductance component and its conjugate

ω = the angular frequency of a voltage source

k_{yv} = an "energy-coefficient" expressing the energy per radian for a transformation from a current at the angular frequency v to an induced voltage at the angular frequency y

$i(v)$, $i(y)$ = complex amplitude of currents at the angular frequencies v and y

P, P_q or P_ω = a "reactance" power component, *i.e.*, the power delivered to the source of the reactance changes at the indicated frequency or its harmonics

W_v = the electrical power delivered to the inductance at the angular frequency v

u, n, i = real positive integers

m, s, r = real integers

$\mathbf{E}$ = column matrix of voltages across an inductance in frequency components (one frequency for each row)

$\mathbf{I}$, $\mathbf{I}_t^*$ = column matrix of currents through an inductance in frequency components (one frequency for each row) in the same order as the voltage vector, and its transposed conjugate

$\mathbf{Z} = \mathbf{VL}$ = square matrix expressing the impedance of a time-varying reactance, in terms of individual voltage and current frequency components

$\mathbf{Y}$ = the inverse of $\mathbf{Z}$

$\mathbf{V}$ = diagonal matrix of all the angular frequencies v_i in a system arranged in descending algebraic sequence

$\mathbf{L}$ = a skew Hermitian matrix expressing the effective inductance of a constant or varying reactive component

$\mathbf{D}$, $\mathbf{D}^*$ = the diagonal form of the current vector and its conjugate

$\mathbf{W} = \mathbf{VK}$ = a matrix of power components expressing the exchange of power between different frequencies in the electrical system

$\mathbf{K}$ = the skew symmetric matrix of energy coefficients

$\mathbf{C}$ = the capacitance matrix which is the dual of $\mathbf{L}$.

Note 1: It is assumed that there are n frequencies v_i (where $i = 1, 2, \cdots, n$) in the system under discussion. The dimensions of all matrices are then, $(\mathbf{E}$ and $\mathbf{I})_{n \cdot 1}$, $(\mathbf{I}_t^*)_{1 \cdot n}$, $(\mathbf{Z\ V\ L\ C\ Y\ D\ D^*\ W\ K})_{n \cdot n}$.

Note 2: j has its usual meaning $\sqrt{-1}$.

Introduction

Varying parameter networks and devices are very well known for their generation (in the steady state) of new frequencies not present in the original inputs. A very important class of these, with varying reactance, exhibits other special properties, resulting in power amplification, energy conversion (electrical-mechanical) and unwanted instability. Systems containing time-varying reactance have been studied at length [6], but their practical solution, particularly in the nonlinear case, is difficult. These varying reactance phenomena occur in many physical situa-

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tions, but they are vital to modern electrical engineering.

Hartley [2] is credited with the first analysis of an electrical network containing a varying inductance. In this and his subsequent treatment [3] of a capacitive unit, the steady-state power relations were examined in terms of the Fourier components in the system. This approach has been used by many workers since, generally with approximating assumptions about the frequencies present in the system. Manley and Rowe [2] discussed a nonlinear inductance and gave general power-frequency relations based on fairly general assumptions about the shape of the hysteresis loop of the core material. Their results have been applied extensively in describing the power relations in varying reactances. The characteristics of reactive frequency transformations were discussed by Weiss [4] in terms of a quantum mechanical model.

Although other approaches have been used [5], [6], the Fourier component analysis for the steady-state case has been the most widely developed. This paper follows the same line and aims at presenting a consistent analysis of varying reactance devices in general, including linear, nonlinear, static and electromechanical systems. A generalized inductive reactance is used in a steady-state circuit analysis approach. This method has proved successful [10] in digital computer studies of parametric devices and the present analysis has been made to provide a theoretical basis for extending this work.

Steady-State Voltage—Current Relations

Let $L(t)$ be a time-varying inductance, across which the voltage drop, resulting from a current $i(t)$, can be written as

$$e(t) = \frac{d}{dt}\left[i(t)L(t)\right].$$

In the steady state and under very general conditions,

$$L(t) = \sum_{u=0}^{\infty} L_u \cos(q_u t + \theta_u). \tag{1}$$

Consider the voltage drop resulting from a current component

$$i(t) = I_v \cos(vt + \phi).$$

This gives

$$e(t) = -\tfrac{1}{2} I_v \sum_{u=0}^{\infty} L_u\left[(v + q_u)\sin(vt + \phi + q_u t + \theta_u) + (v - q_u)\sin(vt + \phi - q_u t - \theta_u)\right]. \tag{2}$$

Eq. (2) is transformed to a more convenient form by substituting "complex amplitude," in place of peak amplitude and phase angle for the inductance and current components; thus,

$$a_u = \tfrac{1}{2}L_u(\cos\theta_u + j\sin\theta_u) \quad \text{and} \quad i(v) = I_v(\cos\phi + j\sin\phi).$$

The transformed expression becomes

$$e(t) = i(v)\sum_{u=0}^{\infty}\left[j(v + q_u)a_u \cos(v + q_u)t + j(v - q_u)a_u^* \cos(v - q_u)t\right]. \tag{3}$$

Note that these expressions could be given entirely in terms of exponentials, but with less convenience. "Complex amplitude" is used in the following sense: If $I_v \cos(vt+\phi) = i(v)\cos vt$, then $i(v)$ is the complex amplitude of the component and equals $I_v(\cos\phi + j\sin\phi)$. Eq. (3) shows that a current of angular frequency v produces voltages at frequencies $(v \pm q_u)$. Therefore, if an electrical network contains a time-varying inductance with components of angular frequency $q_0, q_1, q_2, \cdots, q_u \cdots$, and a voltage source of angular frequency ω is connected to this network, the resulting voltages and currents will, in general, contain the angular frequencies $(\omega + sq_u)$, where s is any real integer, including zero and negative values.

Consider an inductance as given by (1) in a network containing frequencies v_i, where $i = 1, 2, 3, \cdots, n$.

Note: Theoretically, n is infinite, even if only one inductance component exists. In most practical cases, however, provided the inductance remains non-negative throughout its variation, a finite limit can be assumed [1].

If the current through this inductance is resolved into its n frequency components and (3) applied for each, a summation of the resulting voltages gives the total voltage drop across the inductance, in terms of the same n frequency components. The relation between the voltage and current associated with the inductance can be expressed in matrix form,

$$\boldsymbol{E} = \boldsymbol{ZI}, \tag{4}$$

where $\boldsymbol{E}$ and $\boldsymbol{I}$ are $n \cdot 1$ column vectors, whose elements are the frequency components of the voltage and current, respectively. $\boldsymbol{Z}$ is the $n \cdot n$ impedance matrix for the inductance. Its elements are the coupling impedances between the different frequencies, in accordance with (3) which, for one current component (*i.e.*, one value of v), provides one column of $\boldsymbol{Z}$. The elements of both $\boldsymbol{E}$ and $\boldsymbol{I}$ should be arranged in algebraically-descending frequency sequence. This gives a convenient co-diagonal form to $\boldsymbol{Z}$, in which the major diagonal elements correspond to the constant term in the inductance, while each alternating component in the inductance gives rise to a pair of co-diagonals in $\boldsymbol{Z}$.

To demonstrate the form of (4), consider an inductance in a system to which a voltage source of frequency ω is applied. Let the inductance components be at $q_0 = 0$, $q_1 = q$, $q_2 = 2q$. Taking into account only the frequencies between $(\omega \pm 2q)$ and putting $e(s)$ and $i(s)$ for the complex amplitudes of the voltages and currents at the frequencies $(\omega + sq)$, the matrix equation for the inductance is

$$\begin{bmatrix} e(2) \\ e(1) \\ e(0) \\ e(-1) \\ e(-2) \end{bmatrix} = \begin{bmatrix} j(\omega+2q)L_0 & j(\omega+2q)d_1 & j(\omega+2q)d_2 & \cdot & \cdot \\ j(\omega+q)d_1^* & j(\omega+q)L_0 & j(\omega+q)d_1 & j(\omega+q)d_2 & \cdot \\ j(\omega)d_2^* & j(\omega)d_1^* & j(\omega)L_0 & j(\omega)d_1 & j(\omega)d_2 \\ \cdot & j(\omega-q)d_2^* & j(\omega-q)d_1^* & j(\omega-q)L_0 & j(\omega-q)d_1 \\ \cdot & \cdot & j(\omega-2q)d_2^* & j(\omega-2q)d_1^* & j(\omega-2q)L_0 \end{bmatrix} \times \begin{bmatrix} i(2) \\ i(1) \\ i(0) \\ i(-1) \\ i(-2) \end{bmatrix}. \tag{5}$$

In general, and as demonstrated by this example, the impedance matrix $\boldsymbol{Z}$ can be replaced by the product of two square matrices

$$\boldsymbol{Z} = \boldsymbol{VL}. \tag{6}$$

$\boldsymbol{V}$ is simply a real diagonal matrix of the frequencies v_v, present in the current and voltage (including zero and negative frequencies). It will be found that $\boldsymbol{L}$ is skew Hermitian, (*i.e.*, $L_{rs} = -L_{sr}^*$). This special property of $\boldsymbol{Z}$, which holds for any inductance expressible in terms of (1), (with the further trivial restriction that $\theta_u = 0$ if $q_u = 0$) leads to the important "energy coefficient" concept developed later.

For a time-varying capacitance, there is a dual relation to (6) giving $\boldsymbol{Y} = \boldsymbol{VC}$. The impedance matrix for a capacitance is thus $\boldsymbol{Z} = \boldsymbol{Y}^{-1} = \boldsymbol{V}(\boldsymbol{VCV})^{-1}$. A time-varying capacitance thus has an "inductance" matrix $\boldsymbol{L} = (\boldsymbol{VCV})^{-1}$, which again is skew Hermitian and could be used in (6).

The discussion to date has been in terms of a single reactance. It will be found, however, that any two-terminal reactive network with steady-state variations in inductances and capacitances has an equivalent skew Hermitian inductance matrix (see Section E of the Appendix). Furthermore, in place of (4), there is a corresponding matrix equation for a multiterminal network which again has these special properties (see Appendix, Section F).

Where the reactance variations arise from nonlinearity, these algebraic expressions still hold. However, individual $\boldsymbol{Z}$ matrices (to be combined by addition and inversion, etc.) which are components of a composite $\boldsymbol{Z}$ matrix must be evaluated for the composite operating conditions in nonlinear systems. It should be appreciated that, while the present algebraic analysis is only very slightly complicated by nonlinearity, actual numerical solutions remain difficult.

For any purely reactive network, therefore, there is a matrix equation similar to (4), relating the external voltages and currents. The impedance matrix from this equation can be expressed as the product of two matrices as in (6). The "inductance" matrix $\boldsymbol{L}$ from this equation is skew Hermitian for any network in which the individual reactances can be expressed as the sum of a series of frequency components as in (1). These general conditions include practically any reactance system.

Power-Frequency Relations

From (4), the net average flow of power into a reactance system is given by

$$P = \text{Real part: } \tfrac{1}{2}I_t^*\boldsymbol{E}$$
$$= \text{Real part: } \tfrac{1}{2}I_t^*\boldsymbol{ZI} \tag{7}$$

where I_t^* is the transposed conjugate of the current vector I.

Since (7) applies generally to varying reactance systems, regardless of how these variations are produced, the principle of conservation of energy leads to the following interpretations of P. For a static lossless reactance, P is zero. For a static lossy reactance (*e.g.*, hysteresis), P equals the loss. For an electromechanical device in which reactance changes are associated with physical movement, P equals the electrical power converted into mechanical power (and losses). For an electric motor, P is positive, but for a generator P is negative, since electrical power is produced from mechanical power via the varying reactance.

A varying reactance, even if lossless, can take electrical power at one frequency and deliver this back to the circuit at another frequency. To study this effect and the net transfer of power between the electrical and "reactance" systems, we will now examine (7) in detail.

Eq. (7) gives an expression for the net average input power to the inductance. If the current vector I is replaced by its diagonal form $\boldsymbol{D}$ (the same elements arranged down the major diagonal of an otherwise zero matrix), the summation is avoided and a new matrix of power components results, whose grand sum equals P.

$$\boldsymbol{W} = \text{Real Part: } \tfrac{1}{2}\boldsymbol{D}^*\boldsymbol{ZD}. \tag{8}$$

Using (6) and the symmetry of $\boldsymbol{D}^*$ and $\boldsymbol{V}$, (8) can be written as

$$\boldsymbol{W} = \text{Real part: } \tfrac{1}{2}\boldsymbol{V}(\boldsymbol{D}^*\boldsymbol{LD}) = \boldsymbol{VK}. \tag{9}$$

Since $\boldsymbol{L}$ is skew Hermitian (L_{vv} pure imaginary; $L_{yv} = -L_{vy}^*$), it follows that $\boldsymbol{K}$ is skew symmetric ($k_{vv} = 0$; $k_{yv} = -k_{vy}$). The row sums of $\boldsymbol{W}$ are the electrical input powers at the corresponding frequencies and the grand sum is the net "reactance" power P. For any angular frequency then, the power input from the electrical network at that frequency is

$$W_v = v \sum_y k_{vy}. \tag{10}$$

To show the form of the matrix W, the example given in (5) would yield:

$$W = VK = \begin{bmatrix} \cdot & (\omega+2q)k_{2,1} & (\omega+2q)k_{2,0} & \cdot & \cdot \\ (\omega+q)k_{1,2} & \cdot & (\omega+q)k_{1,0} & (\omega+q)k_{1,-1} & \cdot \\ (\omega)k_{0,2} & (\omega)k_{0,1} & \cdot & (\omega)k_{0,-1} & (\omega)k_{0,-2} \\ \cdot & (\omega-q)k_{-1,1} & (\omega-q)k_{-1,0} & \cdot & (\omega-q)k_{-1,-2} \\ \cdot & \cdot & (\omega-2q)k_{-2,0} & (\omega-2q)k_{-2,-1} & \cdot \end{bmatrix} \tag{11}$$

where, for example, $k_{r,s}$ is an abbreviation of $k_{(\omega+rq),(\omega+sq)}$.

The Energy Coefficient "k"

Consider now the element "k" of matrix K. Let $y=v+q_u$. Then, in terms of the currents $i(v)$ and $i(y)$ at the respective angular frequencies v and y, making use of (3) and (9),

$$k_{vy} = -\text{ Imaginary part: } \tfrac{1}{2}i(v)^*d_u{}^*i(y),$$

while

$$k_{yv} = -\text{ Imaginary part: } \tfrac{1}{2}i(v)d_ui(y)^*.$$

Thus, as noted above,

$$k_{vy} = -k_{yv}, \tag{12}$$

and vk_{vy} is the power component delivered from the network at the angular frequency v, into the v/y frequency coupling impedance. Conversely, yk_{yv} is the power component delivered from the electrical network at the angular frequency y into the same frequency coupling. Since k_{yv} and k_{vy} have opposite signs, one of these will generally be a power output[1] to the electrical network. In any event, if v and y are unequal, there will be a component left over which is the power delivered from the v/y transformation to the source of the reactance changes. This "reactance" power component is

$$(yk_{yv} + vk_{vy}) = (y - v)k_{yv} = q_uk_{yv}. \tag{13}$$

The total behaviour of the varying reactance in the steady state, therefore, is the sum of the effects of the individual frequency couplings, each of which can be thought of as taking power from the electrical network at one frequency, delivering part of it back to the electrical network at another frequency, and delivering the balance to the reactance system. The individual transfers are entirely dependent on frequency, according to the equations:

[1] This depends on the signs of v and y.

$$\left.\begin{array}{l} \text{Power delivered from the electrical system at } y \\ \quad = yk_{yv} \\ \text{Power delivered to the electrical system at } v \\ \quad = vk_{yv} \\ \text{Power delivered to the reactance system at } q \\ \quad = qk_{yv} \end{array}\right\}, \tag{14}$$

where $y=v+q$.

Because of its importance, k_{yv} will be referred to as the "energy coefficient" for the particular transformation. It has the units of energy per radian and is related to the "number of quanta per second N" used by Weiss [4], by $k=Nh/2\pi$ where h is Plank's constant.

Before examining the energy coefficient in more detail, some over-all energy relations for varying reactances in general will be established, based on the above characteristics of the energy coefficients.

General Energy Relations

First, from (9) and (10), if each row of W is divided by its respective angular frequency and the sum taken over all frequencies, the result is the grand sum of K, which must be zero. In terms of the electrical input powers at each frequency (*i.e.*, row sums of W),

$$\sum_v \left(\frac{W_v}{v}\right) = 0. \tag{15}$$

This means that, within the electrical system of frequencies associated with a varying reactance, the net input of "energy per radian" over all frequencies is zero. This relates the powers at different frequencies and, used with the conservation of energy, gives useful information about the over-all performance of varying reactance devices.

The second relation concerns the partitioning of the flow of power to the reactance system. Suppose the electrical system contains the frequency components $v=m\omega+sq$ for any real integral m or s. The net power supplied to the reactance can be partitioned into components:

$$P = \sum_v W_v$$
$$= \sum_v \left(v \sum_y k_{vy} \right)$$
$$= \sum_v \left(m\omega \sum_y k_{vy} \right) + \sum_v \left(sq \sum_y k_{vy} \right). \qquad (16)$$

Considering the part

$$\sum_v \left(sq \sum_y k_{vy} \right)$$

(where $y = m'\omega + s'q$), this will contain pairs of terms of the forms $s'qk_{yv}$ and sqk_{vy}. Making use of (13), these have a sum of $(s'-s)qk_{yv}$. Now, if $s'=s$, this sum is zero; but if not, the component is the contribution to the "reactance" power, at the angular frequency $(s'-s)q$. The sum of all these pairs of terms is thus the total power delivered to the source of the reactance changes at the frequencies sq, for all s. In terms of the power inputs from the electrical system, the "reactance" power at sq is

$$P_q = \sum_v \left(\frac{sq}{v} \right) W_v.$$

By similar treatment,

$$P_\omega = \sum_v \left(\frac{m\omega}{v} \right) W_v. \qquad (17)$$

Note: The example of (11) gives $P_\omega = 0$, $P = P_q$. Provided the harmonic frequency series $m\omega$ and sq have no coinciding terms, the expressions for the "reactance" power at the two frequency series will be independent; in any case, however, $P = P_\omega + P_q$. This result can be extended for more frequency components in the system.

The Manley-Rowe [2] results are special cases of (17). Their general equations for a lossless nonlinear reactance (for which the net reactance power must be zero) are two independent expressions of the form of (17). When examining hysteresis loss, Manley and Rowe limited their study to the case in which all the loss was associated with one frequency. The reactance power P_q in (17) for a nonlinear reactance is equivalent to the total hysteresis loss at all the frequencies in the sq series. Note that the present notation and range of summation differs from that of Manley and Rowe. Their work confirms the present results for the particular case of a nonlinear reactance.

The important relations expressed by (15) and (17) result from the special properties of the impedance matrix $\boldsymbol{Z}$ [as discussed in connection with (6)]. The relations, therefore, hold for any reactive network, provided each individual varying inductance (or capacitance) can be represented by a frequency series, as given by (1) (or its dual).

Furthermore, the relations depend on the summation of the energy coefficients, which appear in pairs, one for each "end" of each frequency transformation in the electrical system. For any given device, there may be several independent systems of frequencies within the electrical system; and, in this event, the relations (15) and (17) hold separately for each frequency system, as well as over-all. This aspect will be elaborated in the Appendix.

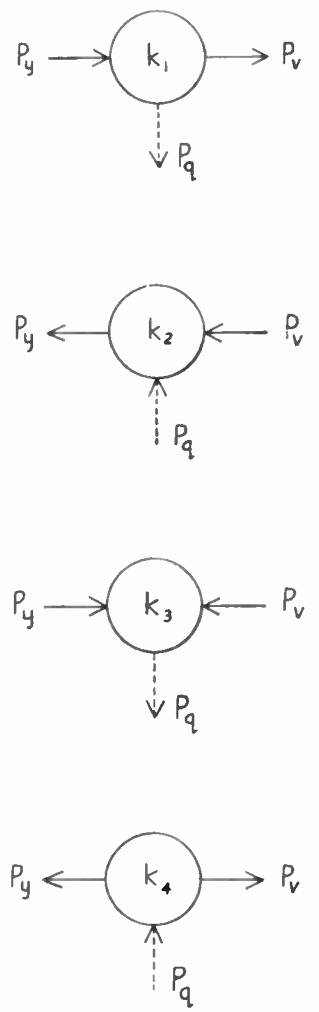
One restriction for which the relations were derived should be noted. Losses in a real reactance (*e.g.*, copper losses) which are associated with the electrical system, were excluded from the analysis. These must be allowed for in the external circuitry. Hysteresis losses or, in fact, any losses which extract power from the electrical system through a frequency transformation will appear as "reactance power." If the copper losses just mentioned were not considered as being outside the reactances, $\boldsymbol{L}$ would not have a purely imaginary major diagonal and the losses would incorrectly appear as part of the reactance power.

Individual Transformation Modes

Eq. (14) shows only one mode of operation for a single frequency transformation. Depending on the relative frequencies and the direction of the power flows, four distinct transformation modes can occur in varying reactance devices. Consider, then, a reactance component at the angular frequency q and currents at the angular frequencies v and y only. If the power input to the transformation at y is P_y, the output at v is P_v and the reactance power output at q is P_q and $y = v + q$; this will be referred to as Mode 1. This and the other modes can be shown diagrammatically as in Fig. 1.

Evidently, all varying reactance power phenomena can be described in terms of a number of individual transformations, each of which is one of these four modes.

For example, the upper sideband of a saturable reactor magnetic amplifier [8] corresponds to Mode 2, where P_q is the input signal power, P_v the carrier input power, and P_y the output power. Mode 1 describes the lower sideband of the same device, if v and y are exchanged. Note that, if the energy coefficient for the lower sideband exceeds that for the upper, the device is potentially unstable, as the net signal input power is negative (neglecting losses). Mode 3 describes the generation of reactance changes in a nonlinear device. If $v = y$, the reactance changes are at twice the current frequency but, if $y = 0$, corresponding to a dc input, the reactance changes can occur at the frequency of the cur-



Mode 1 y and v have equal signs
k positive

Mode 2 y and v have equal signs
k negative

All power flows are reversed.

Mode 3 y and v have opposite signs
k positive

To comply with (14), v is considered to be a negative frequency.

Mode 4 y and v have opposite signs
k negative

Again, (14) requires v to be a negative frequency.

Fig. 1.

rent. These well-established phenomena are thus capable of description in terms of the present work. Mode 4 describes the so-called parametric or negative resistance amplifier [9], with P_q the pump power, P_y the idler power, and P_v the power delivered to the signal circuit.

Some electromechanical devices are: the simple induction motor, Mode 1, with P_y the line input power; P_v the slip frequency power in the rotor; and P_q the mechanical output power. When $v=0$, Modes 1 and 3 describe the normal constant field electric motor and Modes 2 and 4 the corresponding generator. With $q=0$, Modes 1 and 2 describe a normal transformer with $v=y$. Generally, in practical devices, more than one transformation and more than one mode occur simultaneously and the above description is in terms of the principal effect. The examples have been given to demonstrate the application of the above theory.

Conclusions

The general class of devices, consisting of time-varying reactance in an electrical system, has been studied by generalized circuit analysis to establish an expression for the power flow between different frequencies in the electrical system. The over-all operation of such a device was seen to consist of the over-all effect of a number of individual frequency transformations within the electrical system. By introduction of an "energy coefficient," the exchange of power between frequencies in the electrical system and between the electrical and reactance systems were described and four individual modes of operation identified. From the properties of these energy coefficients, over-all power-frequency relations were established, which were independent of the individual energy coefficients. These apply to linear, nonlinear, static and electromechanical systems in general and include the Manley-Rowe results as special cases, thus providing a considerable extension of their work.

The value of the present analysis is that it provides a general theory applying to practically all time-varying reactance systems. It gives over-all characteristics and a simple picture in terms of the energy coefficients of the detailed operation of these systems. The many important practical devices which use time-varying reactance for energy conversion and power amplification appear very simple in terms of the present analysis, which, therefore, will facilitate a more rigorous analysis of existing devices and the development from theoretical considerations of more elaborate devices. The phenomena of instability associated with varying reactances are similarly explained (as negative power flows) from the present analysis, which thus provides a means of studying the incidence and control of instability.

Although this analysis is in terms of an electrical network with time-varying inductance and capacitance, the results can be applied in other fields, *e.g.*, mechanical or aerodynamical, where the same equations apply.

Finally, the analysis provides a consistent approach to varying reactance phenomena which is suitable for the digital computation of results for practical devices. Although the emphasis has been on power relations, the circuit analysis technique used can be applied very simply to finding the voltage, current, and impedance transforming characteristics of these devices. It therefore constitutes a generalization of steady-state circuit analysis to cover varying, as well as constant, reactance components. The Appendix summarizes some of the features of the analysis and its application.

Appendix

Some Applications and Special Features of the Analysis

A. Circuit Characteristics of a Single-Frequency Transformation

If a current of angular frequency v is coupled to a current at the angular frequency y by an inductance component of angular frequency q, where $y=v+q$, putting T as the current amplitude ratio I_y/I_v, the voltage amplitude ratio is $E_y/E_v=(y/v)T^{-1}$ and, if the impedance seen by the transformation at the angular frequency y is $z(y)$, then this is reflected onto the frequency v as $(v/y)T^2z(y)^*$. The power ratios have already been given as proportional to the frequency ratio. The current ratio is the same as a normal transformer action,

$$T^2 = \left(\frac{y^2 d_u d_u^*}{z(y)z(y)^*}\right)$$

where $(d_u d_u^*)$ is the square of the inductance component. These expressions apply even if $v=y$.

B. An Important Special Case

A useful result of (15) and (17) is in the study of a varying reactance device with electrical power input at one angular frequency ω and reactance changes with any waveform periodic at the fundamental angular frequency q. Output power from the reactance to the electrical circuit occurs at the frequencies $(\omega+sq)$ for all real integral s. Thus, the electrical power input W' at ω in terms of the electrical power outputs W_s at all other frequencies is

$$W' = \sum_s \left(\frac{\omega}{\omega + sq}\right) W_s, \tag{18}$$

(*Note:* this sum excludes $s=0$.)

while the power input from the reactance changes at the angular frequencies q, $2q$, $3q$, $\cdots$ is

$$P_q = \sum_s \left(\frac{sq}{\omega + sq}\right) W_s. \tag{19}$$

Thus, a particular power output W_s is supplied jointly by the electrical input W' and the reactance power P_q in a *proportion* fixed by the frequencies, quite independently of the waveform of the reactance. The distribution of power output over the frequencies will, of course, depend greatly on the waveform of the reactance.

C. Negative and Zero Frequencies

Eqs. (15) and (17), as given, depend on the existence of negative frequencies. Although other conventions can be adopted, parts of the present work are consistent only if negative frequencies are allowed. There is no present contradiction, then, in having two separate power components, one at v and another at $-v$. In relating this analysis to the external circuit, we must combine separate components of current or voltage at frequencies of equal magnitude but opposite sign.

Eq. (15) applies to zero frequencies when it is realized that the net power delivered to a reactance (either electrical or reactance power) at zero frequency is zero. The limiting value of the power divided by its angular frequency, as this frequency approaches zero, is the sum of the associated energy coefficients, not infinity.

D. Frequency Topology

The over-all behaviour of a varying reactance device can be described in terms of the power flow between frequencies, once the frequency relations are known. This is conveniently shown by a diagram made up of individual transformations interconnected by electrical or reactance power flows. As an example, the two pump parametric amplifier of Bloom and Chang [7] would be (in its simple form with limited frequencies) as seen in Fig. 2, where k_1 is a Mode 3 transformation with input

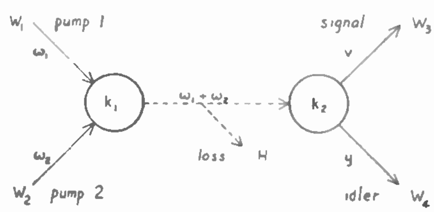


Fig. 2.

powers W_1 and W_2 at the frequencies ω_1 and ω_2, respectively. The sum of these powers is delivered to the reactance changes at the frequency $(\omega_1+\omega_2)$. The second transformation k_2 is a Mode 4. It delivers power to both the signal circuit and the idler circuit at the respective frequencies v and y. The various powers in terms of the energy coefficients are: [*Note:* $y=(\omega_1+\omega_2-v)$]

$$W_1 = \omega_1 k_1 \qquad W_3 = v k_2$$

$$W_2 = \omega_2 k_1 \qquad W_4 = y k_2$$

and the loss $H=(\omega_1+\omega_2)(k_1-k_2)$.

Eq. (17) can be applied to the electrical power inputs and outputs in their form given above, but to apply (15) the standard from is necessary as follows:

$$W_1 = \omega_1 k_1 \qquad W_3 = vk_2'$$

$$W_2 = -\omega_2 k_1' \qquad W_4 = -yk_2,$$

where $k' = -k$ and W_3 and W_4 are negative, being outputs. Eq. (15) is satisfied by the device as a whole, or by either electrical systems (W_1, W_2) or (W_3, W_4). Similarly, (17) can be used over-all to find the loss H, or separately on each electrical system to find the reactance power out of k_1 or into k_2. It will be seen that the frequency topology diagram of power flows fully describes the power relations (noting the characteristics of a single transformation). Eqs. (15) and (17) do not add to this but provide an algebraic approach to the same information viewed externally. The energy coefficients or the frequency-power diagram present the full detail of the internal operation of the device.

*E. An Example of the **Z** Matrix for a Composite Reactance (Fig. 3)*

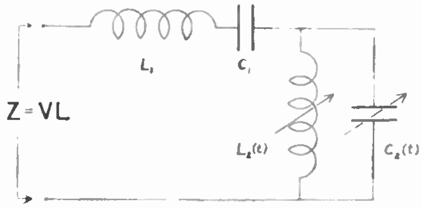


Fig. 3.

Suppose the network contains angular frequencies v_i. These are then the elements of the diagonal matrix $\boldsymbol{V}$. The inductance matrices for each component are, for L_1, $(jL_1)\,\boldsymbol{U}$; for C_1, $[-j(1/C_1)]\boldsymbol{V}^{-2}$; for $L_2(t)$, $\boldsymbol{L}_2$; and for $C_2(t)$, $(\boldsymbol{V}\boldsymbol{C}_2\boldsymbol{V})^{-1}$, where all are skew Hermitian. Assuming that the indicated inverses exist then, the over-all matrix is

$$\boldsymbol{L} = \left\{(jL_1)\boldsymbol{U} - \left(j\frac{1}{C_1}\right)\boldsymbol{V}^{-2} + (\boldsymbol{L}_2^{-1} + \boldsymbol{V}\boldsymbol{C}_2\boldsymbol{V})^{-1}\right\} \tag{20}$$

Note that $\boldsymbol{U}$ is the unit matrix and that the over-all $\boldsymbol{L}$ is also skew Hermitian. Note, from (20), that the constant reactances only contribute to the major diagonal of $\boldsymbol{L}$.

F. Multi-Terminal Networks

In place of (4), for this example, (Fig. 4),

$$\left[\begin{array}{c} \boldsymbol{E}_1 \\ \hline \boldsymbol{E}_2 \end{array}\right] = \left[\begin{array}{c|c} \boldsymbol{V} & \cdot \\ \hline \cdot & \boldsymbol{V} \end{array}\right] \times \left[\begin{array}{c|c} \boldsymbol{L}_1 + \boldsymbol{L}_3 & -\boldsymbol{L}_3 \\ \hline -\boldsymbol{L}_3 & \boldsymbol{L}_2 + \boldsymbol{L}_3 \end{array}\right] \times \left[\begin{array}{c} \boldsymbol{I}_1 \\ \hline \boldsymbol{I}_2 \end{array}\right] \tag{21}$$

where $\boldsymbol{Z}_1 = \boldsymbol{V}\boldsymbol{L}_1$, $\boldsymbol{Z}_2 = \boldsymbol{V}\boldsymbol{L}_2$, $\boldsymbol{Z}_3 = \boldsymbol{V}\boldsymbol{L}_3$.

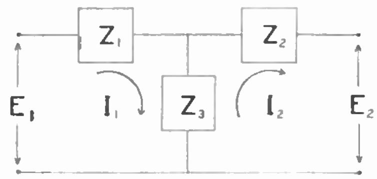


Fig. 4.

Since $\boldsymbol{L}_1$, $\boldsymbol{L}_2$ and $\boldsymbol{L}_3$ are skew Hermitian, so is the compounded inductance matrix. Thus, there is an equivalent form of equations (4), (6) and (9) which lead to the energy relations of equations (15) and (17) for multiterminal networks.

Further information on these generalizations will be given elsewhere by the author [1].

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Error Statistics and Coding for Binary Transmission Over Telephone Circuits*

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Summary—**About 2000 hours of noise data on various digital data telephone links are analyzed with respect to the applicability of coding theory to such links. The analysis indicates that coding for error correction is inefficient for such systems, but that coding for error detection, along with feedback, is remarkably simple and effective.**

Introduction

Commercial transmission of digital data over telephone circuits is capable of providing an overall error probability on the order of 10^{-5}, but in many situations, lower error probabilities are either needed or highly desirable. This paper will enquire whether coding can be efficiently combined with existing digital data systems to provide these lower error rates.

The applicability of coding, however, is very closely related to the particular statistics with which the errors occur. For instance, a simple single error-correcting block code would be quite effective if the errors were independent in time, but almost useless if the errors were to occur clumped together into bursts. Likewise, there are simple burst error-correcting codes[1] that are very effective for errors clumped together into short bursts, but such codes are almost useless if an appreciable number of bursts, or noisy periods, persist for many digits. In fact, even the most advanced error-correction schemes are not very effective if there are noisy periods lasting for hundreds or thousands of digits, since the capabilities of such schemes are overloaded during the noisy periods and wasted at other times. As will be shown later, a simple and effective alternative in such situations is to use coding to detect the presence of errors in a block and then retransmit blocks originally received in error.

Experimental measurements capable of resolving these questions about error statistics have been made for some time. One technique [2], [5] developed at Lincoln Laboratory, is to transmit a 16-symbol message, compare the received message with that transmitted and punch onto paper tape pertinent error data if the transmitted and received messages do not agree. The paper tape is then processed to determine several quantities such as the average error rate, the variation in error rate as a function of the day of the week or the time of the day, and the distribution of errors within the 16-symbol words.

This type of analysis indicated that the errors on digital data telephone circuits are not independent, but it left unanswered the question about the distribution of the duration of noisy periods. About 2000 hours of this data have been re-analyzed in this report in order to answer this question. The analysis of these data will be given in the next section and then the implications to coding will be discussed.

Experimental Noise Data

The noise data [3], [6], [7], [9] that were used in this report were obtained using four different data systems: CTDS, Milgo, A-1 and Kineplex. The types of telephone line and amount of time over which the data were taken are tabulated in Table I. The lines used were private lines leased from the New England Telephone and Telegraph Company. The lines received no special treatment or equipment and are expected to be typical of the private wire plant.

TABLE I

Modulation system	Channel data rate (symbols/sec)	Type of telephone line	Time (hours)
CTDS	1300	Microwave	305
		K-carrier	241
A-1	1300	Microwave	521
Milgo	1000	Microwave	204
		K-carrier	249
Kineplex	2400*	Microwave	449
		K-carrier	118
		Total	2087

* The error data were recorded for only one channel of the Kineplex system. Data were transmitted but not recorded on remaining seven channels [6].

The recording of the error data was accomplished with the ADDER [2], [5] (automatic digital data error recorder). The pattern of ones and zeroes in the transmitted word was determined by toggle switch settings and was usually kept constant during a day's run but was varied from one day to the next. When errors occurred, the ADDER punched the received word and time of occurrence onto paper tape. Since the time required to punch the necessary data for a received word containing errors is approximately thirty times longer than the times required for transmission (assuming the 1300-bit/sec rate), some provision is necessary to handle successive words with errors. The ADDER has storage for three 16 symbol words. If words with errors

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‡ Lincoln Lab., M.I.T., Cambridge, Mass.
[1] See [10] for discussion and bibliography on burst codes.

are received when these three storage registers are in use, a counter, called the "excess error" counter, is used to count the number of such words even though the words cannot be saved. When a storage register is again available for error data, the contents of the "excess error" counter are punched onto tape and the counter is reset to zero.

Successive 16 symbol words were combined into longer sequences of 127, 255 and 511 symbols which will be called code words. Several error statistics of the longer sequences were then computed. These lengths were chosen to fit the length requirements for the Bose-Chaudhuri [1] class of cyclic codes. The lengths of the code words differ by one from a multiple of 16 and in forming the code words, this extra symbol was ignored. If an "excess error" count occurred in forming the longer sequences, the last recorded 16 symbol error pattern was repeated immediately in the longer words as many times as indicated by the counter.

The following error statistics were measured for the code words:

1) The distribution of the number of errors in code words that have errors,
2) The distribution of burst lengths[2] in code words that have errors,
3) The distribution of the number of 16 symbol words with errors in code words that have errors,
4) The distribution of the number of code words from one code word with errors to the next code word with errors.

The results of these four measurements for the CTDS data system using code words with lengths 127, 255 and 511 are shown in Figs. 1 through 4. Continuous curves are drawn through the data points to indicate the general trend of the data even though more accurate representations would have been the usual stair-step curves for discrete distributions. In Figs. 5 and 6 the distribution of errors and burst lengths for code words with 511 symbols is shown for the four data transmission systems that were used.

Discussion of Noise Data

The figures clearly indicate that the errors occur in bursts rather than independently. The average probability of a symbol error in the telephone line channel is about 10^{-5} [3]. If these errors were independent, then, for word lengths of 511 symbols, words with single errors would account for 0.997 of all the words with errors instead of 0.21 as in Fig. 1. It should be noted that the error data shown in Figs. 1 through 4 are nearly independent of the length of the code. Thus if error correction is used to correct errors in 90 per cent of the code words with errors, the code must be capable of correcting up to 30 randomly placed errors independent of whether a 127 or 511 symbol code is used. Although the individual curves differ considerably when the data system is changed as in Figs. 5 and 6, they still illustrate the same characteristics of the noise. It is also to be observed from the numbers of 16 symbol words with errors in code words with errors in Fig. 3 that there are many instances in which errors are spread throughout the code word rather than being confined to one or two short severe bursts.

One additional characteristic of the data should be discussed. There was evidence in some of the data of complete loss of signal as though the line were momentarily open or as though the signal suffered from severe fading. Although the actual time such conditions prevailed was very small, the curves in Figs. 1 through 6 would be noticeably affected by including these data since large numbers of errors are involved. For example, the signal was lost during 0.65 minute of one of the Milgo records. The data collected during this time would result in 76 code words with errors which is more than 10 per cent of the errors found in the remaining 448 hours of noise measurement. Consequently noise data obtained when a complete loss of signal was apparent were removed from the data used in Figs. 1 through 6, and this amounted to 0.65 minute for the Milgo system and 4.30 minutes for Kineplex. In addition, of the 23 A-1 records which were initially selected, five were not used because they each contained several sustained periods of signal dropout.

Complete line failures of this type usually result in a received word with all zeroes or all ones depending upon the modulation system which is used. These failures are easily detected by the decoder by requiring that all transmitted words contain a minimum number of ones (or zeroes). For example, if an open line results in the all zero sequence at the receiver, complementing the parity check symbols after they are computed would insure that the all zero sequence is not a code word for now all zeroes in the information places will result in all ones in the parity places.

Other interesting characteristics of the raw data [6], [7] are the lack of symmetry in the channel with respect to the input symbols, the dependence of errors upon the previously transmitted symbols, and the dependence of errors on the time of day. These will not be discussed further here since they are not very pertinent to the coding and error-detection schemes discussed.

Error Correction

In addition to characterizing the noise, the curves are also useful in forming conclusions about the use of error-correcting codes for telephone circuits. Several error-correcting codes and decoding procedures are available for binary channels and an important question is whether or not these codes are practical in combating the noise found in telephone circuits.

[2] The "burst length" is defined here as the number of symbols from the first to the last symbols in error in any given code word. Thus the minimum burst length for a code word in error is one and the maximum is the length of the code word.

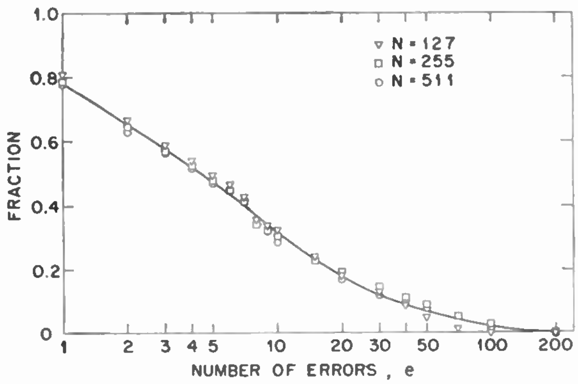

Fig. 1—Fraction of the N symbol code words with errors which have more than e errors (CTDS data).

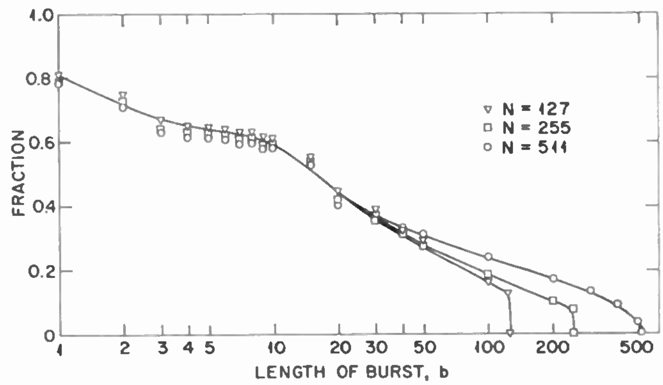

Fig. 2—Fraction of the N symbol code words with errors for which the burst length of errors is greater than b (CTDS data).

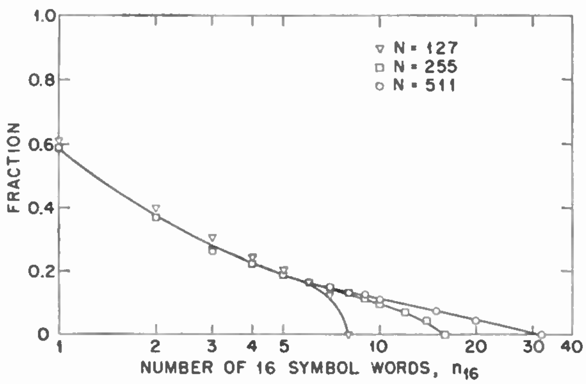

Fig. 3—Fraction of the N symbol code words with errors for which the number of 16 symbol words with errors is greater than n_{16} (CTDS data).

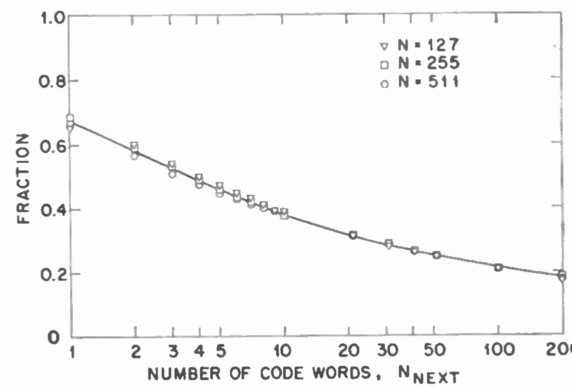

Fig. 4—Fraction of the time for which the number of code words from one code word with errors to the next code word with errors is greater than N_{next} (CTDS data).

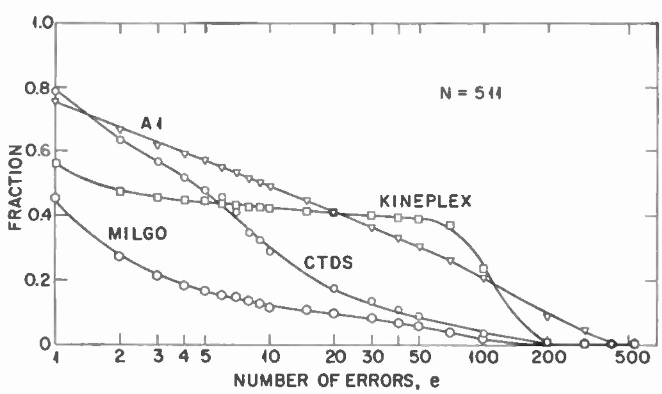

Fig. 5—Fraction of the code words with errors which have more than e errors.

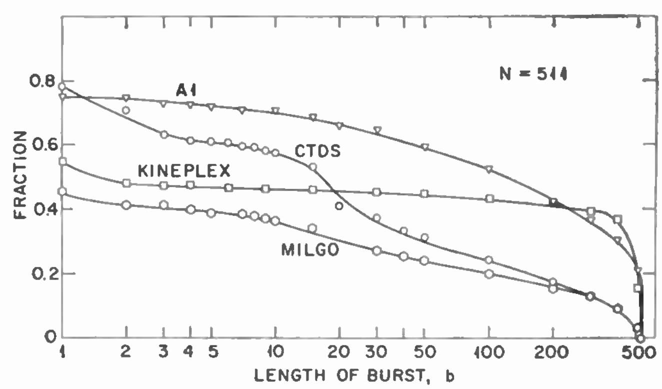

Fig. 6—Fraction of the code words with errors for which the burst length of errors is greater than b.

For a specific example, consider the 546 hours of data for the CTDS system which represents a total of 5×10^6 code words of length 511. Of these only 1475 (less than 1 out of every 3000) contained errors. Assume that an error-correcting code is used to reduce by a factor of 10 the number of code words containing errors. This would require that the code correct 90 per cent of the words that have errors, or from Fig. 1, that the code correct all code words containing up to 30 errors. Since the uncorrected words contain so many errors, even such a code would reduce the number of symbol errors by a factor of less than 2. The amount of equipment to correct 30 errors with present-day decoding techniques would prohibit the use of such a code with telephone circuits except perhaps for some very unusual applications. Even if the error correction were done, a considerable amount of redundancy would be required in the transmitted code word. Although an optimum code capable of correcting up to 30 errors in a block of 511 symbols is not known, the theory of error-correcting codes guarantees that at least 161 of the 511 symbols would have to be parity checks. If a burst error-correcting code were used under the same assumptions, it would have to be capable of correcting a burst of up to 300 symbols which is also unrealistic.

Any conclusion about the feasibility of error-correcting codes with telephone circuits must be prefaced by repeating again that the curves represent 2000 hours of data taken over several private telephone lines with four types of modulation equipment. Improvement in either the noise characteristics of the telephone lines or in the modulation equipment may affect the conclusion. However, subject to the foregoing reservations, error-correcting codes in general do not appear practical for use with telephone circuits because their cost is too great in relation to the number of code words with errors that would be corrected.

Error Detection

Since error-correcting codes appear impractical for telephone circuits, the question of whether redundancy of any type is useful and practical and whether there are other schemes for correcting errors arises and will be discussed.

It is easy to envision communication systems in which the receiver is willing to discard data containing errors but is unwilling to tolerate errors in the data that is accepted. It will be shown that for applications such as these, the addition of very small amounts of redundancy can reduce the probability of an undetected error almost to zero. In other communication systems there may be equally stringent conditions on accepting only error-free data, but it may be possible to signal the transmitter to retransmit the message if errors are detected. In situations such as these, error-detecting codes can be very useful. In general, error-detecting codes can be implemented very easily with a minimum of equipment and, for the same probability of undetected error, require much less redundancy than error-correcting codes. The noise in telephone circuits lends itself very well to error detection. The channel seems to operate in two extreme states: noise-free and very noisy, and about all that can be done from a practical viewpoint if reliable communication is required is to use the channel when it is good and not use it when it is bad. The experimental noise data were used with several error-detecting codes which were simulated on an IBM 709 computer. These results will be presented after a brief description of the codes and their implementation.

One promising class of new codes for error detection is the class of cyclic codes. These are systematic group codes in which a sequence of k information symbols are encoded into a sequence of n symbols by appending $n-k$ redundancy symbols. Cyclic codes can be encoded with an $n-k$ stage shift register and if only error detection is required, the decoding also can be done with an $n-k$ stage shift register.

The encoding and decoding equipment are shown in Fig. 7. A square box represents a single stage shift register and a circled plus sign represents addition modulo two. Encoding is done as follows: [10]

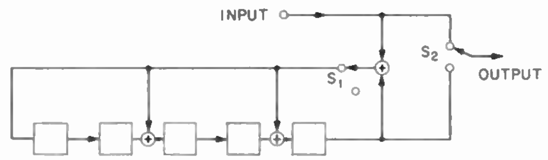

Fig. 7—Typical encoder and decoder for a cyclic shift register code.

1) With the shift register initially set to all zeroes and switches S_2 and S_2 in the positions shown, the k information symbols are fed into the shift register one at a time and simultaneously transmitted over the channel.
2) After the last information symbol has entered the shift register, the shift register contains the $n-k$ redundancy symbols. Switches S_1 and S_2 are now placed in the opposite positions and the contents of the shift register are transmitted over the channel to form the code word with n symbols.

The decoding for error detection is performed in much the same way.

1) With the register initially set to all zeroes and switch S_1 closed, the entire n-symbol received word is shifted into the register one symbol at a time.
2) After the last symbol has entered, a test is made of the contents of the register and an error is assumed to have occurred unless every stage contains a zero.

TABLE II

Code size			Error data				
Length (symbols)	Parity checks (symbols)	Minimum distance (symbols)	Data system	No. of code words	No. of word errors	Average time for undetected errors (years)	No. of undetected errors actually found in data
511	18	5	CTDS Milgo Kineplex A-1	5.0×10^6 2.8×10^6 1.4×10^6 4.8×10^6	1475 693 1238 2979	21 97 35 9	0 0 0 0
255	16	5	CTDS Milgo Kineplex A-1	1.0×10^7 5.7×10^6 2.4×10^6 9.5×10^6	1758 806 1685 4284	4 20 5 1	0 0 0 0
127	14	5	CTDS Milgo Kineplex A-1	2.0×10^7 1.1×10^7 4.8×10^6 1.9×10^7	2118 908 2659 6405	.9 4 .5 .2	0 0 0 1
127	7	3	CTDS	2.0×10^7	2118	.006	10

Bose-Chaudhuri codes were selected for use with the experimental noise data. These are cyclic codes with the following properties. For any positive integers m and t there is a Bose-Chaudhuri code with length $n=2^m-1$ for which the minimum distance[3] between any pair of code words is at least $2t+1$. The number of parity check symbols is never greater than mt. The combinations of length, minimum distance and number of parity checks which were used are tabulated in Table II.

The Bose-Chaudhuri codes with minimum distances three and five are close packed [4] and hence are optimum when used for error correction with a symmetric, independent channel. The structure that these codes must have to be optimum when used for error correction should make them good codes for error detection, and in fact, for a given code size, no greater minimum distance can be achieved with any fewer parity checks.

Any code with minimum distance $2t+1$ will detect any $2t$ or fewer errors that occur. In addition, when a cyclic code with r parity checks is used for error detection, all error bursts of length r or less are detected. The only way for an undetected error to occur is for noise to alter the transmitted word in such a way that the received word is identical to some other code word. This is possible only if more than $2t$ errors or a burst of more than r symbols occur, and even then, the probability that the errors will be undetected can be extremely small.

An exact determination of the probability of an undetected error is difficult for it involves a determination of the probability of an undetected error for the individual code words. A reasonable simplification is to assume that when over $2t$ errors occur or when an error burst of more than r symbols occurs, the code words and noise are essentially independent. Since there are 2^k code words and 2^n possible received words, the probability of an undetected error in such cases is approximately 2^{k-n}. With these assumptions, the probability of an undetected error is now equal to the product of the probability that the received word has more than $2t$ errors or an error burst of more than r symbols by the probability that such a received word is a code word.

[3] The distance between two binary sequences is equal to the number of positions in which the two sequences differ.

To illustrate the above, consider the CTDS data which represents 546 hours of ADDER operation. When 511 symbol code words were formed from the 16 symbol code words, a total of 1475 code words contained errors. Of these, 714 contained four or fewer errors. A Bose-Chaudhuri code with $n=511$ and $n-k=18$ has a minimum distance of five so that all of the 714 errors were detected. If the probability of an undetected error for error patterns with more than four ones is assumed to be 2^{-18} as mentioned above, one would expect that since 761 code words had more than four errors in 546 hours of operation, an undetected error would occur on the average of once every $2^{18}\times546/761=1.87\times10^5$ hours of operation or once every 21.3 years. Similar results were obtained using other data systems, and these results are tabulated in Table II. When a $n=127$, $n-k=7$ code is used with the CTDS data, an undetected error is expected every 50 hours of operation. Actually ten undetected errors occurred, or one every 55 hours of operation, which is in good agreement with the expected rate.

Feedback

In many communication systems it is unacceptable for the receiver to know merely that there has been an error in a block of messages; the receiver must also be able to find the correct message. If a noiseless return channel, or feedback channel, is available from the receiver to the transmitter, the receiver can simply request a retransmission of the offending message from the transmitter. Complications arise, however, if the feedback channel is noisy. Effective methods to handle a noisy feedback channel have been proposed by Metzner and Morgan [8] and by Wozencraft and Horstein [11].

The fundamental characteristic of all these methods is that the transmitter retransmits the message both when the feedback signal is a request for retransmission and when the feedback signal is in doubt. Thus a particular block of messages will fail to be correctly received only when there is an undetected error in either the forward or feedback directions. However, as described before, the probability of an undetected error using coding can be made as small as desired in both channels.

A much more likely event than an undetected error is that the same message will be correctly received several times due to detected errors in the feedback link. Metzner and Morgan [8] suggest handling this problem by means of an alternating tag on successive messages and there are also other procedures that can be used.

The use of feedback in communication generally causes some loss in rate over and above that lost by coding. In the case of telephone circuits, however, the rate loss due to repetitions is quite small. Assuming that both forward and feedback channels have similar characteristics as the data in this report, it should be necessary to repeat less than 1 block in 1500.

If such a feedback scheme is to be used with a constant data rate source, considerable storage is needed to handle the waiting line created by the occasional retransmission of messages. However, if one considers a telephone circuit to be a channel with time-varying capacity, and notes from the data in Fig. 1 that the capacity occasionally drops to practically nothing for periods at least as long as 500 digits and probably longer, then one sees that considerable storage is required by any scheme that transmits reliably over such a channel.

To keep the waiting line from building up with a constant data rate source, it is necessary to have the source rate somewhat lower than the transmitter rate. The smaller the rate sacrifice here, the larger the storage that has to be provided. The storage can be of the fixed access type, but still it will constitute the major cost of the coding system. It is possible, although it appears unlikely, that a small amount of error correction in this situation could reduce the storage requirement enough to be economically feasible. In the situation where the source rate is controllable, however, the argument is very clearly in favor of simple error detection and feedback, since little storage is necessary.

To illustrate the effect of attempting some error correction along with detection for telephone data systems, consider the CTDS data. Out of the 1475 code words of length 511 recorded with errors, only 323 had single errors. In Table II it is estimated that a Bose-Chaudhuri code of length 511 with 18 parity checks will produce an undetected error on such data about once every 21 years. If this code is used for a single error correction plus detection, an undetected error will occur not only when the noise changes one code word to another code word, but also when it changes a code word into any of the 511 words at distance one from a code word. This would produce undetected errors at an estimated rate of once every 15 days. An extra 9 parity checks would be needed to decrease the undetected error rate to the previous value of once every 21 years. Thus the reduction in rate necessary to correct the 323 single errors is equivalent to a loss of 91,000 of the original 5 million code words. This loss of rate will aggravate the waiting line problem and although no data are available here, it does not seem likely that many of the 323 words with single errors occur during the very noisy periods when the waiting line gets long.

Mathematical Models for Telephone Circuits

In the preceding pages, a large body of experimental error statistics has been used to determine methods for decreasing the error probability on digital data links. A seemingly more elegant approach to this problem is that of using the data to construct a mathematical model of error statistics and then determining how to decrease the error probability in the model. This approach has been used in the past, often with quite misleading results, so it is worth pointing out the dangers and difficulties here.

In constructing a model, it is first necessary to separate the important characteristics of the data from the trivia. If this is not done with caution, however, these "trivial" occurrences will be precisely the occurrences that cause errors in any reasonably chosen coding scheme. For instance, one could measure the first and second order error statistics on a digital data link and decide to ignore both the higher order statistics and the second order statistics for lengths greater than 15 or 20 bits. A model constructed from this data would clearly illustrate the short bursts of the noise but would completely ignore the very long periods of low channel capacity that make error correction impractical.

A somewhat better model could involve a channel with two states—one noisy and one noise-free, along with a distribution function determining the time spent in the noisy state. Assuming an error rate in the order of $\frac{1}{2}$ in the noisy state, one is forced to conclude after the reception of a small number of correct digits that the channel is back in the noise-free state. This model again illustrates the short severe bursts of noise, but fails to bring out the more moderate bursts of much longer duration. Both these models would indicate the effectiveness of burst error-correcting codes, whereas the actual data clearly indicates just the opposite.

The previous paragraphs do not imply that these models are completely inapplicable to telephone lines. Such simple models might be useful in learning how to improve both telephone lines and their digital data transmitters and receivers. Such models are certainly inapplicable to coding questions, however, due to the great sensitivity of coding to slowly varying channel characteristics.

One can easily visualize models that are broad enough to include most of the quantities of interest in coding studies. For instance, the channel could be considered as having two states again, but the noisy state could now be characterized by a distribution function over both the error rate and the duration of the noisy state. The attractiveness of using a model rather than direct data fades somewhat for a model this complex that still omits many important characteristics. The results of this paper would have been both difficult to derive and somewhat questionable if a model approach had been used.

Conclusions

The data demonstrate that error-correcting codes are impractical with the circuits and modulation systems that were tested. The fact that all of the systems and circuits behave in a comparable manner and that over 2000 hours of operation is represented suggests that this conclusion might be extrapolated to most of the "private wire" plant.

On the positive side, it has been shown that error-detection codes either with or without feedback are practical. Using moderate size block lengths (100 to 500 symbols) and a very small amount of redundancy (15 to 20 symbols), it is possible to transmit over "private wire" telephone circuits with a probability of an undetected word error so small it almost can be neglected. Moreover, the encoding and decoding for such codes can be easily implemented with a modest amount of equipment most of which consists of 15- to 20-stage shift registers.

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Correction

I. L. Auerbach, author of "European Electronic Data Processing," which appeared on pages 330–348 of the January, 1961, issue of these Proceedings, has brought the following to the attention of the Editor.

On page 345 the trade name C.d.C. was incorrectly ascribed to be a licensee of Potter for magnetic tape units used by SEA. This error is regretted. Compagnie des Compteurs has informed the author that their tape transports are a product of their own design and development.

Communications Satellites Using Arrays*

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Summary—**Passive reflectors require large ground antennas and powers but permit use of unstabilized reflectors. Active systems, on the other hand, offer attractive performance through high directivity antennas but generally require stabilized vehicles. Further, the reliability of a satellite repeater amplifier is a critical factor.**

Automatic angle return arrays are investigated for both passive and active systems. These arrays, called Van Atta arrays, return a signal in the direction of incidence and are effective over at least ±45°. Thus only partial stabilization or single axis (spin) stabilization need be used, greatly simplifying the station-keeping orbit control problem.

An active Van Atta (AVA) scheme has inherently high reliability since many of the distributed amplifiers can fail without serious performance degradation. The distributed structure also allows use of low-power solid-state amplifiers.

This paper points up the salient advantages of the AVA system, and de'ineates the major areas where further work is indicated.

I. Introduction

MUCH has been written about communications satellites, both active and passive.[1–9] Generally considered for the active case is a single antenna, whereas metallic spheres and trihedral corner reflectors are considered for the passive case. Ryerson,[6] however, also considered the Van Atta reflector, a passive array wherein each element is connected to one other element. Obviously the next step is to insert amplifiers in each connection, thereby making an active Van Atta array (AVA). This paper discusses the uses of such arrays for both active and passive communications satellites.

In the next section, the basic array will be described in several geometries, followed by a discussion of effective aperture and element factor. Section IV covers passive arrays. Active systems are treated in Section V. A comparison of passive and active systems is made in Section VI.

* Received by the IRE, October 13, 1960; revised manuscript received February 27, 1961.

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II. The Van Atta Array

The Van Atta passive array[10,11] is the discrete equivalent of a corner reflector; it has a directivity commensurate with the physical aperture, yet returns energy at the incidence angle. Fig. 1 depicts the basic principle wherein conjugate pairs of elements are connected by equal (electrical) lengths of cable (or waveguide, etc.). In this figure, all incident rays have equal phase at the wave front, and each ray acquires an additional phase shift of $\alpha+3\beta$ in getting back to the same wave front; thus exit rays are in phase at the same angle. An odd number of elements can be accommodated by connecting the center element back on itself with the same cable length, then connecting conjugate element pairs as before.

The array can be mechanized in a number of geometries, among which are planar, cylindrical, and spherical. Fig. 2 shows the connections required for a planar array wherein each box represents an element, and boxes containing the same number are connected together, with all cable lengths equal. It can be seen that each quadrant of a rectangular array is a mirror image of the opposite quadrant. In this case, the principal plane beamwidths are just what one would expect from a conventional uniformly excited array. The scattering cross section of the array is found to be essentially that of a flat plate of area equal to the effective aperture (see below). Fig. 3 shows a ring source on a cylinder, and it can be seen that if the line lengths are adjusted to include the departure of the array surface from the projected planar aperture, reinforcement again occurs at the incidence angle. For a two-dimensional array on a cylinder, there would, of course, be a mirror-image connection in both azimuthal and axial directions. Several arrays would be needed to cover the cylindrical surface. Other useful geometries include the tetrahedron with an array on each face, the cube, and the sphere. Several arrays would be used on the spherical surface. Before discussing applications, the effective aperture variation will be investigated.

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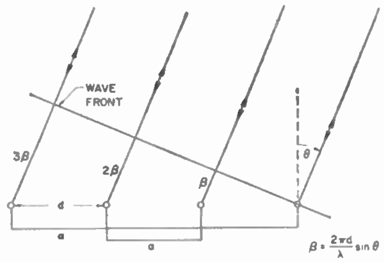

Fig. 1—Linear Van Atta reflector.

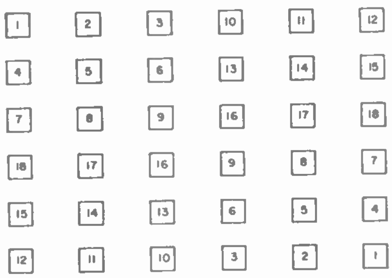

Fig. 2—Square Van Atta reflector.

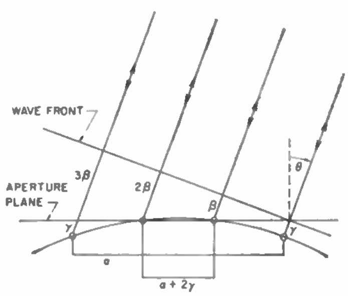

Fig. 3—Cylindrical array.

III. Effective Aperture and Element Factor

A. Effective Aperture

The effective aperture of an electronically scanned array depends upon the element factor, the array geometry and scan or incidence angle. The element factor will be discussed later in terms of specific realizations of these arrays. The effective aperture can be defined on the basis of beamwidths[12] or on the basis of directivity (maximum gain). Since the latter corresponds to the most common usage, this will be used here. Directivity is given by

$$G = 4\pi A/\lambda^2. \tag{1}$$

Here A is the effective aperture, which depends upon the manner in which the array is excited. The Van Atta array functions as a linear phase device and is probably not amenable to supergain. Further, since high sidelobe ratios do not appear necessary for communications satellite purposes, this paper will be concerned exclusively with the so-called uniform array in which all elements are excited with constant amplitude and with linear phase. For a continuous radiator, the effective aperture at broadside (A_0) is just the physical aperture, and the effective aperture varies as $\cos\theta\cos\phi$ out to a transition point beyond which the array acts in an endfire fashion. A discrete broadside array of half-wave spacing has a somewhat greater effective aperture than a continuous aperture of the same size. In the limit of end fire, the effective aperture normalized wrt A_0 is given by[13]

$$\frac{A}{A_0} = \frac{3}{4\sqrt{2}}(\lambda/a)^{1/2}. \tag{2}$$

Beamwidth is given by the standard formula

$$\theta_0 = \frac{102°}{N} \simeq \frac{51°\lambda}{L}, \tag{3}$$

where N is the number of elements on a side. Fig. 4 gives an estimate of effective aperture variation.

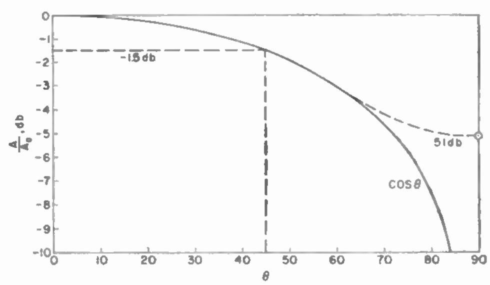

Fig. 4—Effective aperture.

B. Element Factor

The response of arrays is comprised with good accuracy of the product of an array factor described above and an element factor. The elements used would typically be either half-wave dipoles or slots, and it appears at first glance that a printed slot or dipole array on the surface of the satellite is most attractive. For a half-wave dipole, the element factor is

$$\frac{\cos(\pi/2\sin\theta)}{\cos\theta}. \tag{4}$$

This is plotted in Fig. 5 along with the cos θ factor for an elemental dipole. At 45° off normal, the element factor is down 4 db. Typical resonant slot patterns are more complex than (4), depending on the method of feeding, but are all numerically very close to the half-wave dipole pattern of Fig. 5.

The element spacing in the array would be nominally $\lambda/2$. As the spacing increases up to a value of about 0.9, the gain of a broadside array remains proportional to

[12] R. W. Bickmore, "A note on the effective aperture of electrically scanned arrays," IRE Trans. on Antennas and Propagation, vol. AP-6, pp. 194–196; April, 1958.

[13] M. J. King, and R. K. Thomas, "Gain of large scanned arrays," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 635–636; November, 1960.

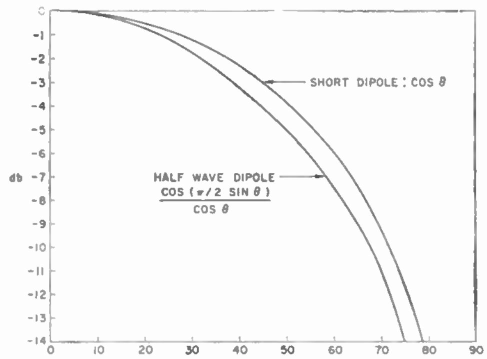

Fig. 5—Element factor.

the area.[14] Beyond this point, the gain no longer increases with area as before. However, spacings of greater than $\lambda/2$ produce secondary main beams at scan angles off broadside and hence represent a gain loss of at least 3 db and possibly more; thus the $\lambda/2$ figure is often used for scanning arrays.

IV. Passive Array

The principal advantages of a passive satellite are that no power supply or equipment reliability problems are incurred, that the satellite may have a very wide bandwidth, and that the attitude may not need to be controlled. The two common types of passive reflectors are the metal sphere and the triplane assembly of eight trihedral corner reflectors. The disadvantage of the spherical reflector is its low-scattering cross section compared to its diameter. The corner reflector has an excellent scattering cross section; however, it suffers some diffraction from the edge at angles close to 45°, and the diffracted component may produce dips[15] as large as 20 db. Compare this with a large planar array whose response is down only 5.5 db at 45°.

The wide angle response of the Van Atta reflector allows the satellite array to return the signal with high gain over a wide range of angles, at least ±45° from broadside. This means that a satellite need only have limited attitude stabilization, in that as long as the earth is within ±45° of the array axis, the response will be within 5.5 db. Essentially complete response could be obtained by disposing six arrays around a cube or four arrays around a tetrahedron,[16] or several arrays around a cylinder or sphere.

For purposes of calculation, a stationary equatorial satellite has been assumed; the numbers can readily be changed to accommodate other orbits. From the "stationary" altitude, the earth subtends an angle of 17.5°; for a vehicle with attitude stabilized to ±1.5°, an antenna of about 20° beamwidth is needed which realizes a gain of about 17 db. For two fixed terminal points, a narrower beam could be used. For example, with a satellite at a longitude of 320°, the angle at the satellite[17] between New York and Lisbon is 7°44′. Fig. 6 depicts such a case; the reflector beamwidth needs to be twice the angle, or 15°28′≃16°. For angles of incidence other than broadside, the scattering cross section is somewhat reduced, but there is a compensating change in that there is less signal loss 8° away from the beam axis due to the broadened beam. For an orientation of the array axis in the direction of wave incidence, the bistatic return is 3 db below the monostatic return, at an angle equal to half the beamwidth. With an orientation of array axis at 45° to the incident wave, the bistatic return is 1.5 db below the monostatic return at the same angle as above (calculated using sin X/X pattern), but the effective aperture-element factor reduces both by 5.5 db making a total of 7.0 db. For the example above where the beamwidth is 16°, an array with seven elements on a side (49 elements in a square configuration) would be needed. At broadside the monostatic equivalent gain is 21.9 db. Bistatic gain 8° away is 18.9 db, and at a 45° inclination angle this decreases to $G=14.9$ db.

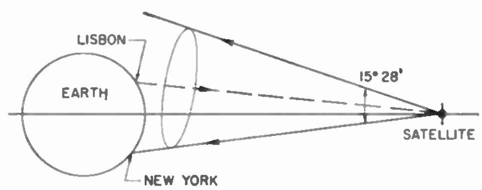

Fig. 6—Passive array.

Directivity and effective area were related in (1); scattering cross section is related to these by

$$\sigma = \lambda^2 G^2/4\pi, \tag{5}$$

where G here is the equivalent directivity as used in the two-way active radar range equation.

Appendix I contains passive range calculations for the New York-Lisbon example. The results are given in Fig. 7 for both a Van Atta reflector and a 100-foot Echo balloon. Parameters assumed are for a digital biphase modulation system. The figure clearly demonstrates a well-known fact: high gain reflectors are not large in wavelengths (due to coverage limitations) and thus do not intercept much of the primary radiated energy. Isotropic scatterers, on the other hand, are inefficient, but can be very large.

[14] H. E. King, "Directivity of a broadside array of isotropic radiators," IRE Trans. on Antennas and Propagation, vol. AP-7, pp. 197–198; April, 1959.

[15] S. D. Robertson, "Targets for microwave radar navigation," *Bell Sys. Tech. J.*, vol. 26, pp. 852–869; October, 1947.

[16] One problem here is the interference produced when two or more arrays are illuminated by and return energy to the earth.

[17] Calculated by Dr. Rojansky of Space Tech. Labs. Inc., Los Angeles, Calif.

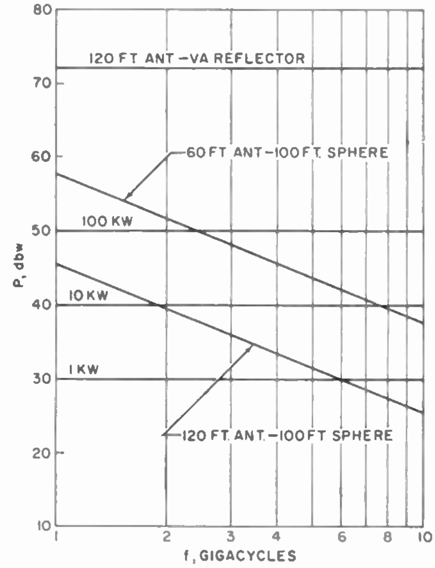

Fig. 7—Ground power per kc bandwidth.

V. Active Van Atta Arrays

A. Active Systems

There are at least three ways in which an active array can be used. First, a broadside array as a single port input (or output) device can simply replace a parabolic dish-type antenna. This usage will not be discussed. Second, an electronically steered array could be used on a vehicle with incomplete attitude stabilization. Here a sensing device would be used to determine the appropriate direction of radiation,[3] and the requisite phase shifts would be produced by ferrite phase shifters, etc. The complexity of the sensory/phase shift mechanism probably precludes serious consideration of this scheme for some time to come. Third, the Van Atta array of the previous section can be made active; this array also applies to partially-stabilized and spin-stabilized vehicles. The AVA (active Van Atta array) appears the best of the three, and will be the subject of this section.

AVA needs an amplifier in each cable or transmission line connecting two elements. Either bilateral gain is needed or two back-to-back unilateral amplifiers could be used. Circulators to separate signals in the two directions could also be used. An advantage of a distributed array amplifier or AVA is that enhanced reliability results, as will be shown later.

B. Single Frequency Schemes

The amplifier for each pair of AVA elements must either be bilateral or consist of two unilateral amplifiers. Due to the change of element input impedance with scan angle, bilateral amplifiers are not practical. It appears that either circulators must be used at each element, or that a frequency offset be used to obtain the necessary stability (weight and size eliminate consideration of isolators at present). Circulators of modest (20-db) isolation which would allow 15 db of gain are readily made in compact printed strip line form. In fact, a slot or dipole array could be printed with the circulators built in. Each circulator would connect to a low-gain (say 15-db) printed transistor or tunnel-diode amplifier. The gain useable is limited by feedback through the element-to-element mutual impedance. Appendix II gives a cursory look at that problem with the result that single frequency gain would be limited to 15–20 db. Phase stability must be maintained within about $\pm 10°$ through each amplifier. Deviations greater than this will cause non-negligible gain reductions. Feedback circuits would probably be indicated; no serious reduction in reliability would occur due to additional components, as will be seen later.

Since the art and literature on solid-state microwave amplifiers are extensive and are increasing rapidly, details of device configurations and of printed circuit realizations of typical amplifiers have not been included.

C. Frequency Shift Schemes

Frequency shifting through the amplifier will reduce the mutual coupling problem and may make circulators unnecessary. One frequency-shifting amplifier is the up-converter (upper sideband) where the gain obtained is equal to the up-frequency multiplication. Since it is difficult to design arrays to operate simultaneously at widely separated frequencies, the up-converter appears to be severely limited in gain. Further, the AVA system must be examined for operation under a frequency translation. Refer to Fig. 1, and observe that if the exit frequency is different from the entrance frequency, the exit angle differs from the entrance angle. In fact, the relationship is

$$\frac{f_1}{f_2} = \frac{\sin \theta_1}{\sin \theta_2} \tag{6}$$

which, for small frequency shifts Δf, reduces to

$$\frac{\Delta f}{f} = \Delta\theta \cot \theta \tag{7}$$

where $\Delta\theta$ is the beam shift angle. This is plotted in Fig. 8 for 5 and 10 per cent frequency shifts. The array beamwidth must be made large enough that the beam shift will not move the beam off the receiving point on the earth. One might take advantage of this frequency squint since a squint of 8° is desired. However the shift occurs in both θ and ϕ planes and, as seen in Fig. 8, varies with angle θ. If stabilization in one plane were available, an "average" shift could be used in the other plane, allowing use of a narrower beam, hence higher system gain.

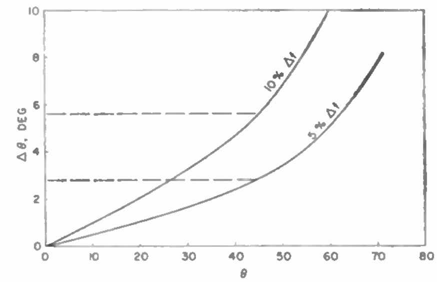

Fig. 8—Beam shift due to frequency offset.

Thus the input and output frequencies should not be too different. One scheme would use an up-converter, either variable capacitance or tunnel diode, followed by a negative resistance down-converter (for good noise figure), where the down-frequency ratio is slightly different from the up ratio. Another possibility would use an RF amplifier with a frequency translator or offset stage. With these types of schemes, it should be possible to achieve gains of the order of 70 db with wide bandwidths with a modest number of components. Bandwidth will be largely determined by power (and gain) lost due to impedance mismatches, as the element impedances change with frequency. Because these considerations are strongly affected by the mutual impedance, bandwidth is discussed along with mutual impedance in Appendix II. All schemes must preserve the phase information in the correct relationship.

D. Performance Calculations

For the "stationary" satellite case, the AVA system could be used two ways, both involving spin stabilization (some form of stabilization is desirable and spin stabilization is probably the simplest form). If the control of the orientation of the spin axis is poor, AVA arrays connected for automatic angle return in two dimensions would be disposed on a tetrahedron, rectangular box, sphere, cylinder, etc. The minimum gain, each way, for the New York-Lisbon example would be 14.9 db. A square array would contain 49 elements with 49 amplifiers.

If the spin axis can be aligned parallel to the Earth's axis, say within ±1.5°, then a disk-shaped pattern coaxial with the satellite would always illuminate Earth. Such a pattern would have 7.1-db directivity. However, an AVA array with a fixed beam in θ (cylindrical coordinates) and automatic angle return in ϕ would realize 14.9 db. A cylindrical geometry would be ideal, with several "printed" slot arrays on the cylindrical surface. Each array would have elements on each circle (of latitude) connected in Van Atta fashion. The several circles would then be connected to produce a narrow-beam response in a plane containing the axis. Automatic angle return in azimuth would be obtained as the cylinder rotates, with the fixed narrow beam in the other plane. Each array would have seven elements in the ϕ direction. Seven amplifiers would be needed. This scheme offers over the fixed disk array 7.8 db. A significant advantage of AVA is that the power levels at each amplifier will probably allow solid-state devices to be used whereas a single port amplifier would require a higher power capability.

Appendix III contains calculations for the previous example, for 1-Mc bandwidths. Fig. 9 presents vehicle output power for four cases: a completely stabilized vehicle (a very difficult job due to solar and lunar perturbations); AVA; a fixed disk pattern; a zero-db antenna. AVA is observed to compare favorably with the stabilized dish, without the attitude control system complexity.[18] For example, with a 60-foot ground dish, an AVA power output of 65-mw provides 1 Mc of PCM bandwidth. This is substantially independent of frequency, depending only on maintaining the assumed performance of antennas, noise temperatures, etc. For the spin-stabilized case above, with seven amplifiers, the output of each would be 9.2 mw (for a 60-foot ground dish). With an amplifier gain of 70 db, the input level to the amplifier would be about −60 dbm, which would allow use of crystal video vehicle receivers and low ground power.

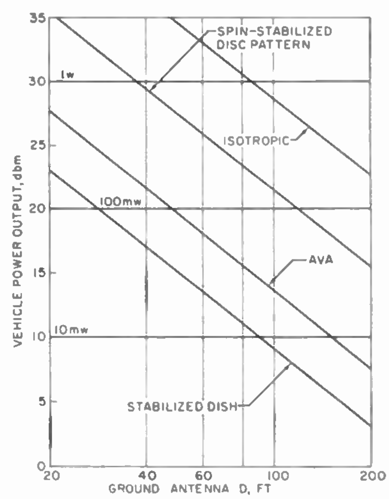

Fig. 9—Vehicle power per Mc bandwidth.

The attitude-control problem of orienting an antenna to always point at Earth is exceedingly difficult. The AVA system is within 4.6 db of the stabilized dish, and requires only spin stabilization. Further, the use of solid-state transmitters and the enhanced reliability are salient advantages.

E. Reliability

Although arrays are usually built with uniform spacing of elements, it is known that a pseudo-random spacing can be utilized in such a way that the number

[18] The effect of solar and lunar perturbations on spin-stabilized vehicles remains to be investigated.

of elements can be appreciably reduced with only minor changes in directivity. Resolution is largely determined by the over-all size as before. Now invert the problem: start with regular $\lambda/2$ spacing and let amplifiers fail in a random fashion. Roughly 40 per cent of the elements in a conventional array can fail in a random fashion without reducing gain appreciably. Sidelobe degradation is considerable, but affects gain only slightly.

The problem of calculating the probability of failure of more than a stated percentage compared to probability of failure of one item is a binomial distribution problem. The probability of exactly M out of N amplifiers working is given by

$$P = \binom{N}{M} p^M (1-p)^{N-M} \tag{8}$$

where p is the probability of a single amplifier operating, and

$$\binom{N}{M} = \frac{N!}{M!(N-M)!}$$

as usual. This problem is the same as that of a population of Rp red balls and $R(1-p)$ black balls. The probability of a random sample of size N with replacement having exactly M red balls is given by (8) above. The probability of at least M red balls, or working elements, is given by[19]

$$P = \sum_{i=0}^{N-M} \binom{N}{M+i} p^{M+i}(1-p)^{N-M-i}. \tag{9}$$

Tables are available;[20] Fig. 10 shows the variation of over-all probability with element probability.[21] These curves display the advantage of a distributed AVA: that with a 70 per cent element probability, the probability of over 60 per cent surviving is 95 per cent. Since 40 per cent can fail without serious performance degradation, we have a 95 per cent probability of satisfactory operation from an individual amplifier probability of only 70 per cent. Thus the individual reliability is considerably enhanced. Using a transistor failure rate of 0.060 per cent per 1000 hours, a single amplifier (element) can have over 25 transistors for a 70 per cent probability of lasting one year, if half the failures are due to transistors! It might be more economical, in view of the very good reliability, to trade off this against number of elements, and to use a pseudo-random array spacing with consequent reduction in number of amplifiers.

[19] W. Feller, "Introduction to Probability Theory and Its Applications," John Wiley and Sons, Inc., New York, N. Y., vol. 1, p. 172; 1950.

[20] "Tables of the Binomial Probability Distribution," Natl. Bur. Standards, AMS-6; 1952.

[21] Calculated by Richard Jaeger of Space Tech. Labs., Inc., Los Angeles, Calif.

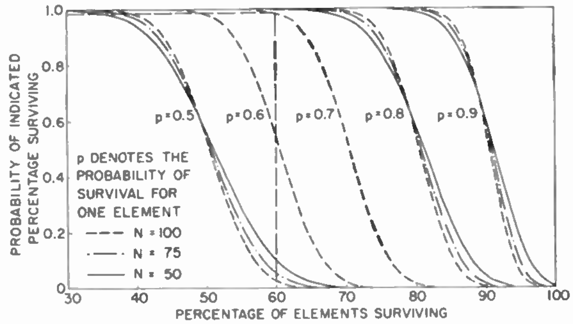

Fig. 10—AVA reliability.

VI. Comparison of Active and Passive Systems

In order to eliminate common elements such as ground antennas, transmitter powers, etc., the two-way passive range equation is compared with the product of the two one-way active equations. This gives a result relating scattering cross section to active gain, distances, etc. Changes in ground power, etc., are readily factored in. The following definitions are used:

G_1 = Ground antenna directivity.
G_2 = Satellite antenna directivity.
P_{1T} = Ground transmitter power.
P_{2T} = Satellite transmitter power.
σ = Scattering cross section.
R = Range.
P_{1R} = Ground signal into receiver.
P_{2R} = Satellite signal into receiver.

For a passive system,

$$\frac{P_{1R}}{P_{1T}} = \frac{G_1^2\lambda^2\sigma}{64\pi^3 R_1} \cdot \tag{10}$$

The up and down links of an active system are represented by

$$\frac{P_{2R}}{P_{1T}} = \frac{G_1G_2\lambda^2}{16\pi^2R^2} = \frac{P_{1R}}{P_{2T}} \cdot \tag{11}$$

Now for satisfactory operation P_{2R} must be sufficiently greater than receiver noise $kT_{eff}B$. This is usually no problem due to large ground powers readily available. The down link is usually the limiting factor due to limited P_{2T}. Very low-noise ground receiver, large antenna combinations mitigate this situation, however.

The combined active equation is

$$\frac{P_{1R}}{P_{1T}} = \frac{P_{2T}G_1^2G_2^2\lambda^4}{P_{2R}256\pi^4R^4} \cdot \tag{12}$$

For comparative purposes, let all common factors between (10) and (12) be equal. Then to make the passive and active systems equal, we must have

$$\sigma = \frac{G_3G_2^2\lambda^2R_p^4}{4\pi R_a^4} \tag{13}$$

where $G_3 = P_{2T}/P_{2R}$ is the gain of the satellite repeater amplifier.

Example: Spherical reflector and AVA

For a spherical reflector, $\sigma = \pi D^2/4$ so that

$$D = \frac{G_3^{1/2} G_2 \lambda R_p^2}{\pi R_a^2}. \tag{14}$$

For the example used previously, $f = 1$ Gc, $G_3 = 70$ db, $G_2 = 14.9$ db, we get

$$D = 30{,}600 \frac{R_p^2}{R_a^2} \text{ feet}. \tag{15}$$

A passive altitude of 2000 nautical miles and an active altitude of 19,400 nautical miles requires a balloon diameter of 325 feet even for the low (70-db) gain repeater assumed, for equal ground power. Larger ground power for the passive case would reduce the 325-foot size proportionately.

The two-way equation (12) can be rewritten,

$$P_{1T} = \frac{\frac{S}{N} B(F_N - 1) R^4}{G_3 D^4 G_2^2 (22.7 \text{ db})}. \tag{16}$$

Units are given in Appendixes I and III. For the equatorial synchronous case again, with biphase PCM and an over-all ground $F_N = 2$ db, the power per Mc bandwidth is

$$P_{1T} = \frac{(185.1 \text{ db})}{G_3 D^4}. \tag{17}$$

With a 70-db gain repeater and a 120-foot dish on the ground, 1.55-kw power is needed per Mc. Since the satellite position only changes slightly over a year's time (due to lunar effects), the 120-foot antenna could be a fixed dish with movable feed, for example. This installation would cost less than a 60-foot steerable dish.

VII. Conclusions

A passive satellite requires a high-power ground transmitter even when low altitudes are used. In contrast, the AVA allows the possibility of achieving such typical performance as 1-Mc bandwidth, 1.6-kw ground power, and 120-foot ground antenna, at 1 Gc, with only spin stabilization required for the vehicle.

The Van Atta array response is down 5.5 db at 45° and can realize a minimum gain of 14.9 db for coverage between New York and Lisbon. This 14.9-db gain is realized each way in addition to the distributed amplifier gain. An important factor is that many of the amplifiers can fail without serious performance degradation; this allows an excellent over-all reliability from lesser individual amplifier reliabilities. If each amplifier has a 70 per cent probability of lasting five years, the probability of satisfactory system operation for five years is 95 per cent. In addition, the low-power output of each amplifier allows only solid-state devices to be used.

Probably an amplifier with small frequency offset is most desirable, and it appears that these amplifiers could be built in a number of configurations, including cascaded transistors and cascaded tunnel diodes. It should be possible to achieve in the order of 70-db gain with a modest number of components in each amplifier. Phase stability of such amplifiers is as yet an unknown problem. Although the AVA distributed amplifier scheme involves a large total number of components, it should be remembered that both the array and the circuitry lend themselves to printed techniques and, in fact, such configurations could probably be made using a sandwich of printed circuit boards and components.

It is hoped that this paper will provoke a more detailed investigation into solid-state printed amplifier techniques, and into final-stage vehicle configurations, weights, etc., for AVA.

Appendix I

A Passive System

Starting with the standard range equation

$$\frac{P_R}{P_T} = \frac{G_1^2 \lambda^2 \sigma}{64 \pi^3 R^4}, \tag{18}$$

one may write

$$P_T = \frac{64\pi^3 f^2 \frac{S}{N} k T_0 B(F_N - 1) R^4 :}{G^2 c^2 \sigma}. \tag{19}$$

With proper values for the constants, the result is

$$P_T = \frac{f^2 R^4 \frac{S}{N} B(F_N - 1)}{G^2 \sigma (29.8 \text{ db})}, \tag{20}$$

where

P = power in watts,
f = frequency in gigacycles,
R = distance in natutical miles,
B = bandwidth in cps,
σ = cross section in m²,
F_N = over-all noise figure (antenna, loss, receiver noise).

The following assumptions appear reasonable.

$R = 19{,}400$,
$S/N = 8.4$ db for a 10^{-4} error rate[22]—biphase PCM,
Preamp and receiver noise figure = 1.5 db,

[22] C. R. Cahn, "Performance of digital phase-modulation communication systems," IRE Trans. on Communication Systems, vol. CS-7, pp. 3–6; May, 1959.

Antenna and loss temperature $= 50°K$ so
Over-all[23] $F_N = 2.0$ db, $F_N - 1 = -2.3$ db.

For an antenna of 50 per cent aperture illumination efficiency,

$$G = D^2 f^2 (7.2 \text{ db}) \tag{21}$$

with D in feet, f in Gc. Then

$$P_T = \frac{(163.5 \text{ db})}{D^4 f^2 \sigma}. \tag{22}$$

For a 100-foot sphere, $\sigma = 2920$ m². For the example of a 14.9-db Van Atta array,

$$\sigma = \frac{(8.3 \text{ db})}{f^2}. \tag{23}$$

So the ground power needed, assuming constant F_N, decreases with frequency squared for Echo, and is constant for the array. The above calculations do not include various losses, nonoptimum equipment performance, etc., but these can readily be included.

Appendix II

Element Input Impedance and Bandwidth

The input impedance of each element of an array is conveniently represented in the conventional fashion as the combination of a self and a mutual impedance. The self impedance is not that of the element in the absence of all other elements, but rather that of the element with all other elements present and open circuited, as the capacitively loaded elements will have a non-negligible effect. The mutual impedance can accurately be taken as that between adjacent elements only. For quarter-wave monopoles or half-wave dipoles, the input self-reactance is substantially independent of the antenna diameter; however, the resistance is a function of diameter. For dipole elements of practical size, a typical single-element self impedance is $73 + j42$ ohms. As mentioned, this should be corrected for the presence of other dipoles. Slot impedance can of course be found for the complementary slot by Babinet's principle. The mutual impedance between half-wave dipoles is strongly dependent upon whether the elements are parallel or collinear. King quotes an impedance for parallel half-wave dipoles[24] with $0.5\ \lambda$ spacing of

$$Z_{12} = 12 - j30 \text{ ohms}$$

while for collinear half-wave dipoles with spacing $0.8\ \lambda$, he quotes[24]

$$Z_{12} = -0.6 - j7 \text{ ohms}.$$

Thus the mutual impedance for parallel elements is nearly five times larger than for collinear elements. Unfortunately, as the direction of arrival of the incident wave changes, so does the relative phase between adjacent elements in the array, and in the equivalent circuit, this is equivalent to changing the phase angle of the load impedance. A simple calculation of the power transferred from one element to the adjacent element through the mutual impedance yields the formula

$$\frac{P_2}{P_1} = \frac{R_{11} c^2}{2a|2a^2 \cos\alpha - c^2 \cos(\beta - 2\gamma)|}, \tag{24}$$

where c and γ are the magnitude and angle of the mutual impedance, a and α are the magnitude and angle of the self impedance, and β is the added phase shift. The usual conjugate impedance match is assumed. For the worst case of half-wave parallel dipoles, the fraction of power coupled varies from -14.3 db to -14.5 db to -14.7 db as the phase angle changes from -45 to 0 to $+45°$. For the collinear pair, the broadside coupling is down -27.8 db. Previous experience has shown that the use of baffles, etc., can reduce coupling between parallel dipoles to a value near that for colinear dipoles.[25] Thus it appears that loop gain useable at a single frequency is limited to about 20 db.

A more careful investigation of the change in input impedance with scan angle is that of Carter.[26] Fig. 11 is reproduced from his work. This shows the input impedance vs scan angle for an element in a finite array. It appears that for scan angles less than 45° the input impedance changes are sufficiently small as to not constitute a serious problem.

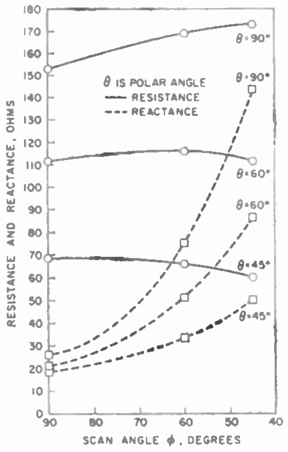

Fig. 11—Impedance vs scan angle of element 31 of half-wave, dipole array with reflector (after Carter).

[23] Note that it might be simpler to use effective noise temperature exclusively, however, many articles use noise figure.

[24] R. W. P. King, "Linear Antennas," Harvard University Press, Cambridge, Mass., pp. 440 and 287; 1956.

[25] I. P. Kaminow and R. J. Stegen, "Waveguide Slot Array Design," Microwave Lab., Hughes Aircraft Co., Culver City, Calif., Tech. Memo. 348, p. 73; July, 1954.

[26] P. S. Carter, Jr., "Mutual impedance effects in large beam scanning arrays," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 276–285; May, 1960.

Calculations of input impedance vs frequency are straightforward but tedious; results have been computed by Blasi[27] for an element in a parallel array and in a collinear array. From these data, it may be observed that the impedance variation is slow with frequency, and it should be possible to impedance compensate fairly simply to cover bandwidths as wide as 20 per cent or more.

Appendix III

An Active System

For the satellite-to-ground link, always the limiting factor,

$$\frac{P_T}{P_R} = \frac{G_1 G_2 \lambda^2}{16\pi^2 R^2} \tag{25}$$

which becomes

$$P_T = \frac{16\pi^2 \frac{S}{N} kT_0 B(F_N - 1)R^2}{G_1 G_2 \lambda^2} \tag{26}$$

which reduces to

$$P_T = \frac{\frac{S}{N} B(F_N - 1)f^2 R^2}{G_1 G_2 (106.1 \text{ db})}. \tag{27}$$

[27] E. A. Blasi and R. S. Elliott, "Scanning antenna arrays of discrete elements," IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 435–436; October, 1960. Also "Effects of Mutual Interaction on the Design of Various Dipole Arrays," Microwave Lab., Hughes Aircraft Co., Culver City, Calif., Tech. Memo. 336; December, 1953.

Units are the same as in Appendix I; G_1 is ground-antenna directivity, G_2 is vehicle-antenna directivity. Assumptions will be the same; the minimum AVA gain of 14.9 is used.

$$S/N = 8.4 \text{ db},$$
$$F_N - 1 = -2.3 \text{ db},$$
$$G_1 = D^2 f^2 \ (7.2 \text{ db}),$$
$$G_2 = 14.9 \text{ db}.$$

Then

$$P_T = \frac{(53.7 \text{ db})}{D^2}. \tag{28}$$

D in feet, for a 1 Mc bandwidth. Table I shows the results. The disk pattern to cover Earth with ±1.5° stabilization has a 20° beamwidth, and a 7.1-db directivity. So it is 7.8 db poorer than the AVA. A completely stabilized platform with 20° pencil beam has a 17.2-db directivity, or 4.6 db better than AVA.

TABLE I

Vehicle Power per Mc Bandwidth

	Stabilized Dish	AVA	Disk Pattern	Isotropic
D = 60 feet	5.6 mw	64.5 mw	390 mw	2 w
D = 120 feet	22.4 mw	16.2 mw	98 mw	500 mw

Acknowledgment

It is a pleasure to mention discussions with Dr. V. Rojansky of Space Technology Laboratories, Inc., and with G. Rove and R. G. Stephenson of Aerospace Corporation. M. L. Buschkotter had the patience to assist in several cycles of manuscript preparation.

Correspondence

Nonreciprocal Parametric Amplifier Circuits*

Two types of cavity difference frequency parametric amplifiers have been discussed extensively in the literature: the two-port transmission amplifier and the single-port reflection amplifier which uses a circulator to achieve separation of the input and output signals. Of these two, the circulator amplifier (shown in Fig. 1) has been shown to have a number of advantages over the two-port amplifier:[1] an increase in the voltage gain-bandwidth product by a factor of at least two, the isolation of the amplifier from the noise of the output termination, and an improvement in stability with respect to variations in source and load impedances.

Unfortunately, the characteristics of available circulators are often unsatisfactory for parametric-amplifier use because of excessive insertion loss, restricted bandwidth, or large size. At low frequencies, circulators are presently not available.

The purpose of this letter is to point out two circuit configurations which, while not using a circulator, have similar properties to the single-port reflection amplifier with a three-port circulator. These circuits are shown in Figs. 2 and 3. Both use two time-varying reactances with a relative phase shift of 90° between them. In addition, the signal frequency is also shifted by 90° at the two reactances. A common idler circuit resonant at $\omega_l = \omega_p - \omega_s$ is used.

In the circuit of Fig. 2, a 90° hybrid is used to split the incoming signal into two equal parts which then emerge from ports 2 and 4. The signal phase at port 4 leads that at 2 by 90°. Since a similar phase shift is present in the pumping, idler currents of the same phase are generated and dissipated in the idler load. Thus, the signal power at ports 2 and 4 is amplified and reflected to emerge from port 3.

In the reverse direction, power entering port 3 is split between ports 2 and 4 but with the phase at port 2 leading that of port 4 by 90°. In this case the idler currents generated are 180° out of phase, and they cancel in the idler resonant circuit. The signal power is then reflected at ports 2 and 4 without amplification and combines to emerge from port 1. The circuit is thus seen to be nonreciprocal with a power gain in the forward direction and a gain of slightly less than unity in the reverse direction. A more quantitative analysis shows that with the exception of the slight difference in reverse gain, the performance of the hybrid amplifier is identical to that of the circulator amplifier.

The circuit shown in Fig. 3 uses a 90° phase-shift network of characteristic conductance $g_{s3}+g_{o3}$ to obtain the required phase shift at the signal frequency. This network may consist of a quarter wavelength of transmission line or a lumped constant equivalent network. Here the phasing is correct for idler cancellation in the reverse direction only at the center frequency. However, in the important practical case where the idler circuit bandwidth is much smaller than that of the signal circuit, the same voltage gain-bandwidth product and a reverse gain of slightly less than unity are also obtained with this circuit.

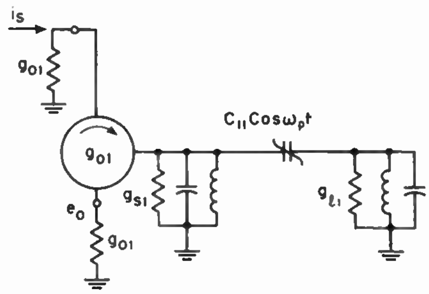

Fig. 1—Circulator amplifier.

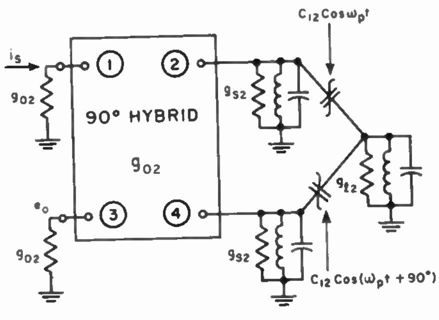

Fig. 2—Hybrid amplifier.

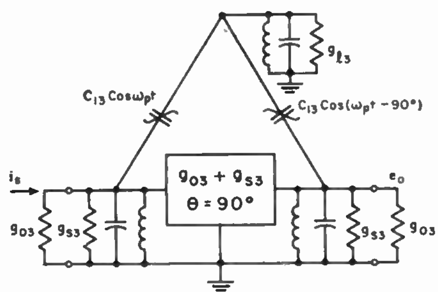

Fig. 3—90° network amplifier.

For large gain the equivalent input noise temperature for the three circuits is given by

$$T_e = T\left[\frac{g_s}{g_o} + \frac{\omega_s(g_o + g_s)}{\omega_l g_o}\right]. \tag{1}$$

However, large gain also requires that

$$\frac{g_{s1}}{g_{o1} + g_{s1}} = \frac{g_{s2}}{g_{o2} + g_{s2}} = 2\frac{g_{s3}}{g_{o3} + g_{s3}} = \frac{4\omega_s\omega_l}{\sigma^2\omega_c^2}, \tag{2}$$

where $\sigma = C_1/C_0$ and ω_c is the varactor angular cutoff frequency. Thus, for varactors of equal quality, the 90° network amplifier of Fig. 3 has a somewhat higher equivalent input noise temperature because of a more unfavorable ratio of g_{s3} to g_{o3}. This difference in noise temperature is dependent on the relative size of the two terms in (1) and may be quite small in some cases.

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* Received by the IRE, January 20, 1961; revised manuscript received, March 3, 1961.

[1] A. E. Siegman, "Gain bandwidth and noise in maser amplifiers," Proc. IRE, vol. 45, pp. 1737–1738; December, 1957.

A Millimeter-Wave Esaki Diode Amplifier*

Oscillation of Esaki diodes at fundamental frequencies in excess of 100 kMc[1] has stimulated the use of these diodes as amplifiers at millimeter wavelengths. Amplification with high gain has now been attained with Esaki diodes at frequencies ranging from 55 to 85 kMc (5.5- to 3.5-mm wavelength). The amplifier closely resembles the millimeter-wavelength oscillator previously described,[1] with the essential differences being in its adjustment and connection in the circuit.

"Formed" point-contact Esaki diodes of n-type gallium arsenide were mounted in reduced-height rectangular waveguide having a LF cutoff of 48 kMc, as shown in Fig. 1.

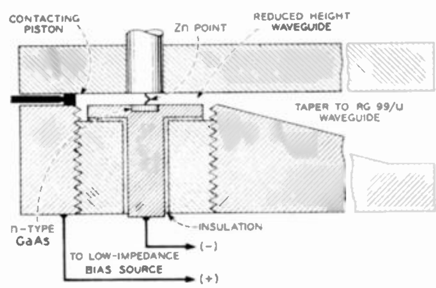

Fig. 1—Cross section of a millimeter-wave Esaki-diode amplifier circuit. The waveguide was RG 99/U reduced in height to 0.01 inch.

The frequency of operation was determined largely by the length and shape of the zinc contact and by the conditions under which the diode was formed. Oscillations were suppressed by the positioning of a contacting piston. The piston also tuned the amplifier frequency to some extent.

A block diagram of the test circuit in the 55-kMc frequency range appears in Fig. 2. The input and output of the reflection am-

* Received by the IRE, April 10, 1961.

[1] C. A. Burrus, "Millimeter wave Esaki diode oscillators," Proc. IRE (Correspondence), vol. 48, p. 2024; December, 1960. Also, "Gallium arsenide Esaki diodes for high frequency applications," J. Appl. Phys., to be published in June, 1961.

plifier are separated advantageously with a circulator. The amplifier showed high gain at piston positions for which the circuit was on the verge of oscillation, and the gain was adjusted to the desired level with tuning stubs between the amplifier and the circulator.

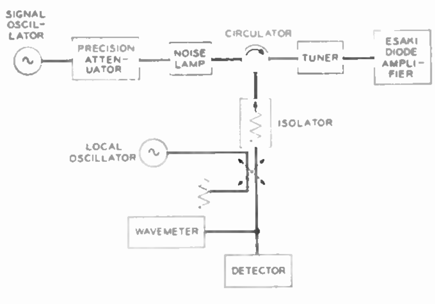

Fig. 2—Block diagram of the 55-kMc test circuit used for noise and gain measurements of the amplifier.

Noise figures were measured near 55 kMc with a special calibrated argon noise lamp[2] having an excess noise ratio of 14 db. To avoid saturation of the amplifier by local oscillator leakage during noise measurements, at least 60 db of isolation was maintained between the circulator and the local oscillator input.

The performance of this amplifier was similar to that previously reported at lower microwave frequencies.[3] Stable gains as high as 35 db were attained with gain compression appearing at a power output of about −35 dbm (0.3 μw). The bandwidth was approximately 40 Mc at a peak gain of 20 db. The highest frequency at which amplification was observed was 85.5 kMc. This limit was imposed by the available test equipment, however, and not by the diodes or the amplifier circuit itself.

Near 55 kMc, the noise figure of the amplifier exclusive of the circulator was found to range from 16 to 18 db. Because of large circuit losses, these noise figures were considerably higher than the diode shot noise contribution, which was about 7 db for the particular diodes employed in the noise measurements.

Although the amplifier was not of optimum design, its operation has demonstrated that extremely simple solid-state Esaki diode circuits can amplify with substantial gain at high microwave frequencies. In passing, it is worth noting that the noise figure of this amplifier is comparable to that of a conventional 55 kMc receiver having a similar bandwidth.

The authors are indebted to W. M. Sharpless for the noise measurements and to E. H. Turner and A. P. King for the loan of many of the necessary millimeter wave components.

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[2] W. W. Mumford, "Gaseous discharge source," in "Millimeter Wave Research," Bell Telephone Labs., Inc., New York, N. Y., Final Rept. No. 24261-15, Contract Nonr-687(00), pp. 111–113; 1955.

[3] R. Trambarulo, "Esaki diode amplifiers at 7, 11, and 26 kmc," Proc. IRE (Correspondence), vol. 48, pp. 2022–2023; December, 1960.

Two-Wire Millimeter-Wave Surface Transmission*

The purpose of this note is to report some characteristics of open-wire lines in the millimeter-wave region. Open-wire lines such as single-wire or two-wire lines have been practically abandoned for frequencies higher than 10 kMc. Two-wire lines have been considered for frequencies below 100 Mc, and single-wire lines for frequencies above 100 Mc.[1]

Recently, it has been found that these open-wire lines might be helpful for millimeter-wave communication. The high attenuation due to atmospheric absorption of millimeter-wave energy ordinarily makes communication difficult. Open-wire lines may help to solve this problem.

When No. 28 enameled wire is connected between two 15-db-gain pyramidal horn antennas located one meter apart and facing each other, the received detected output voltage increased to three times the value observed without the wire. When another wire was added in parallel to the former wire, this two-wire line increased the voltage 5.6 times the value observed without wires. The spacing of the wires was 9.5 mm and the operating frequency was 58 kMc. A two-wire line with 4.5-mm spacing showed 10 db less attenuation than the single-wire line, and a two-wire line of 2-mm spacing showed 20 db less attenuation than a single-wire line at 59 kMc. No. 26 wire was the optimum size for minimum attenuation. This wire had 5 db less attenuation than No. 38 wire and 3 db less attenuation than No. 16 wire. When a polystyrene sheet of 0.015-inch thickness was attached between the two wires of 2 mm spacing, the attenuation increased 7 db. The stranded-silk-covered two-wire line of 2-mm spacing showed a 3.5 db increase of attenuation when it was dry and a 25.7 db increase for the wet condition, in comparison with the No. 28 enameled two-wire line of 2-mm spacing. When the silk cover was removed, the attenuation was almost equal to the enameled wire line.

The field extent was measured by measuring the attenuation due to an obstacle of 15-mm×180-mm area conductor placed to disturb the transmission. The experimental results are shown in Fig. 1. According to these results, the field extent was 8 mm for the single-wire line and 3.5 mm for the two-wire line of 4.5-mm. spacing. The field extent decreased to 2.2 mm when the spacing of the wires was 2.2 mm. The test frequency was 58.29 kMc and the wavelength was 5.15 mm.

It is interesting to note that these field extents are very small in comparison with the value of $\frac{3}{4}$ inch reported by Wiltse for the dielectric image line operated at 70 kMc measured in a similar way.[2] As Goubau has remarked,[1] two-wire ines exhibit better

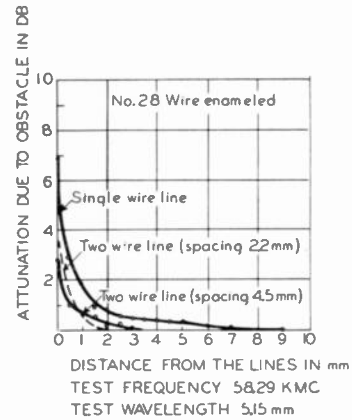

Fig. 1—Field extent.

power concentration than comparable single-wire lines.

The theoretical work relating to the two-wire line has yet to be completed. It appears that the line operates in a hybrid mode in which longitudinal components of both the electric and magnetic fields are present, but that an approximate solution can be obtained by assuming a transverse magnetic mode. The nature of the launching device suggests a transverse electric mode, but an analysis has indicated that the lowest possible mode is rapidly attenuated and that the next possible mode, the dipole mode, would exhibit a field extent much larger than that observed. It appears that an analysis based on the TM_{00} mode, which exists on the Goubau line, will yield good approximations for attenuation and field extent.

The authors wish to thank the Raytheon Company for contributing the klystron tubes as a signal source.

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* Received by the IRE, March 20, 1961.

[1] G. Goubau, "Open wire lines," IRE Trans. on Microwave Theory and Techniques, vol. MTT-4; pp. 197–200; October, 1956.

[2] J. C. Wiltse, "Some characteristics of dielectric image lines at millimeter wavelengths," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 66–69; January, 1959.

Computer Simulation of a Television Coding Scheme*

This letter reports initial results on the performance of a television channel-compressing scheme using "dual-mode" transmission.[1] The scheme is easily implemented with pulse code modulation. For a "bit-rate" reduction of approximately 2:1, pictures obtained by computer simulation show varying amounts of degradation depending on picture material.

Past efforts aimed at reducing the channel capacity required for TV transmission by removal of redundancy have met with at least two major obstacles: the need for complex instrumentation, even for testing some

* Received by the IRE, March 23, 1961.

[1] E. R. Kretzmer, "Television System Having Reduced Bandwidth," U. S. Patent No. 2,946,851; July 26, 1960.

of the schemes; and the extreme difficulty of realizing a channel capacity reduction in the form of bandwidth saving. Three new factors have entered the scene in recent years: the utilization of certain tolerances of the human visual mechanism;[1,2] increasing interest in the use of pulse code modulation;[3] and the possibility of testing proposed systems by digital computer simulation.[4]

The tolerance in the visual mechanism mentioned above has been demonstrated in a transmission system which separates the low and high video frequencies and quantizes the "highs" much more coarsely than the "lows."[2] This is permissible because we tolerate much less accurate amplitude rendition of picture areas containing fine detail or edges than of regions of quasi-uniform luminance. The dual-mode scheme reported here also effects coarse quantization of fine detail, but uses time division instead of frequency division. At the transmitter, the video signal is sampled in the usual way at a rate twice its highest frequency. A running decision is then made as to whether or not the present sample qualifies as fine detail. The criterion is whether it differs from the preceding sample by more than a predetermined magnitude—six per cent of the video range for the results reported here. If so, the sample is encoded coarsely—to only four binary digits; if not, full seven-bit encoding is used, but every other sample in a low-detail area is omitted—to be approximated at the receiver by interpolation. A method for informing the receiver which mode is being transmitted has been described,[1] but was not included in the simulation. Its inclusion requires equal blocks of, say, eight digits for both modes (including one mode-identification bit) and would cause some further degradation which is believed to be small.

The present encoding uses four bits per sample instead of the normal seven bits, at the expense of some picture quality. Fig. 1 shows picture material in dual-mode processed form, while Fig. 2 shows the same material in seven-bit representation (considered as perfect). These pictures have been obtained by programming the IBM 7090 digital computer to perform the operations specified by the processing scheme on a large number of numerical data, representing the luminance values of all picture elements. The computed values appear in digital form on an output magnetic tape which is then played back on a special monitor.[4] Each picture contains approximately 10,000 samples and is scanned in 2.4 seconds.

There are two important effects which are not shown by this computer simulation method; namely, the effects of repeated scanning of still pictures and the effects of motion. Therefore, computer pictures can be used to disqualify poor systems but do not necessarily indicate the best system.

Fig. 1—Picture material in dual-mode processed form.

Fig. 2—Same picture material quantized to 7 bits per sample.

Acknowledgment

Thanks are due to E. E. Sumner and J. S. Mayo for their guidance, H. S. McDonald for helpful suggestions and the use of the pictures and playback equipment, W. A. Anderson for taking the pictures, and M. J. Magelnicki for programming the computer.

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[2] E. R. Kretzmer, "Reduced alphabet representation of TV signals," 1956 IRE Convention Record, pt. 4, pp. 140–147.
[3] R. L. Carbrey, "Video transmission over telephone cable pairs by pulse code modulation," Proc. IRE, vol. 48, pp. 1546–1561; September, 1960.
[4] R. E. Graham and J. L. Kelly, Jr., "A computer simulation chain for research on picture coding," 1958 IRE WESCON Convention Record, pt. 4, pp. 41–46.

Secondary Electron Emission and the Satellite Ionization Phenomenon*

The observations of J. Kraus and colleagues at The Ohio State University on large anomalous electromagnetic scattering from artificial earth satellites have been extensively reported.[1,2] Most of these observations have been obtained using reflected signals from WWV at 15, 20, and 25 Mc. These observations are assumed to be connected in some manner with a satellite-induced ionization phenomena in the medium traversed by the vehicle.

Several independent and distinct mechanisms have been proposed and analyzed for the production of local electron and ion densities varying from ambient in the neighborhood of a satellite. One of the earliest models was the result of the theory of electrohydrodynamic interactions of a charged satellite with the ionosphere developed originally by Kraus and Watson.[3] The model was subsequently expanded and analyzed from a somewhat different viewpoint by Rand[4] and Pappert.[5] A theory to account for the ionization buildup of a traveling cloud or satellite ghost has been proposed by Walker and Singer.[6] Magnuson and Medved[7] have considered the question of sputtering as having some bearing on ionization buildup for vehicles below 200 km.

It seems most unlikely that any unique simple physical model can completely account for all of the observations reported, particularly in view of the large variation in the altitude regime involved. It is the purpose of this note to outline a model for ionization buildup based on secondary electron emission and to delineate two altitude domains for its validity. The term "secondary electron emission" is used here in a rather broad and general sense to signify the emission of electrons from the vehicle surface initiated by any neutral species such as: 1) neutral atoms and molecules in the ground or excited states, or 2) energetic photons above the photoelectric threshold ($\lambda < 4000$ A). The two altitude regimes of interest are defined by process 1) to lie between 90 and 250 km and by process 2) to correspond to altitudes greater than 1000 km during exposure of the satellite to solar radiation. These domains correspond rather roughly to the regions where Kraus' observations have been reported.

Emission of electrons from solids due to ion impact has been beautifully treated in an extensive series of papers by Hagstrum[8] in terms of the Auger emission process. The yield of secondary electrons γ is relatively independent of the ion kinetic energy and depends mainly on the ionization potential of the bombarding atom and the work function of the solid. Typical yield values obtained by Hagstrum are $\gamma \sim 10^{-1}$ for ion energies ranging from 10 ev to several hundred ev for clean surfaces. We here assume that some fraction of the neutral species in the upper atmosphere (alt <250 km) are in metastable states. The satellite is exposed to the particle stream with impact energies on the order of 10 ev. The emission of an electron from such an impact process proceeds in an analogous manner to the positive ion case here via the process of Auger de-excitation. There is some experimental evidence that γ values for ions and metastables are similar. Effective values of $\gamma \gtrsim 10^{-3}$ for the upper atmosphere appear plausible if we assume that at least one in a hundred of the neutral atoms are metastables. As a consequence of $\gamma \gtrsim 10^{-3}$ satellite

* Received by the IRE, April 3, 1961.
[1] J. D. Kraus, and E. E. Dreese, "Sputnik I's last days in orbit," Proc. IRE, vol. 46, pp. 1580–1587; September, 1958.
[2] J. D. Kraus, R. C. Higgy, and W. R. Crone, "The satellite ionization phenomenon," Proc. IRE, vol. 48, pp. 672–678; April, 1960.
[3] L. Kraus, and K. Watson, "Plasma motions induced by satellites in the ionosphere," *Phys. Fludis*, vol. 1, pp. 480–488; November–December, 1958.
[4] S. Rand, "Wake of a satellite traversing the ionosphere," *Phys. Fluids*, vol. 3, pp. 265–273; March–April, 1960.
[5] R. A. Pappert, "Application of Bohm-Pines Theory to Plasma Motion," Phys. Sec., Convair, San Diego, Calif., Rept. No. ZPh-026; December, 1958.
[6] E. H. Walker, and S. F. Singer, "Wake of a charged body moving in a plasma," *Bull. Am. Phys. Soc.*, vol. 5, p. 234; April, 1960.
[7] G. D. Magnuson, and D. B. Medved, "Sputtering as a Mechanism for Increase of Ionization in the Vicinity of Low Altitude Satellites," Phys. Sec., Convair, San Diego, Calif., Rept. No. ZPh-050, 11-227; December, 1959.
[8] H. D. Hagstrum, "Auger ejection of electrons from metals by ions," *Phys. Rev.*, vol. 96, pp. 336–365; October, 1954.

surfaces should charge positive at altitudes less than 200 km with a resulting electron sheath formed around them.

The emission of electrons can be analyzed for several assumed conditions. For an infinite flat plate bombarded by neutral particles where the ejected electrons are monoenergetic, the electron density distribution surrounding the vehicle follows the Langmuir-Childs Law for the plane parallel diode where the equivalent of the second plate is given by the turn-around point of the electrons trapped in the potential well formed by the positive surface charge buildup on the satellite.

The electron density variation $n(x)$ is given by:

$$n(x) = Bx^{-2/3}$$

$$B = \left(\frac{8}{9}\right)^{1/3}\left[\frac{m\epsilon_0[j(x_s)]^2}{e^2}\right]^{1/3} \text{(mks units)} \quad (1)$$

and the effective sheath thickness x_s (which can be considered as an effective Debye distance characteristic of the problem) is given by

$$x_s^2 = \frac{m[v(x_s)]^2}{n(x_s)e^2/\epsilon_0}. \quad (2)$$

where x_s is the distance from the satellite to the electron turn-around point, $j(x_s)$ is the electron current density emitted at the satellite surface, and $v(x_s)$ is the initial velocity of the emitted electrons.

Finally in terms of assumed values of $v(x_s) = 1.6 \times 10^8$ cm/sec and satellite velocity of 8×10^5 cm/sec we have

$$n(x_s) = 5 \times 10^{-3}\gamma\rho_n \text{ in cm}^{-3} \quad (3)$$

$$x_s^2 = 1.6 \times 10^9(\gamma\rho_n)^{-1} \text{ in cm}^2, \quad (4)$$

where ρ_n is the neutral density.

The variation of the sheath thickness and the electron concentration at the surface with altitude are plotted in Fig. 1. Above 1000 km where the satellite may go positive due to photoemission, there will be little variation with altitude depending only on the flux of ultraviolet solar radiation.

For electron densities relevant to 10-Mc radiation and higher, the sheath thickness is less than 10 cm. It can then be shown that for a spherical or cylindrical satellite, the plane parallel analysis and equations given are quite valid. A detailed discussion of these questions is given by Langmuir and Compton[9] where it is shown that for equidistant curved surfaces having a radius of curvature large compared to their separation the problem may be treated to a high degree of accuracy as though they are plane surfaces. Although the thickness of the sheath is small compared to the satellite dimensions, such a configuration may resonantly scatter in the Rayleigh limit. The case of a uniform spherical plasma shell surrounding a metal sphere has been analyzed by Pappert[10] and in the long wavelength limit it is found that for back scattering and 20 Mc waves, the cross section for such a case would be on the order

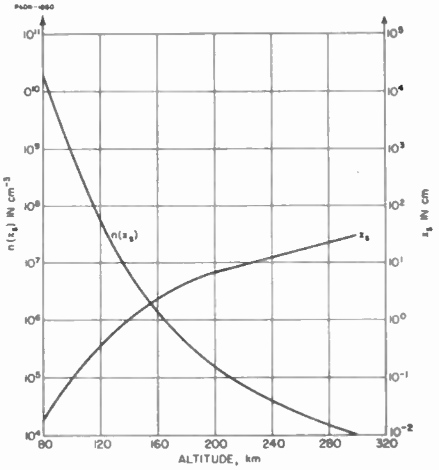

Fig. 1—Electron density $n(x_s)$ and sheath thickness (x_s) as a function of altitude. Assume $\gamma = 10^{-2}$ and constant.

of 100 square meters. Several important considerations must still be resolved before these mechanisms can be invoked in an explanation or prediction of the satellite ionization phenomenon in a particular altitude regime; it is essential that an understanding of actual secondary electron yields and velocity distributions for 10-ev particles be achieved. The effects of the satellite field and electron density and velocity gradients in the electron shells on the calculated resonant scattering must also be determined. Another problem which should be considered is whether the relatively large velocities (10^8 cm/sec) of the electrons at and near their points of origin on the satellite surface can somehow give rise to Doppler shifts.

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[9] I. Langmuir and K. T. Compton, "Electrical discharges in gases II. Fundamental phenomena in electrical discharges," *Rev. Mod. Phys.*, vol. 3, pp. 191–257; April, 1931.

[10] R. A. Pappert, "Rayleigh Scattering Associated with Possible Shock Formation in a Collisionless Plasma," Phys. Sec., Convair, San Diego, Calif., Rept. No. ZPh-062, III-29; June, 1960.

Tunable S-Band Traveling-Wave Maser for Telemetry Systems*

Traveling-wave masers,[1–4] utilizing distributed structures and yielding nonreciprocal gain have resulted in a significant advance in the state-of-the-art. The purpose of this note is to report on a ruby traveling-wave maser that is electronically tunable over an 11 per cent band (2120 to 2370 Mc) having a net gain of 30 db with an amplification bandwidth of 21 Mc.

The maser utilized a comb slow-wave structure[1] that was loaded with X-ray oriented ruby (nominal chromium concentration of 0.05 per cent) on both sides of the comb.[4] The active length of the structure was 6½ inches. The magnetic fields and pump frequencies that were required for three-level maser operation varied from 2.38 to 2.48 kilo-oersteds and 12,420 to 12,730 Mc, respectively; the incident pump power required for saturation was about 100 mw.

The net forward gain of the maser, measured as a function of operating frequency at a bath temperature of 1.8°K, is plotted in Fig. 1. From this figure it is clear that the maser gain is nearly constant across the tuning range, and there are no regenerative effects in the amplification band. The slow-wave structure incorporated polycrystalline yttrium-iron garnet (YIG) disks to obtain the nonreciprocal reverse loss which is necessary for amplifier stability.[1] The net reverse loss of the maser was in excess of 100 db, thus insuring short-circuit stability.

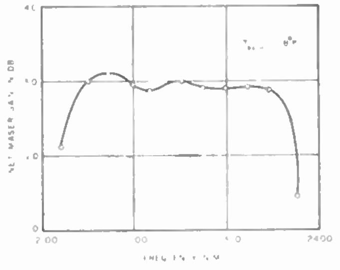

Fig. 1—Net maser gain vs operating frequency.

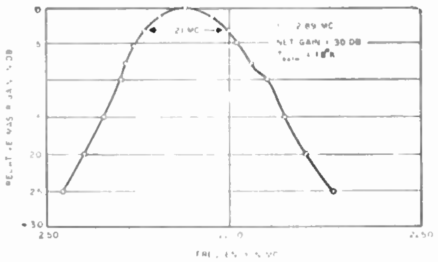

Fig. 2—Typical maser bandwidth.

Fig. 2 is a plot of a typical bandwidth curve measured at 2189 Mc. It can be seen that there is a slight "hump" on the high-frequency side of the amplification band. Furthermore, the measured bandwidth is slightly greater than predicted by the nominal 20-oersted ruby linewidth.[5] These are the results of a stagger-tuning effect caused by slight variations in the relative C-axis orientation of the four ruby slabs used.

Measurements on the over-all effective receiver input noise temperature, (T_e), using

* Received by the IRE, March 31, 1961. This work was supported by the U.S.A.F., Aeronautical Systems Div., under Contract AF33(600)-38862.

[1] R. W. DeGrasse, E. O. Schulz-DuBois, and H. E. D. Scovil, "Three level solid-state traveling-wave maser," *Bell Sys. Tech. J.*, vol. 38, pp. 305–334; March, 1959.

[2] E. O. Schulz-DuBois, H. E. D. Scovil, and R. W. DeGrasse, "Use of active material in three-level solid state masers," *Bell Sys. Tech. J.*, vol. 38, pp. 335–352; March, 1959.

[3] W. S. C. Chang, J. Cromack, and A. E. Siegman, "Cavity and Traveling Wave Masers Using Ruby at S-band," Electronics Labs., Stanford University, Stanford, Calif., Tech. Rept. 155-2; July 28, 1959.

[4] S. Okwit, F. R. Arams, and J. G. Smith, "Electronically tunable traveling-wave masers at L and S bands," Proc. IRE, vol. 48, pp. 2205–2206; December, 1960.

[5] J. E. Geusic, R. W. DeGrasse, E. O. Schulz-DuBois, and H. E. D. Scovil, "The Three Level Solid-State Maser," Bell Telephone Labs., Murray Hill, N. J., 9th Interim Rept. on Microwave Solid-State Devices, U. S. Army Signal Corps Contract DA-36-039-sc-73224, pp. 7–10; August, 1958.

a hot and cold load noise generator in a Y-factor measurement, yielded a $T_e = 18°K \pm 2°K$, of which 8.5°K was due to the noise contributions of a relatively poor second-stage amplifier (14.5-db noise factor). Thus, the T_e of the maser proper plus its input feed line losses is less than 10°K.

Measurements on the gain characteristics have shown a gain saturation (deviation of 3 db from linearity) at an output power of −27.5 dbm.

Tuning of the maser is accomplished by varying a single dial which simultaneously adjusts the applied magnetic field (current through an electromagnet) and the pump frequency (delay line voltage of a backward-wave oscillator). No mechanical tuning is necessary because of a multiple-resonance effect in the pump transmission line at the pumping frequency.

Further work is in progress, and a complete report on the design considerations of the final packaged system capable of military field use will be presented at a later date.

The authors gratefully acknowledge the help of J. Wolczok and J. Baris in the construction and measurements.

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Measurements on Resonators Formed from Circular Plane and Confocal Paraboloidal Mirrors*

In a recent letter by Fox and Li[1] an analysis of the resonator modes in resonators formed from circular plane mirrors and confocal paraboloidal mirrors was presented. Their analysis was in connection with an interest in optical masers. This note is to call attention to the same results obtained in work connected with guided waves and beams.

Resonances between circular-plane conducting surfaces was discovered some years ago in a resonator being used to measure the loss of the Sommerfeld surface wave on a cylindrical conductor.[2,3] The resonance appeared as a spurious response along with the expected resonance of the Sommerfeld wave. A typical case is shown in Fig. 1. It is easily shown that to obtain such an over-all response curve requires that the interfering

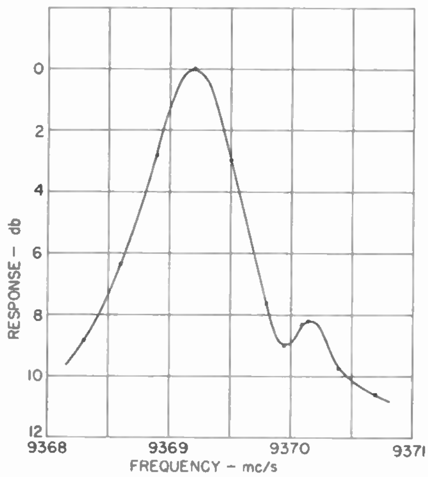

Fig. 1—Response of the Sommerfeld wave on a cylindrical conductor along with the spurious response in a resonator formed by two plain circular end plates.

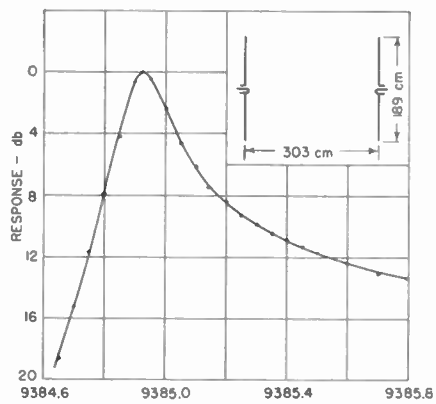

Fig. 2—Response curve of a resonator formed by two plain circular plates.

response must have a higher Q than the desired response. No resonance with a higher Q than that associated with the Sommerfeld wave was known to exist in such a resonator at that time.

In an attempt to identify the spurious response in the Sommerfeld wave resonator, the copper wire between the flat circular plates was removed. Very high Q responses were found to occur for certain spacings of the loop coupled parallel plane resonator. The resonator was rather large, making it difficult to adjust, and hence limited the completeness of the experimental investigations. A typical response curve for the parallel-plane resonator is shown in Fig. 2. The curve is unsymmetrical about the center frequency and hence the phenomenon was called "Pseudo-Resonance."[4] The particular case shown in Fig. 2 was for a resonator end plate diameter of 189 cm, a resonator length of 300 cm, and an operating wavelength of 3.2 cm. Using the notation of Fox and Li, a for the radius of the circular mirrors and b for resonator length the ratio $a^2/b\lambda$ is 9.4, the measured Q for the resonator was 60,000 or a loss of 0.0426 db. This corresponds to a loss per transit of 1 per cent. This point lies on the curve for circular plane mirrors given by Fox and Li in their Fig. 2. At the time these measurements were made the analysis of the resonance phenomena was not put in the form given by Fox and Li.

The second curve given by Fox and Li in their Fig. 2 is for confocal paraboloidal mirrors and is the same curve given by Goubau and Christian in connection with their work on the beam waveguide for millimeter waves.[5] The curve represents the loss per iteration for the beam waveguide and is presently being verified experimentally by the resonator technique. The resonator is formed by using a paraboloidal mirror at one end and a plane mirror at the other. The plane mirror is set at the focal point of the paraboloid. This resonator simulates one iteration along the beam waveguide and is equivalent to a resonator made up of two confocal paraboloids.

Measurements on this resonator at a wavelength of 3.2 cm are now in progress. A resonator with mirror diameters of 31.5 cm and an $a^2/b\lambda$ ratio of 0.905 had a measured Q of 260,000. The unloaded Q when end-plate reflection loss is accounted for is 364,000, or a loss of 0.18 per cent. This agrees very well with the curve first given by Goubau and Christian and again by Fox and Li.

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* Received by the IRE, March 3, 1961. The work described in this letter was supported by the Signal Corps under contract DA-36-039-sc-56734.

[1] A. G. Fox and T. Li, "Resonator modes in an optical maser," Proc. IRE, vol. 48, pp. 1904–1905; November, 1960.

[2] E. H. Scheibe, "Study of Surface Wave Transmission lines," University of Wisconsin, Madison, Final Rept. on Signal Corps Contract DA-36-039-sc-56734; June 30, 1955.

[3] B. G. King, E. H. Scheibe, I. Tatsuguchi, "The physical realizability of the Sommerfeld wave on a cylindrical conductor," *Proc. Natl. Electronics Conf.*, vol. 11, pp. 949–957; October, 1955.

[4] B. G. King, I. Tatsuguchi, E. H. Scheibe and G. Goubau, "Pseudo-Resonance Between Parallel Plates," presented at URSI-IRE Spring meeting, Washington, D. C.; May, 1955.

[5] G. Goubau and J. R. Christian, "A New Waveguide for Millimeter Waves," presented at URSI-IRE Fall meeting, San Diego, Calif.; October, 1959.

A Solid-State Analog to a Traveling-Wave Amplifier*

Previous multidiode parametric amplifiers have used relatively few diodes per wavelength on the propagating circuit with the result that the circuit is dispersive and the bandwidth is limited accordingly. In an attempt to overcome this limitation, an exploratory L-band, nondegenerate parametric amplifier has been built with a shielded helix as the propagating circuit, and three parallel stacks of 8 pill varactor diodes, each placed end to end, mounted symmetrically in the helix as the active medium (see Fig. 1).

Gains of 4–8 db with bandwidths of 15–25 per cent were obtained at L band with a pump frequency of 2020 Mc. A noise figure of 9.8 db was measured. The poor noise figure is believed to be due to a poor idle frequency termination and internal reflections of the pump frequency signal. Because the diodes operate under self-biased conditions, they must be carefully selected to ensure the proper operating point for each

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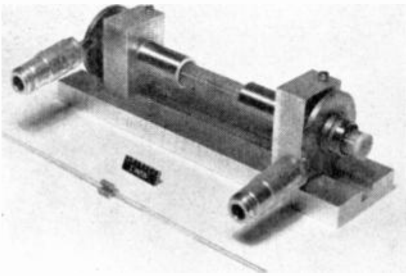

Fig. 1

and must be carefully positioned in the helix to achieve the proper interaction impedance. Further development is required to perfect the amplifier.

It is a pleasure to acknowledge many helpful discussions with Dr. K. Chang and the assistance of R. D. Hughes and E. Diamond.

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The Equivalent Noise Current of Esaki Diodes*

In the theoretical analysis of the noise performance of Esaki semiconductor diodes used as amplifiers and mixers, it is necessary to know the voltage dependence of the *sum* of the two-tunnel current components which constitute the net (measured) tunnel current. Unfortunately, the measured diode current only yields the *difference* of these components. Though it is possible to ascertain the tunnel current sum by shot-noise measurements, in practice, this presents experimental difficulties because of the negative differential conductance exhibited in part of the diode I-V characteristic, and also because of the inherently high differential conductance associated with the positive slope region at small forward and reverse bias voltages. The former situation necessitates stabilization of the diode with a loading conductance whose noise contribution must be known accurately,[1] whereas the latter situation requires that a suitable low-loss impedance step-up transformer be inserted between the diode and the following low-noise preamplifier, in order that the diode noise will not be obscured by the noise of the amplifier.[2]

As an alternative we shall show, using quite general considerations, that each tunnel component, and hence the sum of these components, can be *calculated* simply in terms of the tunnel-diode current for the voltage range, which excludes the "excess" current region.[3] Our results should apply for "reverse" bias voltages and for "forward" bias voltages which encompass most of the negative differential conductance region of the I-V characteristic.

The two oppositely-directed tunnel currents are given by certain integrals over the electronic energy states participating in the tunneling process. These have the general form:[4,5]

$$I_e = A\int_E F(E,V)f(E)[1 - f(E+qV)]dE, \quad (1a)$$

$$I_z = A\int_E F(E,V)f(E+qV)[1-f(E)]dE. \quad (1b)$$

Here E is the energy variable, V the applied voltage, and q the electronic charge. The factor A is a constant which incorporates the physical properties of the semiconductor material. The function F represents the quantum mechanical probability that an electron will tunnel through the diode junction barrier potential. The quantity $f(E)$ is the Fermi occupation function given by

$$f(E) = \frac{1}{1+\exp\left(\frac{E-E_F}{kT}\right)}, \quad (2)$$

where E_F is the Fermi level, k is Boltzmann's constant, and T is the absolute temperature.

The current I_e is caused by tunneling electrons which originate in the conduction band of the n-doped semiconductor, and hence is in the direction of the applied bias when the p-side of the junction is made positive with respect to the n-side ($V>0$).[6] The component I_z is directed oppositely, and originates with electrons which tunnel from the valence band of the p-doped semiconductor. If we define the net bias current I to be positive when the diode is "forward" biased, $V>0$, then

$$I = I_e - I_z. \quad (3)$$

To relate the individual tunnel components to I, we note that the Fermi factors in the tunneling integrals (1) are related as follows:

$$f(E+qV)[1-f(E)]$$
$$= f(E)[1-f(E+qV)]\exp\left(-\frac{qV}{kT}\right). \quad (4)$$

This is the key result. Since V is not an integration variable, then

$$I_z = I_e\exp\left(-\frac{qV}{kT}\right) \quad (5)$$

hence,

$$I_e = \frac{I}{1-\exp\left(-\frac{qV}{kT}\right)}. \quad (6a)$$

$$I_z = \frac{I}{\exp\left(\frac{qV}{kT}\right)-1}. \quad (6b)$$

These results have also been stated by others, though without proof.[7] It should be emphasized that these results are not based upon any assumptions concerning the functional dependence of the tunneling probability factor $F(E, V)$. They are dependent only on the way the Fermi functions appear in the tunneling integrals, and hence, should hold whenever the tunnel currents can be expressed in the form given by (1). Thus, they would apply, for example, to "direct" and "indirect" tunneling in semiconductors[5,3] and to tunneling between normal and superconducting metals separated by an insulating film.[8,9]

The author has used the above relations in a theoretical study of the noise performance of Esaki diode mixers (to be published). Thus, if the two tunnel components exhibit full shot noise, and if the two noise currents are uncorrelated, then the mean square noise current in a frequency interval δf is expressed as

$$\overline{i^2} = 2qI_n\delta f. \quad (7)$$

The equivalent diode noise current I_n is equal to the sum of the two tunnel currents, and hence is given by

$$I_n = I\coth\frac{qV}{2kT}. \quad (8)$$

If the diode I-V characteristic is given in the tunneling region, then the equivalent noise current I_n can be obtained from it with the help of (8). For $T=300°$K, the equivalent noise current and net diode current coincide for $V>100$ mv (excluding the excess current region, of course), a result which has been verified for a germanium diode.[1] It would be very useful to verify (8) by careful noise measurements taken in the positive conductance region near $V=0$, since this region is of importance in mixer applications.

At zero bias, it is important to observe that (7) reduces (correctly) to the thermal noise formula $\overline{i^2}=4kTg\delta f$. In this limit I vanishes; hence,

$$I_n \underset{V\to 0}{\longrightarrow} \frac{\partial I}{\partial V}\frac{2kT}{q} = \frac{2kTg(0)}{q}.$$

Here $g(0)$ is the incremental conductance at $V=0$. This result has also been derived by Tiemann, although in a less direct manner and only for a specific tunneling model.[1]

* Received by the IRE, March 6, 1961.

[1] J. J. Tiemann, "Shot noise in tunnel diode amplifiers," PROC. IRE, vol. 48, pp. 1418–1423; August, 1960.

[2] B. Schneider and M. J. O. Strutt, "Theory and experiments on shot noise in silicon p-n junction diodes and transistors," PROC. IRE, vol. 47, pp. 546–554; April, 1959.

[3] E. O. Kane, "Theory of tunneling," *J. Appl. Phys.*, vol. 32, pp. 83–91; January, 1961.

[4] L. Esaki, "New phenomenon in narrow germanium p-n junctions," *Phys. Rev.*, vol. 109, pp. 603–604; January, 1958.

[5] P. J. Price and J. M. Radcliffe, "Esaki tunneling," *IBM J.*, vol. 3, pp. 364–371; October, 1959.

[6] R. A. Pucel, "Physical principles of the Esaki diode and some of its properties as a circuit element," *Solid-State Electronics*, vol. 1, pp. 22–23; March, 1960.

[7] D. I. Breitzer, "Noise figure of tunnel diode mixer," PROC. IRE, vol. 48, pp. 935–936; May, 1960.

[8] J. Nicol, S. Shapiro, and P. Smith, "Direct measurement of the superconducting energy gap," *Phys. Rev. Lett.*, vol. 5, pp. 461–463; November, 1960.

[9] J. Fisher and I. Giaver, "Tunneling through thin insulating films," *J. Appl. Phys.*, vol. 37, pp. 172–177; February, 1961.

It is interesting to note that (7) and (8) also hold for the low-frequency shot noise in conventional *p-n* junction diodes when operated at low injection levels, and when the effects of recombination and generation in the depletion layer are unimportant.[2,10]

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[10] A. van der Ziel, "Noise in junction transistors," Proc. IRE, vol. 46, pp. 1019–1021; June, 1958.

A Junction-Transistor Frequency Divider*

A low-ratio binary frequency divider may be constructed utilizing the carrier-storage delay in a junction diode.[1] Since such a device possesses two stable states, it may be used to perform logical operations in an analogous manner to that suggested by von Neumann[2] and Goto.[3] Like the parametric subharmonic oscillators, phase synchronization may be obtained by cyclically varying the operating conditions so that the device passes through a region in which it is sensitive to a small subharmonic synchronizing signal. The number of units which may thus be controlled is limited by the load presented to the output terminals of the frequency divider. In order to reduce the effect of the load, some isolation is necessary. One method is to use a junction transistor as the frequency divider.

Typical low-power audio-frequency transistors have storage times which limit the lower input frequency to about 100 kc. Some power transistors may operate with input frequencies as low as 10 kc. For an initial investigation a low operating frequency is to be preferred to enable waveforms to be studied. The following observations were made with an input frequency of 50 kc.

Two typical circuits are shown in Fig. 1. The operating conditions for the input circuit are substantially the same as for the diode. The source impedance should be low, 600 Ω or less. Frequency division may be obtained within the range $1<E/V<2$ and $1<\omega/\omega_0<2$, where E is the peak value and ω the angular frequency of the input voltage, V is the applied bias and ω_0 the series resonant frequency of L and C. It can be shown[4] that the subharmonic output approaches a maximum as E/V and ω/ω_0 tend to 1 and 2, respectively.

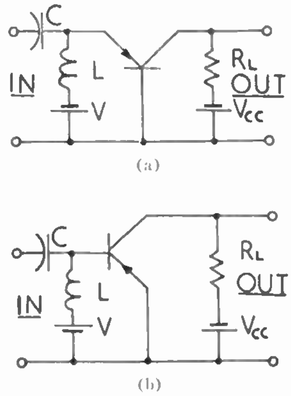

Fig. 1—Transistor frequency divider. (a) Common base. (b) Common emitter.

Fig. 2 shows typical waveforms obtained under the following conditions: $E/V=1.1$, $\omega/\omega_0=1.4$, $V=4.5$ v, $V_{cc}=7.5$ v, $R_L=25$ kΩ. The shape of the collector voltage waveform is influenced principally by the available charge stored in the base; and by the base-collector barrier capacitance in the common emitter circuit. It is found that the common base circuit provides the better isolation, it being possible to reduce R_L to zero without materially influencing the operation of the input circuit. Care should be taken not to exceed the maximum instantaneous rating of the transistor, since a large voltage may be developed in the input circuit especially as ω_0 approaches $\omega/2$. It may prove necessary to operate the transistor in an inverted configuration with the collector-base junction in the input circuit.

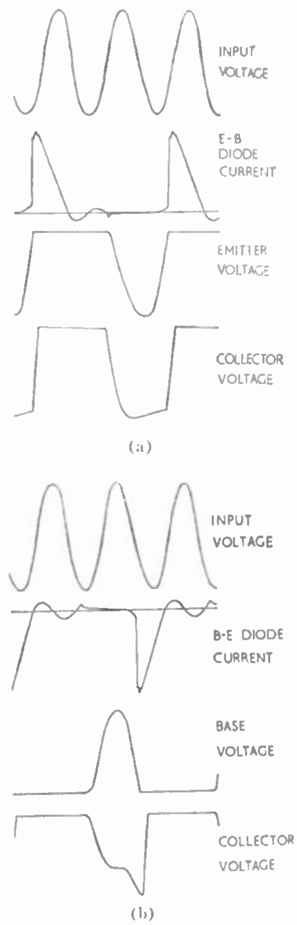

Fig. 2—(a) Common base circuit waveforms. (b) Common emitter circuit waveforms.

If a relatively pure subharmonic output is required, R_L may be replaced by a parallel tank circuit resonant at the subharmonic frequency. Although interaction with the input is increased, stable operation may be obtained.

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* Received by the IRE, January 16, 1961.
[1] W. D. Ryan, "Frequency division by carrier storage," *Electronic Engrg.*, vol. 33, pp. 40–41; January, 1961.
[2] R. L. Wigington, "A new concept in computing," Proc. IRE, vol. 47, pp. 516–523; April, 1959.
[3] E. Goto, "The parametron, a digital computing element which utilizes parametric oscillation," Proc. IRE, vol. 47, pp. 1304–1316; August, 1959.
[4] W. D. Ryan and H. B. Williams, "The carrier-storage frequency divider, a steady-state analysis," in preparation.

Sideband Correlation of Lunar and Echo Satellite Reflection Signals in the 900-Mc Range*

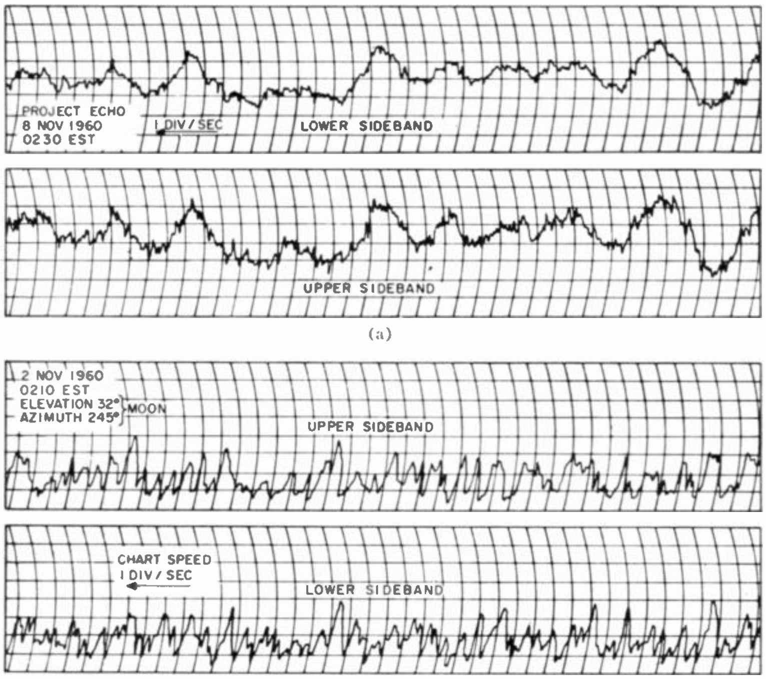
The detection of electromagnetic signals reflected from the moon was reported in 1946 by Mofenson.[1] Fourteen years later voice communication via a moon relay was discussed by Radford.[2] Doppler information and range depth of the moon have been measured,[3] and these have bearing on the effectiveness of a moon relay for communication.

In an attempt to determine the correlation bandwidth of signals reflected from the moon, some preliminary experiments have been carried out recently. These experiments have shown poor correlation in the amplitude fluctuations of two sidebands at 915 Mc. Comparison with similar tests of tropospheric scatter signals and signals reflected from the Echo satellite indicate that the lack of correlation is due to the moon and not to the Earth's atmosphere or ionosphere.

The lunar reflection tests were made cooperatively by General Electric Company and Air Force Cambridge Research Laboratories. AFCRL, at Bedford, Mass., illuminated the moon at 915 Mc using double-sideband suppressed carrier modulation. The transmitted power was ten kilowatts and the transmitting antenna is a 28-foot diameter paraboloid with a linearly-polarized feed.

Signals reflected from the moon were received at General Electric's Radio-Optical Observatory, near Schenectady, N. Y. The receiving antenna is a 28-foot paraboloid with rotatable linearly-polarized feed. A parametric amplifier with a 1.5-db noise

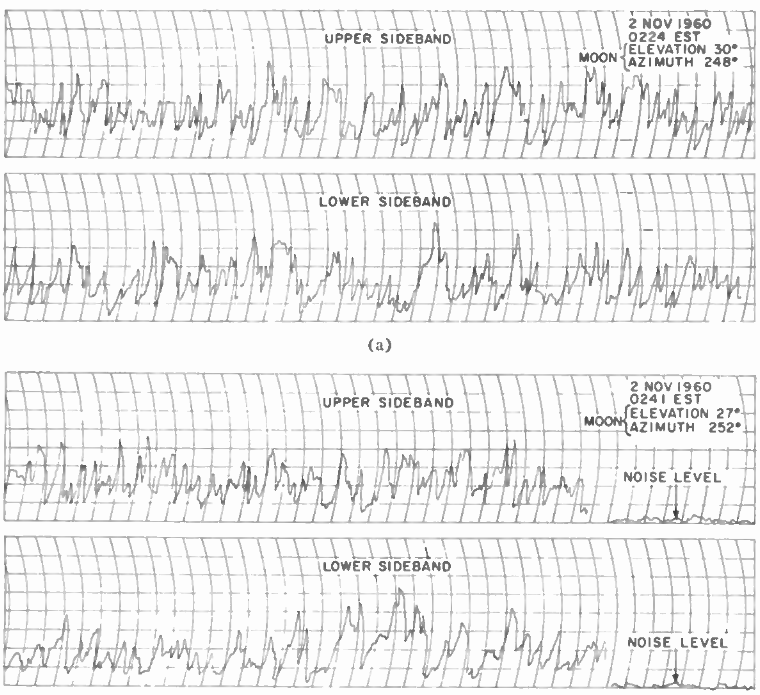
* Received by the IRE, February 16, 1961.
[1] J. Mofenson, "Radar echoes from the moon," *Electronics*, vol. 19, pp. 92–98; April, 1946.
[2] W. H. Radford, "Moon relay communication," 1960 IRE International Convention Record, pt. 5, pp. 277–283.
[3] Gordon H. Pettengill, "Measurements of lunar reflectivity using the millstone radar," Proc. IRE, vol. 48, pp. 933–934; May, 1960.



(a)

(b)

Fig. 1—(a) Echo satellite reflection –20 kc sideband separation. (b) Lunar reflection – 500 cps sideband separation.



(a)

(b)

Fig. 2—(a) Lunar reflection – 2000 cps sideband separation. (b) Lunar reflection – 8000 cps sideband separation.

figure is followed by a crystal mixer and the mixer in turn by a 30-Mc IF amplifier. The amplifier feeds two R390 receivers that have their IF bandwidths set at 100 cps. The AGC circuits of the receivers have been modified so that the AGC voltage is a linear function of signal level in db even at very low signal levels. The AGC time constant is 0.3 second. The receivers are separately tuned, one to each sideband, and the AGC voltages are recorded on a two-pen chart recorder and also on magnetic tape.

Amplitude fluctuation measurements over the tropospheric scatter path from Bedford to Schenectady were made in the same way as the lunar reflection experiments.

Echo experiments were made cooperatively by GE and Bell Telephone Laboratories at Holmdel, N. J. BTL illuminated the satellite at 960.05 Mc using a 60-foot paraboloidal antenna with a circularly-polarized feed. The receiving equipment at General Electric was identical to that used in the lunar reflection tests except that the receiver bandwidths were set at 1000 cps. A carrier phase lock receiving system was used to compensate for Doppler shift. The carrier phase lock equipment is sometimes used on the lunar reflection experiments but is not essential for the moon signals.

Tropospheric scatter recordings were made with sideband separations up to 80 kc. Amplitude fluctuations of both sidebands were always well correlated, even though 30-db fades were common.

Typical results from Echo and the moon are shown in the chart recordings. Chart speed was one division per second for all the examples shown here.

On this particular record, the Project Echo reflections show a fading period averaging approximately 5 seconds. The test shown in Fig. 1(a) was made on November 8, 1960, at 0230 EST. The fine structure is believed to be due to noise. There is reason to believe that this fading is due to distortions in the surface of the satellite. It should be noted that when the measurement was made, the 100-foot-diameter metallized plastic sphere had been in orbit 88 days. The fading rate is much lower than that observed from the moon and has a more regular pattern. The fading of both sidebands was well correlated when the sideband separation was 20 kc.

An examination of the lunar reflection charts show that the amplitude fluctuations as received from the moon appear to have a poorer correlation for a 500-cps sideband separation than from Echo with 20-kc separation and over the tropospheric scatter path with 80-kc separation. The noise level observed on the lunar tests when the receivers were detuned may be observed in Fig. 2(b) on the charts. When both receivers were tuned to the same sideband, *i.e.*, zero-cps sideband separation, the AGC voltage fluctuations appear to be almost perfectly correlated. However, as the sideband separation is increased, the correlation decreases rapidly until it appears to be very low when the sidebands are separated by 4 or 8 kc. Correlation coefficients for lunar data were computed on a digital computer and found to be as follows:

Separation	Correlation
0 cps	0.99
1000	0.325
2000	0.161
4000	0.099
8000	0.111

The author would like to acknowledge the aid of R. W. Garrett and B. H. Claxton of GE, R. J. Miner of AFCRL, and the BTL personnel who carried out most of these measurements.

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The Teflon Waveguide Plunger*

Microwave waveguide plungers composed of a teflon block and a small piece of metal have been used in this laboratory for several years.

The structure of the plunger is shown in Fig. 1. A small piece of metal is embedded, capped, plated or sputtered on the tip of a teflon block, which slides inside the waveguide.

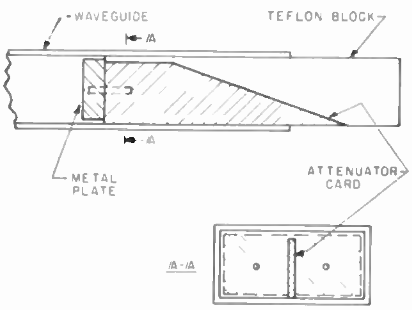

Fig. 1—Teflon plunger for rectangular cross-section waveguide.

The metal part of the plunger is isolated from the waveguide wall by a clearance of a few mils.

The reflection of the wave at the plunger is due only to the large impedance mismatch, but it is large enough to cause a VSWR greater than 35. This is sufficient to match out most load impedances.

The characteristics are constant over the waveguide frequency, because there are no frequency-sensitive components in the plunger.

The leakage wave is absorbed by a microwave attenuator which is embedded in a teflon block, as shown in Fig. 1. The leakage power can be made as small as needed by adjusting the attenuator plate. A leakage of less than -50 db is easily achieved.

In the frequency range above 8 kMc, especially in the millimeter range, the teflon plunger becomes much more stable, and easier to adjust than metal plungers. It shows extremely good stability against mechanical shock and preserves constant electric characteristics for a long time.

Even if the surface of the waveguide is not clean, the electric characteristics as a plunger do not change, because there is no metal-to-metal contact point which is sensitive to surface conditions.

The wax-like smooth surface of teflon makes it very easy to move the plunger in the metal waveguide and to set it in the required position quickly. Therefore, we can very often eliminate mechanical adjusting devices even in the millimeter-wave range. It saves considerable time and labor, both in the experiments and in the production.

Besides this, the elasticity of teflon eases the restrictions on waveguide tolerance and on the surface finish.

Since the plunger fits snugly to the waveguide wall, the waveguide can easily be hermetically sealed without any special consideration.

It is not necessary for the dielectric material to be teflon. Other materials which have the same mechanical characteristics as teflon are equally good or may be even better. A lossy dielectric might eliminate the need for an attenuator behind the metal cap.

The metal part of the plunger can be made in several ways to meet the requirement. Embedding, sputtering, painting, and capping the metal on the tip of the plunger are equally easy to do.

Another version of the teflon plunger in the circular waveguide is shown in Fig. 2.

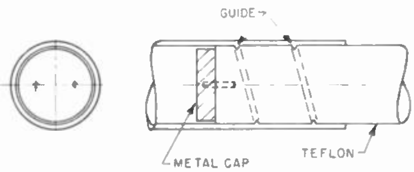

Fig. 2—Teflon plunger for circular cross-section waveguide.

The shallow groove on the waveguide wall guides the teflon plunger smoothly with a screw action.

Carefully made teflon plungers have been found to be as good electrically as metal plungers having spring fingers or chokes on the tips, and mechanically they are very much superior to the metal plungers.

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A New Microwave Power Amplifier*

Quite often the simplest solution to a problem is overlooked. In the case of microwave power amplification a very simple solution, which involves neither resonance nor phase-synchronism and has never been tried, is as follows.

Part of a Lecher line or of a coax is made such that a dc potential can be applied between the two conductors. Also, one of the conductors is made electron emissive. We shall call this electrode the cathode and the other the anode. The dc potential applied is negative and the anode repulses the electrons toward the cathode. When a sufficiently large signal travels down the guide, the positive peaks of the signal will be large enough for electrons affected by it to reach the anode. The current thus produced at the anode separates into a current flowing in phase with the signal and a current flowing in the opposite direction. The current flowing in phase with the signal is increased continuously, while that part of the signal which created it propagates in the tube. The current flowing in the opposite direction is made of about equal increments which flow at equal velocity and can never bunch together. This current is thus a dc current which returns to the source of dc potential between anode and cathode.

Let one affect the system by a pulse of signal of duration τ. The dc potential ϕ_0 is so chosen that the signal potential ϕ_1 surpasses it for a duration θ_T and within that time has an average value $\bar{\phi}_1$. The input energy is of the form

$$E_i = \frac{\bar{\phi}_1^2}{Z}\tau, \tag{1}$$

where Z has the dimensions of an input impedance but also involves the ratio between ϕ_1 and $\bar{\phi}_1$. If the phase velocity of the pulse is c, the length of the pulse in the guide is

$$l = c\tau. \tag{2}$$

The current which flows to the anode within that length is of the form

$$I_p = c\theta_T R f((\bar{\phi}_1 - \phi_0)), \tag{3}$$

with R a length due to the geometry and in general equal to the perimeter of the cross section of the cathode. The function f relates the current density to the potential seen by the electrons. One would, of course, operate the tube in the space-charge-limited region so that the current density would follow the potential difference.

Half of the current reaching the anode flows back to the cathode through the dc source of potential and provides the dc energy necessary. The total dc energy thus provided is

$$E_{dc} = \frac{I_p}{2}\,\frac{L}{c}\,\phi_0, \tag{4}$$

in which L is the length of the active section of the guide. Part of this energy is used for plate dissipation and the rest is transformed into ac energy. The plate dissipation is given by

$$E_p = I_p(\bar{\phi}_1 - \phi_0)\,\frac{L}{c}\cdot \tag{5}$$

Thus the energy acquired by the pulse is

$$E_0 = \frac{I_p}{\epsilon}\,\frac{L}{c}\,(3\phi_0 - 2\bar{\phi}_1). \tag{6}$$

The power gain is the energy acquired over the input energy and, when large, is of the form

$$G = \frac{1}{2}\,I_p Z\left(\frac{L}{l}\right)\frac{(3\phi_0 - 2\bar{\phi}_1)}{\bar{\phi}_1^2}\cdot \tag{7}$$

The efficiency is the ratio of output power to dc power used and is given by

$$\eta = \frac{3\phi_0 - 2\bar{\phi}_1}{\phi_0}\cdot \tag{8}$$

It is simple to plot the current, power gain, and efficiency characteristics on a $(\bar{\phi}_1, \phi_0)$ space.

An efficiency of 60 per cent for a power gain of 50 seems quite reasonable and is obtained, for instance, in the numerical example (Table I). The tube is a limiter either when it becomes temperature limited or when

* Received by the IRE, February 14, 1961.

$\bar{\phi}_1$ exceeds $3/2\phi_0$, whichever occurs for the lowest $\bar{\phi}_1$. The tube of course is not directional and will oscillate when reflections produce a reflected $\bar{\phi}_1$ larger than ϕ_0. A directional tube can be made by lightly bending the paths of the electrons forward or backward with a dc magnetic field. A small signal theory for such a tube, with ϕ_0 necessarily positive, shows gaining waves.

TABLE I

L	$=10$ cm
I_p/l	$=200$ a/cm
Z	$=600\ \Omega$
τ	$=10^{-2}$ second
ϕ_1	$=6000$ volts
ϕ_0	$=5000$ volts
E_0	$=3\cdot10^4$ joules (output power: $3\cdot10^6$ watts)
E_p	$=2\cdot10^4$ joules ($2\cdot10^3$ joules/cm of axial length)
G	$=50$
η	$=60$ per cent

Basically, the tube is a class-C amplifier at microwave frequencies so that the output current wave form corresponding to a sinusoidal voltage contains a large amount of harmonics. It is obviously possible to keep the energy in the fundamental by having a sufficiently-resonating load or antenna. In the case where the load or antenna is extremely broad band, the useful output-signal is made of pulses repeating at a microwave frequency, a rather interesting possibility.

This new amplifier is related to the oscillator invented by A. G. Clavier and used by him in the late twenties on the microwave link across the English Channel. In this case the anode circuit was ended so as to resonate and the dc potential ϕ_0 did not exist. The noise power in the electron gas surrounding the cathode was sufficient to excite the resonating anode.

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A Superposition Property of Angle Modulation*

There has been some previous analysis reported concerning the properties of the waveform

$$V(t) = \cos\left[\omega_c t + P(t) + f(t)\right], \tag{1}$$

where ω_c is a fixed angular frequency, $P(t)$ is a periodic angle modulation of period T, and $f(t)$ is a second angle modulation which may be either random or causal. Blachman[1] and Karr[2] have shown with certain approximations that the effect of the second modulation is to angle modulate each of the discrete spectral components of the waveform

$$y(t) = \cos\left[\omega_c t + P(t)\right] \tag{2}$$

by the modulation $f(t)$. Thus, given

$$y(t) = \sum_n d_n \cos\left[\left(\omega_c + \frac{2\pi n}{T}\right)t - \beta_n\right], \tag{3}$$

it has been shown to be approximately true that

$$V(t) = \sum_n d_n \cos\left[\left(\omega_c + \frac{2\pi n}{T}\right)t + \beta_n + f(t)\right]. \tag{4}$$

The purpose of this correspondence is to show that (4) is indeed an exact relationship.

Begin by rewriting (1) as

$$\begin{aligned} V(t) &= \mathrm{Re}\left\{\epsilon^{j[\omega_c t + P(t) + f(t)]}\right\} \\ &= \mathrm{Re}\left\{\epsilon^{j[\omega_c t + f(t)]}\epsilon^{jP(t)}\right\}. \end{aligned} \tag{5}$$

The periodic function $\epsilon^{jP(t)}$ may now be expanded in a complex Fourier series,

$$\epsilon^{jP(t)} = \sum_n C_n \epsilon^{j(2\pi n/T)t} \tag{6}$$

where the C_n are given by

$$C_n = \frac{1}{T}\int_0^T \epsilon^{j[P(t)-(2\pi n/T)t]}dt. \tag{7}$$

Substitution of the series in (5) and term-by-term multiplicaiton by the first exponential gives

$$V(t) = \mathrm{Re}\left\{\sum_n C_n \epsilon^{j[(\omega_c + 2\pi n/T)t + f(t)]}\right\}. \tag{8}$$

Taking the real part as indicated, we have

$$\begin{aligned} V(t) = \sum_n R_n \cos&\left[\left(\omega_c + \frac{2\pi n}{T}\right)t + f(t)\right] \\ - I_n \sin&\left[\left(\omega_c + \frac{2\pi n}{T}\right)t + f(t)\right], \end{aligned} \tag{9}$$

where

$$\begin{aligned} R_n &\triangleq \mathrm{Re}\{C_n\} \\ I_n &\triangleq \mathrm{im}\{C_n\}. \end{aligned} \tag{10}$$

Use of the trigonometric identity

$$\begin{aligned} &A\cos\theta + B\sin\theta \\ &= \sqrt{A^2+B^2}\cos\left[\theta + \tan^{-1}\left(\frac{-B}{A}\right)\right] \end{aligned} \tag{11}$$

gives

$$\begin{aligned} V(t) = \sum_n \sqrt{R_n^2 + I_n^2}\cos&\left[\left(\omega_c + \frac{2\pi n}{T}\right)t \right. \\ &\left. + \tan^{-1}\left(\frac{I_n}{R_n}\right) + f(t)\right]. \end{aligned} \tag{12}$$

The definitions

$$\begin{aligned} d_n &= \sqrt{R_n^2 + I_n^2} \\ \beta_n &= \tan^{-1}\left(\frac{I_n}{R_n}\right) \end{aligned} \tag{13}$$

lead to the desired result,

$$V(t) = \sum_n d_n \cos\left[\left(\omega_c + \frac{2\pi n}{T}\right)t + \beta_n + f(t)\right]. \tag{4}$$

Note that when $f(t)$ is identically zero, (4) reduces to (3), thereby ensuring that the d_n and β_n are the same in both equations.

Thus, the effect of the periodic modulation $P(t)$ is to divide the carrier into a number of subcarriers separated in frequency by an amount equal to the fundamental frequency of $P(t)$. The added modulation $f(t)$ then angle modulates each of these subcarriers identically.

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[1] N. M. Blachman, "Limiting Frequency Modulation Spectra," Electronic Defense Lab., Sylvania Electric Products Inc., Mountain View, Calif., Tech. Rept. No. EDL-M43, pp. 3–5; April, 1955.

[2] P. R. Karr, "A note on angle modulation by a mixture of a periodic function and noise," IRE Trans. on Information Theory, vol. IT-5, pp. 140–143; September, 1959.

Cerenkov Radiation in a Dielectric Tube Waveguide*

The Cerenkov radiation of a prebunched beam in a dielectric tube waveguide has two important features: one, the mode interference problem and physical size restrictions are less severe than in closed metallic waveguides;[1] two, this mode of interaction provides a radiation resistance which is about an order of magnitude higher than that of the conventional Cerenkov radiation in an infinite medium.[2,3]

The experiments reported on herein consist of passing a tightly bunched, 0.8-mev beam along the axis of a dielectric tube waveguide of inner and outer radii a and b, respectively, and a relative dielectric constant κ. The usual synchronism conditions are imposed, *i.e.*, at the beam harmonic frequency of interest the phase velocity of the TM_{0m} waveguide mode that is excited is equal to the beam velocity. A schematic drawing of the experimental arrangement is shown in Fig. 1.

Cerenkov radiation, escaping from the end of the dielectric tube, passes through a

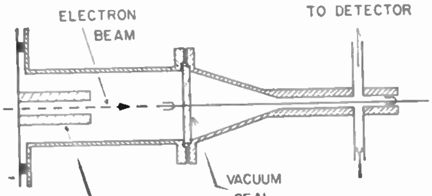

Fig. 1—Schematic diagram for the Cerenkov radiation of a 0.8-mev tightly bunched beam in dielectric tube waveguide. The beam source is not shown.

* Received by the IRE, February 18, 1961. This work was sponsored by the U. S. Atomic Energy Commission under Contract AT(11-1)-392.

[1] R. C. Becker and P. D. Coleman, "The dielectric tube resonator," *Proc. Symp. Millimeter Waves*, Polytechnic Press, Brooklyn, N. Y., pp. 191–222; April, 1959.

[2] H. Lashinsky, "Generation of microwaves by Cerenkov radiation," *Proc. Symp. Millimeter Waves*, Polytechnic Press, Brooklyn, N. Y., pp. 181–190; April, 1959.

[3] P. D. Coleman and C. E. Enderby, "Megavolt electronics Cerenkov coupler for the production of millimeter and submillimeter waves," *J. Appl. Phys.*, vol. 31, no. 9, pp. 1695–1696; September, 1960.

glass vacuum seal into a coaxial line coupled to a waveguide detection system tuned to a wavelength of 9.01 millimeters. This arrangement does not completely couple out all of the power generated, but it is sufficiently good to evaluate adequately the behavior of the device.

First to be examined were the phenomena of first and second coherence. When the length L of the interaction region is greater than the spacing D between two consecutive electron bunches, the radiated power P_n or total radiation resistance R_n vary as the square of the distance L, which phenomenon is called second coherence.[4] If the length L is less than D, then R_i is a linear function of L, resulting in first coherence, the condition that applied to these experiments.

When the nth harmonic beam frequency is in synchronism with the TM_{0m} mode of the dielectric guide, then the Cerenkov power generated by first coherence can be expressed in the form

$$P_n = \frac{I_n^2}{2}\left[\frac{(1-\beta^2)DL}{4\pi\epsilon_0 c\lambda_n^2} G_m(a, b, \lambda_n, \kappa)\right] = \frac{1}{2} I_n^2 R_n, \quad (1)$$

where the transverse dimensions of the beam are assumed to be small, $v=\beta c$ is the beam velocity, I_n is the harmonic current amplitude, λ_n is the harmonic wavelength, and G_m is a function of tube geometry, dielectric material, and mode of operation.

Fig. 2 gives the experimental results for radiation resistance vs length of interaction L. Within experimental accuracy, the radiation resistance was found to increase linearly with length; the mode of operation was the TM_{01} mode. It was noticed that for large waveguide lengths there was some evidence of beam interception by the sidewalls. In spite of the fact that this can reduce the power level, it is doubtful that enough beam interception would occur to change a square law variation to the observed linear variation.

In Fig. 3 is shown the theoretical and experimental variation of radiation resistance for a teflon tube waveguide, operating in a TM_{01} mode, as the inner radius, and hence also the outer radius, are varied. The theoretical curve is for the bunch-by-bunch type of radiation, since the length of the structure is less than the distance of separation between two consecutive bunches. It is seen that agreement between the theoretical and experimental curves improves considerably as the inner and outer radii of the waveguide are increased. This is due to three factors: In the first place, increasing the radii improves the matching of electromagnetic radiation between the waveguide and free space. In the second place, an increase of inner radius reduces the extent of beam interception by the side walls and hence the debunching of the beam. Finally and most important, at small hole sizes the inner surface of the waveguide seemed to charge-up; the surface charge tends to

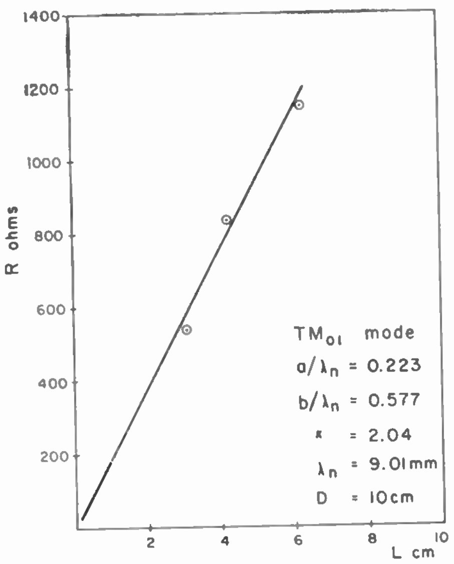

Fig. 2—Experimental variation of the radiation resistance vs length of a dielectric tube waveguide.

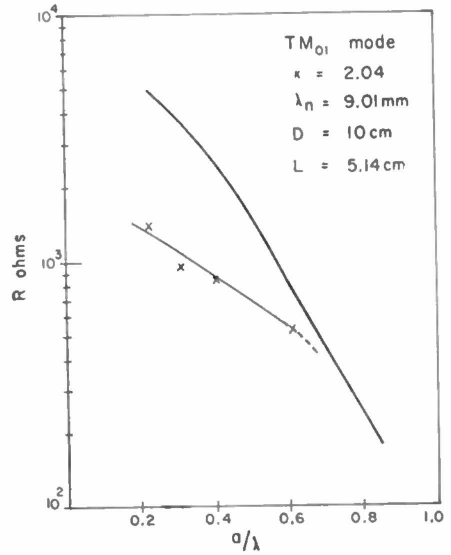

Fig. 3—Experimental and theoretical variation of radiation resistance vs a/λ for a dielectric tube waveguide. (A different glass seal from the one in the results of Fig. 2 was used.)

shield the beam and hence to reduce the power level appreciably. This last effect seemed to be quite independent of the length of the structure.

As has been shown in the literature,[5] Cerenkov radiation in a slow-wave guiding structure and TWT action are the same. Hence, the relatively high radiation resistance of the dielectric tube waveguide may lead one to suspect that this structure may be suited for TWT schemes. However, at low phase velocities the dielectric constant of the waveguide must correspondingly be high, which results in a reduction of the circuit impedance, unless the dimensions are reduced considerably.

On the other hand, this scheme has a higher radiation resistance than Cerenkov radiation in an infinite dielectric material. The price of the high radiation resistance is the mode interference problem which is nonexistant in an infinite slab of dielectric material.

These experiments again verify that higher interaction resistances can be obtained at the expense of physical size and mode interference. However, it appears that the Cerenkov effect in a dielectric tube guide can readily be used down to a wavelength of 1 millimeter with power levels of the order of 1 to 10 watts for a drive current of 0.1 ampere.

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[4] G. A. Bernashevsky, *et al.*, "Radiation in the millimeter waveband by a relativistic electron flow," *Proc. Symp. Millimeter Waves*, Polytechnic Press, Brooklyn, N. Y., pp. 169–180; April, 1959.

[5] J. R. Pierce, "Interaction of moving charges with wave circuits," *J. Appl. Phys.*, vol. 26, no. 5, pp. 627–638; May, 1955.

Autocorrelation Pattern Recognition*

The recent article on autocorrelation pattern recognition[1] was of considerable interest, since this author did some work in the same area several years ago.[2] The optical methods discussed in the article imply, in some sense, the storage of large amounts of information, *viz.*, the value of the autocorrelation function for all possible translations of the ideal pattern relative to itself. In contrast, the previous work was an attempt to distill this vast amount of information into a smaller quantity which could be stored easily. This study was entirely in the framework of computer program recognition, where economy of storage space is of some importance. One advantage of the scheme to be described is that it permits (indeed requires) the use of several "ideals" for definition of the patterns.

Consider the presentation of several samples of a given pattern which are to be used as the "ideals" for recognition. Rather than use the complete autocorrelation functions, we could examine the set of autocorrelation functions for the various patterns presented and try to find those displacements (amounts of two-dimensional shift) for which the several autocorrelation functions have the most "similar" values.

For the moment, let us consider only one displacement. For this displacement, we can examine the distribution of the values of the autocorrelations of the samples which are presented. We wish to find some measure of "similarity" from the distribution of values at the one displacement.

Several measures of "similarity" of a set of numbers, measured from their distribution, have been proposed. One measure is the

* Received by the IRE, February 14, 1961.

[1] L. P. Horwitz and G. L. Shelton, Jr., "Pattern recognition using autocorrelation," Proc. IRE, vol. 49, pp. 175–185; January, 1961.

[2] R. Y. Kain, "A New Approach to Pattern Recognition," M.S. thesis, Dept. of Elec. Engrg., Mass. Inst. Tech., Cambridge; June, 1959.

standard deviation of the numbers, which indicates the "spread" of the numbers from their mean value. If the numbers are all multiplied by a constant, their standard deviation (and mean) will be multiplied by that same constant. So that the similarity measure will be invariant to scale changes (in the numbers), the ratio of the mean to the standard deviation is chosen as the measure of "similarity." A second measure has been proposed by Lewis,[3] and involves computation of the amount of information "contained" in a distribution of values. This requires a better estimate of the distribution than can be obtained from several samples, and also requires computation of logarithms. The first method, using the ratio of the mean to the standard deviation, was used exclusively in these experiments.

Recognition using the mean to standard deviation ratio was programmed for the TX-O computer. Samples of patterns were traced from a book of type faces. The height of a line (and not of a letter) was standardized in the tracing process. For recognition of printed characters, this is not an unreasonable constraint, since it merely corresponds to adjustment of the magnification of the optical system on the input. Seven different type fonts were used for the experiments, but only capital letters were used. In all cases, all seven examples of each letter were presented as ideals, and were later presented for recognition. Numerals were not used in any phase of the experiments. A 16×16 square matrix was used, and the line height was standardized at ten squares.

The seven samples of a given letter were used to obtain seven samples of the autocorrelation function for a particular displacement. The ratio of the mean to the standard deviation of these values was then computed. The N (number chosen) displacements with the highest ratios were selected as the "characteristics" of that letter. Presentation of samples of another letter will result in another selection, giving N displacements for this letter. In general, there is some overlap in the selection.

For the moment, consider recognition of only one letter. When a picture is presented for recognition, the value of the autocorrelation for each of the previously selected displacements is computed. If the value lies within R (the range) standard deviations of the mean for each of T (the threshold) displacements, the pattern is said to match the samples previously used to define the characteristics. When a number of different letters have been given as samples, this procedure is followed for each possible answer. Thus, a number of possible answers are given, not the "most likely" answer. In almost all cases, the correct answers were given, but there were multiple answers in most cases.

A considerable improvement in the results was obtained by using the autocorrelation function normalized to the number of black points in the picture. The number of correct answers was increased, while the number of incorrect answers was decreased. It is interesting that a large percentage of the incorrect answers were either I or J. Reduction of the threshold to pick up all correct answers can be accomplished either by reducing the threshold uniformly, or by reducing it only for those letters where correct answers were missed. The experimental results are summarized in Table I, where equal letter probabilities have been assumed. The columns labeled "Without I, J" indicate the results if all I or J answers are disregarded, but I and J are still allowed as inputs. A threshold of 23/24 indicates reduction of the threshold to twenty-three for those letters (A, F, H, J, O, U) where a correct answer was missed. The last column in the table indicates the results using a simple system for picking one answer from the set of possibilities. The probability that a given answer will be correct is found by feeding all possible letters to the computer and assuming that all letters are equally probable (for simplicity). When a number of answers are given, the one with the highest probability of being correct is chosen.

TABLE I

EXPERIMENTAL RESULTS

	Method			Percent of Correct Answers Appearing		Average Number of Answers per Sample		Per cent Recognition
	N	T	R	With I, J	Without I, J	With I, J	Without I, J	With I, J
Unnormalized	16	16	4.5	87.0	79.9	3.14	2.40	—
	16	15	4.5	95.6	88.3	4.73	3.82	—
Normalized	24	24	3.5	98.3	90.0	1.98	1.34	89.9
	24	23	3.5	100.0	92.3	2.75	1.94	84.6
	24	23/34	3.5	100.0	92.3	2.16	1.46	93.4

There is one major problem inherent in this method. Identical values for all of the samples at a particular displacement give a standard deviation of zero. As this is not a good measure of the deviation, such displacements were not used for recognition. To estimate the number of such cases, the division subroutine was modified so that $a/0$ was a large positive number when a was nonzero, and zero when a was zero. When this was done, all selected displacements had deviations of zero, showing that a large number of "good" displacements are discarded by the procedure used. One solution might assign some arbitrary small deviation to these cases.

As indicated above, this method requires less storage space than the use of the complete autocorrelation functions. It possesses the further advantage that it is possible to specify the pattern by a number of samples, rather than by only one sample. This situation seems to correspond more to the specification of patterns to people. The method here also involves little computation, a complete recognition including selection among all possibilities in the alphabet of twenty-six letters requiring only about one second on the TX-O. Since the computer does not have instructions for multiple shifting, multiplication, or division, the time should be much less in most high-speed computers. Selection of displacements from seven samples requires about three minutes per pattern, but this is done infrequently. It should be possible to modify the method of selection to include the possibility of presentation of samples of "not A," etc. The values of the autocorrelation function of these "negative samples" should be used in the selection process, credit being given to how well separated they are from the values for the "positive samples." There are several ways that this might be done. It might also be possible to employ similar processes in finding complete autocorrelation functions from multiple samples using some sort of averaging and weighting of the various points, according to the amount of "similarity" of the autocorrelations of the samples at the particular displacement. Such a weighting could be incorporated in the optical scheme discussed by Horwitz and Shelton using an extra negative immediately adjacent to the standard autocorrelation negative.

The author is indebted to his thesis advisor, Prof. P. M. Lewis, for many helpful comments and suggestions.

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[3] P. M. Lewis, "Efficient Information Storage," General Electric Res. Lab., Schenectady, N. Y., Rept. No. 58-RL-2141; December, 1958.

Closed-Form Solution for the Output of a Finite-Bandwidth Pulse-Compression Filter*

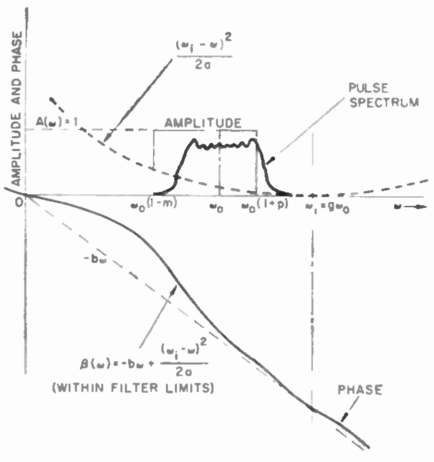
The analysis in a review paper by Cook[1] concerning a pulse compression technique which utilizes linear FM phase coded pulses, assumes a matched filter approximation of infinite bandwidth and linear group time delay. This model can be refined considerably by a restriction of filter bandwidth. A closed-form solution has been obtained, and the theoretical output pulse computed more closely resembles the output of actual bandwidth-limited, pulse-compression filters for low and moderate time-bandwidth products.

The coded pulse signal at the input to the filter can be represented by

$$f(t) = \text{cosine } (\omega_0 t + \tfrac{1}{2}at^2) \quad \text{for } |t| < T \qquad (1)$$

* Received by the IRE, February 3, 1961.

[1] C. E. Cook, "Pulse compression—key to more efficient radar transmission," PROC. IRE, vol. 48, pp. 310–316; March, 1960.



(a) $H(j\omega)=A(\omega)e^{j\beta(\omega)}$

(b) $A(\omega)=A(-\omega)$

(c) $\beta(-\omega)=-\beta(\omega)=b\omega-\dfrac{(\omega_i-\omega)^2}{2a}$

Fig. 1—Filter transfer characteristic.

and zero elsewhere. The term $a=\pi\Delta f/T$ is the sweep rate, where Δf is the total frequency deviation. The Fourier transform of $f(t)$ can be written as

$$F(\omega)=\int_{-T}^{T}\left(\exp\left[j(\omega_0-\omega)\tau+j\frac{a\tau^2}{2}\right]+\exp\left[-j(\omega_0+\omega)-j\frac{a\tau^2}{2}\right]\right)d\tau. \quad (2)$$

The second term in this integral can be shown to be very small and will be neglected. For the bandwidth-limited filter characteristic shown in Fig. 1, the output of the filter is given by

$$G(t)=\frac{1}{2\pi}\,\mathrm{Re}\left\{\exp\left[j\omega_i(t-b)-j\frac{a(t-b)^2}{2}\right]\int_{-T}^{T}[\exp(ja\delta\tau)]\times\left[\int_{\omega_0(1-m)}^{\omega_0(1+p)}\exp\left(j\frac{[\omega-\omega_0+a(\delta-\tau)]^2}{2a}\right)d\omega\right]d\tau\right\} \quad (3)$$

where

$$\delta=t-b-(g-1)\omega_0/a.$$

The double integral can be evaluated in closed form by integrating by parts to give the exact solution of (3):

$$\begin{aligned}G(\delta)=\frac{\sqrt{T\Delta f}}{2\pi\Delta f\delta}\Bigg(&\sqrt{S^2+C^2}\cos\left[(\omega_0+\pi\Delta f)\delta-\frac{\pi\Delta f}{2T}\delta^2+\phi-\tan^{-1}\frac{C}{S}\right]\\&-\sqrt{(S')^2+(C')^2}\cos\left[(\omega_0-\pi\Delta f)\delta-\frac{\pi\Delta f}{2T}\delta^2+\phi-\tan^{-1}\frac{C'}{S'}\right]\\&+\sqrt{(S'')^2+(C'')^2}\cos\left[\omega_0(1+p)\delta+\phi-\tan^{-1}\frac{C''}{S''}\right]\\&-\sqrt{(S''')^2+(C''')^2}\cos\left[\omega_0(1-m)\delta+\phi-\tan^{-1}\frac{C'''}{S'''}\right]\Bigg)\end{aligned} \quad (4)$$

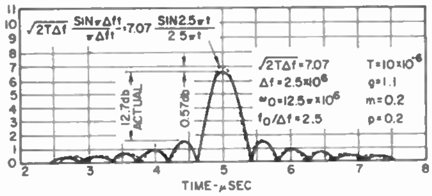


Fig. 2—Typical example of output pulse for bandwidth-limited pulse compression filter.

where ϕ is a constant phase shift which is contained in each term and can be ignored; and where $S, S', S'', S''', C, C', C'', C'''$ are, respectively, the sine and cosine Fresnel integrals of the following limits:

$$S=\int_{\alpha_1}^{\alpha_2}\sin\left(\frac{\pi}{2}u^2\right)du,$$

$$\alpha_1=\sqrt{T\Delta f}\left[\frac{\delta}{T}-1-\frac{2mf_0}{\Delta f}\right],$$

$$\alpha_2=\sqrt{T\Delta f}\left[\frac{\delta}{T}-1+\frac{2pf_0}{\Delta f}\right].$$

S' is the sine Fresnel integral with limits β_2 and β_1 where β_1 and β_2 are given, respectively, by adding $\sqrt{2T\Delta f}$ to α_1 and α_2;

$$S''=\int_{\gamma_1}^{\gamma_2}\sin\left(\frac{\pi}{2}u^2\right)du,$$

$$\gamma_1=\sqrt{T\Delta f}\left[-1-\frac{2pf_0}{\Delta f}\right],$$

$$\gamma_2=\gamma_1+2\sqrt{T\Delta f}.$$

Likewise, S''' is given by the same expression as S'' with $(+m)$ replacing $(-p)$ in the limits.

The result given in (4) can be used to compute the output of a pulse-compression filter for any pulse-compression ratio, or time-bandwidth product $(2T\Delta f)$, when the filter band-pass characteristic is centered on, or offset, with respect to the signal spectrum; thus, the effects of Doppler shifts on the output can be evaluated. The output envelope for a specific case, with moderate pulse-compression ratio of 50:1, is plotted in Fig. 2 and compared with the sine x/x form. It may be noted that the peak is reduced somewhat from the sin x/x envelope. In addition, the zeros of the sidelobe structure are displaced unequally. The sidelobe structure will complicate the application of techniques for sidelobe reduction such as Taylor weighting[2] and transveral filtering.

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[2] T. T. Taylor, "Design of Line Sources for Narrow Bandwidth and Low Side Lobes," Hughes Aircraft Co., Culver City, Calif., Tech. Memo No. 316; July, 1953.

Silicon Epitaxial Microcircuit*

A completely semiconductor microcircuit was devised from a slab of multilayered silicon which was constructed by the Merck, Sharp and Dohme Research Laboratory. The completely silicon structure was entirely fabricated by the use of vapor phase deposition techniques. An eight-layered epitaxial structure was constructed to yield electrical characteristics of an oscillatory nature. Simple processing of the multilayered slab directly produced the microcircuit structure.

The three-terminal network oscillated in different modes yielding output voltage variations as shown in Fig. 1. Sinusoidal and gated type electrical variations occurred. Voltage amplitude variations of 1 to 5 volts occurred at pulse and cyclical time rates of 10,000 to 30,000 repetitions per second. By the use of an additional linear external resistance, it was possible to observe a waveform consisting of a 100-kc variation combined with a 10-kc variation.

The silicon integrated circuit was designed to contain regions capable of performing switching and negative resistance action, capacitive storage, and resistive dissipation. The basic circuit arrangement is that of the conventional saw-tooth generator-relaxation oscillator. Shunting capacitive and resistive leakage effects play a role in the microcircuit operation. Electrical characteristics of the negative resistance element or switch were: holding-current magnitude was 1.25 milliamperes, breakover voltage magnitude 9.0 volts, and negative resistance value 5000 ohms. The voltage-current characteristics of the capacitive region was typical of that of a large-area silicon p-n junction. The measured value of zero bias capacitance was 1500 $\mu\mu f$. A conductive material applied properly to a p-n junction of the C-region circumvented the occurrence of low-voltage breakover. Similarly, conductive material applied properly

* Received by the IRE, January 16, 1961; revised manuscript received, February 13, 1961.

to a junction of the negative-resistance region permitted control of transition parameters and enabled low-voltage breakover to occur. The resistive region possessed a linear type electrical characteristic. Resistance magnitude was about 40,000 ohms. The structure resistive coupling between the output terminal and the resistance input terminal was very low.

Various features of circuit operation were determined largely by the initial distribution of impurities in the multilayered structures. Maximum applied voltage, for example, was limited to a value below the breakdown voltage value of the junction that isolated the resistance region. The applied voltage magnitude was also limited by the breakdown voltage value of the center junction of the capacitive region.

Circuit operation depended largely upon surface conditions of all junctions. The shunt leakage effect of the capacitive region was significant in affecting circuit operation. Excessive shunt capacitive leakage introduced a voltage-dividing action that resulted in the occurrence of a stable dc level. Relatively large capacitive leakage did not permit oscillations to occur. The off impedance of the switching element (R_{4L}), the charging resistance (R), and the capacitive shunt resistance (R_{CSH}) combined and imposed the following necessary condition upon circuit operation

$$V_{\text{applied}} \times \frac{\dfrac{R_{4L} \times R_{CSH}}{R_{4L} + R_{CSH}}}{R + \dfrac{R_{4L} \times R_{CSH}}{R_{4L} + R_{CSH}}} \geq V_{4LBO}$$

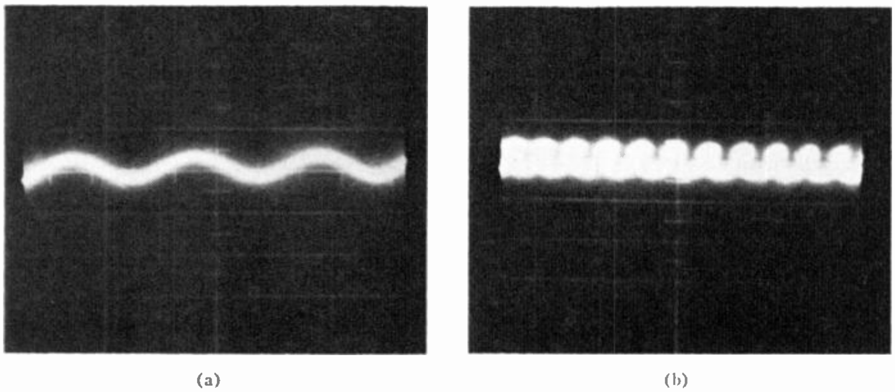

Fig. 1—Output waveforms: (a) sinusoidal, horizontal (30 μsec/cm), vertical (5 volts/cm); (b) gated, horizontal (30 μsec/cm), vertical (1 volt/cm).

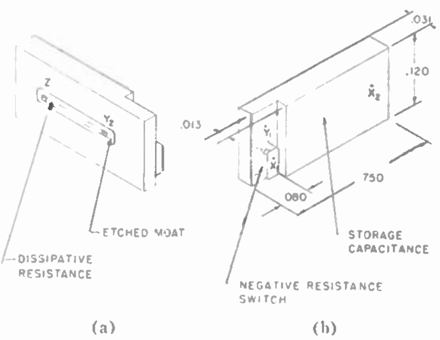

Fig. 2—Sketch illustrating slab external geometry. (a) Resistive region. (b) Capacitive and switching-negative resistance regions.

where V_{applied} is applied voltage, and V_{4LBO} is the breakover voltage.

The geometry of the multilayered microcircuit is illustrated in Fig. 2. Fig. 3 shows the individual layer thicknesses and indicates type conductivity. In constructing the circuit from the multilayer silicon sandwich, several simple asphalt mask applications were used as well as several chemical etch applications consisting of relatively fast and slow silicon etches.

Preliminary work indicates that a new process tool is available for broadening and simplifying the approaches to microcircuit construction. Results suggest the possible use of layerized silicon sandwiches that can be chemically or electrochemically etched a number of ways to obtain numerous forms of electrical performance.

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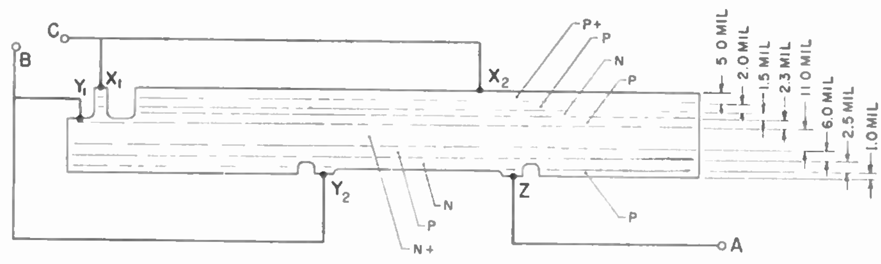

Fig. 3—Sketch showing features of epitaxial slab.

On the Integration and Summation of Trigonometric and Exponential Functions*

For many purposes, the expressions given below for the definite integrals of trigonometric and exponential functions are more convenient than the well-known forms. Similar expressions for the summations of such functions are also useful and are shown below. Each formula applies to three cases, and the same function must be used on both sides of the equation; exp i means the natural exponential of i times the stated argument. The proofs are left as exercises for the reader.

$$\int_a^b \begin{matrix}\sin \\ \cos \\ \exp i\end{matrix} (cx + d)dx = \frac{\sin c\,\dfrac{b-a}{2}}{\dfrac{c}{2}} \begin{matrix}\sin \\ \cos \\ \exp i\end{matrix} \left(c\,\frac{b+a}{2} + d\right),$$

$$\int_a^b \begin{matrix}\sinh \\ \cosh \\ \exp\end{matrix} (cx + d)dx = \frac{\sinh c\,\dfrac{b-a}{2}}{\dfrac{c}{2}} \begin{matrix}\sinh \\ \cosh \\ \exp\end{matrix} \left(c\,\frac{b+a}{2} + d\right),$$

$$\sum_{k=a}^{b} \begin{matrix}\sin \\ \cos \\ \exp i\end{matrix} (kc + d) = \frac{\sin \dfrac{b-a+1}{2}c}{\sin \dfrac{c}{2}} \begin{matrix}\sin \\ \cos \\ \exp i\end{matrix} \left(\frac{b+a}{2}c + d\right),$$

$$\sum_{k=a}^{b} \begin{matrix}\sinh \\ \cosh \\ \exp\end{matrix} (kc + d) = \frac{\sinh \dfrac{b-a+1}{2}c}{\sinh \dfrac{c}{2}} \begin{matrix}\sinh \\ \cosh \\ \exp\end{matrix} \left(\frac{b+a}{2}c + d\right).$$

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The Old and the New*

Poles and Zeros' "The Old and the New," in the PROCEEDINGS for February, 1961, recalls the old crystal detector, now resurrected in combination with modern advances in the art for uses where the latter alone are unworkable.

But it fails to point out the unmistakable and striking similarity between those detectors and modern diode transistors, not alone in physical form, but in electrical characteristics as well.

One need only go back to the early (1910) papers of Harvard's George Pierce, G. W. Picard, or J. A. Fleming, to see such devices pictured and described in detail, with characteristic *E-I* curves, very much like those of present-day transistors, using the same point contacts but on natural instead of man-made crystals. I, myself, also made such measurements and, additionally, on electrolytic detectors in those early days, and still have the curves. Modern solid diode rectifier devices are only refinements of those old crystal detectors, so why all this hocus pocus in new nomenclature? Some of the "wireless" workers of that period even made those detectors *oscillate!*

May I also remind our forgetful oldsters and untutored newsters in this art that when Whitney, Langmuir and their associates and contemporaries at Schenectady and New York's West Street got higher vacuums in de Forest's Audion, by means of a newly-developed German pump, they also used such hocus-pocus nomenclature, by the word "Kenotron," in an effort to label a mere improvement as a wholly new genus of electronic devices. The Patent Courts, however, judged otherwise!

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* Received by the IRE, February 20, 1961.

Generating a Rotating Polarization*

In a recent letter[1] Allen discussed the problem of generating an EM wave with rotating polarization. The solution involves the use of two radio sources, of different frequency, with an arrangement of phase shifters between the sources and a radiating device. The two resulting EM waves from the radiator are circularly polarized and of such phases as to produce a rotating polarization. The present note describes a system developed to produce a rotating-polarization, 22-Mc wave for transmission from the ground to a rocket.

The application involved the determination of the electron density of the ionosphere from the Faraday rotation of the signal. A nominally-zero spin rate was anticipated for the rocket. By initially rotating the plane of polarization of the transmitted signal, a series of nulls was received on a linear dipole carried by the rocket. From a comparison of the times of occurrence of these nulls with those observed on a ground-based monitor the electron density could be computed.

The method is illustrated by Fig. 1. It was based on the well-known Magic-*T* or hybrid junction. This was constructed of coaxial cable (RG 11/*U*) with $L_1=\lambda g/4$ and $L_2=3\lambda g/4$. λg is the wavelength in the cable of the *T*. Two of the terminals were excited through RG 8/*U* cables, by signals of frequency f_1 and f_2 which differed by 160 cps. From the other two terminals, signals were fed to two crossed half-wave dipoles through cables of RG 8/*U*, such that $L_4=L_3+\lambda g/4$. λg is the wavelength in the RG 8/*U* cable. The analysis indicates that the wave propagating from the antenna system is of the form

$$2\cos\frac{(W_1+W_2)}{2}t\ \angle\ \frac{(W_1-W_2)}{2}t.$$

The rate of rotation of the plane of polarization is thus 80 cps, with the direction determined by sign of f_1-f_2.

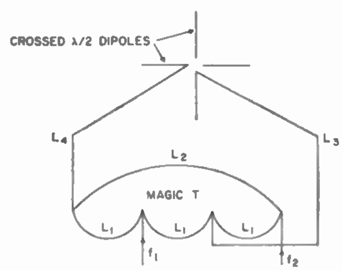

Fig. 1—Arrangement of the magic *T* and dipoles.

The choice of cable types was dictated by the condition that the characteristic impedance of the cable of the *T* must be $\sqrt{2}$ times that for the feed cables. The two cables chosen satisfied this condition closely. It was found that the *T* could be adjusted to an isolation of 40 db between opposite terminals and an equality of power division to within 0.25 db.

Two transmitters were used, capable of producing up to 45 watts each. One transmitter provided f_1 in Fig. 1, and also provided a reference signal to a SSB generator circuit which produced the drive for the second transmitter at a frequency 160 cps different from f_1 (Fig. 2).

Type-7360 tubes were used at 11 Mc as balanced modulators; their output tanks were detuned above and below the carrier frequency, and the outputs combined in a cathode follower to provide a SSB signal 80 cps removed from the original carrier.

The cathode follower output was injected into the grid circuit of a class-*C* Hartley oscillator.

The synchronized oscillator provided good limiting and hence suppression of residual carrier and unwanted sidebands, and

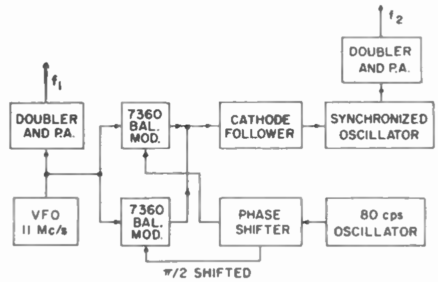

Fig. 2—Circuit for the generation of the signals of frequency f_1 and f_2.

gave a high-level signal to drive the doubler of the second transmitter without further amplification.

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* Received by the IRE, February 13, 1961.

[1] P. J. Allen, "Generating a rotating polarization," PROC. IRE (Correspondence), vol. 48, p. 941; May, 1960.

A Comparison Between Theoretical and Experimental Data on Phase Velocity of VLF Radio Waves*

The U. S. Navy Electronics Laboratory has obtained data which permit the absolute determination of the phase velocity of VLF radio waves.[1,2] The observations were made by transmitting one-second bursts of an unmodulated carrier from a master station. When the master station was silent, similar transmissions were made at the same frequency at two slave stations. The relative phases of the master and the two slave transmissions were observed at two receiving sites. The results of these experiments are to be published in a forthcoming paper by Pierce, Casselman, Tibbals, and Heritage. Mr. Tibbals sent me a copy of the manuscript for my comment. Since the subject is of great importance, I thought that my interpretation of their data might be of interest. For sake of completeness, a concise description of their technique is also given.

In the first set of experiments, in the spring of 1959, the master station was at Haiku, Hawaii, and a slave station was at San Diego, Calif. The receiving site was located at Wahiawa, also in Hawaii. The frequencies employed were 10.2, 11.2, 13.2, 14.2, and 15.2 kc. In the second set of experiments, in early 1960, a second slave station was located at Forestport in New York state and an additional receiving site was

* Received by the IRE, February 9, 1961.

[1] J. A. Pierce, C. J. Casselman, M. L. Tibbals, and D. P. Heritage, "The velocity of propagation of VLF radio waves," to be published.

[2] C. J. Casselman, D. P. Heritage, and M. L. Tibbals, "VLF propagation measurements for the Radux-Omega navigation system," PROC. IRE, vol. 47, pp. 829–839; May, 1959.

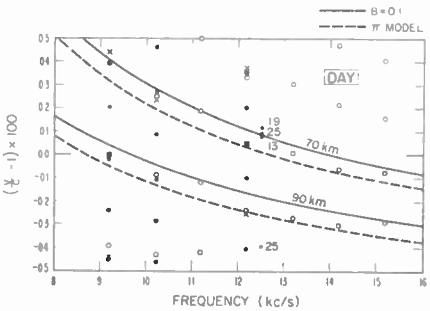
located near San Diego. The frequencies were then 9.2, 10.2, and 12.2 kc.

The data obtained in such experiments define only the phase, ϕ, of one signal with respect to another. The corresponding phase velocity v is related to ϕ by the equation

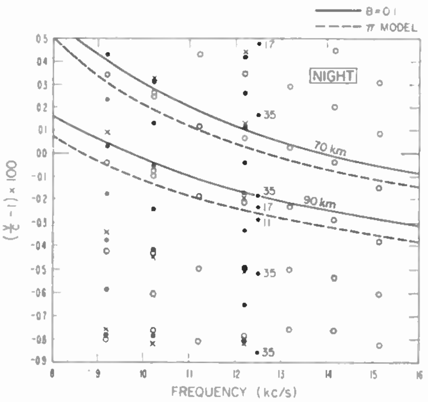
$$2\pi f \frac{d}{v} = 2m\pi + \phi,$$

where d is the total distance of travel and m is some unknown integer. An initial estimate of m can be made by taking the phase velocity v to be the same as the velocity of light. The remaining ambiguity can be resolved by plotting derived values of v as a function of frequency for various values of m. When this is repeated for several values of d, it becomes quite apparent which sets of points should be chosen. Essentially, this is the technique used by Pierce, Casselman, Tibbals, and Heritage.[1]

The situation is illustrated in Fig. 1(a) and Fig. 1(b), where possible values of the



(a)



(b)

Fig. 1—Phase velocity as a function of frequency for day and night. The curves are based on mode theory and the points are obtained from the experimental data of Pierce, Casselman, Tibbals, and Heritage.

phase velocity deviation $(v/c-1)$ are plotted vs frequency. The small circles correspond to the path from Hawaii to San Diego to Hawaii. The dots correspond to the path from Hawaii to Forestport, N. Y., to Hawaii. The crosses correspond to the path Hawaii to Forestport, to San Diego. Also shown in Figs. 1(a) and 1(b) are some points which represent data taken on a ship moving between California and Hawaii in August, 1959. The distance to San Diego in hundreds of kilometers is indicated beside each point. Because the ship's position was not known accurately, this set of data is not as precise.

Curves of $(v/c-1)$ derived from mode theory are now also drawn in Figs. 1(a) and 1(b). The full effect of earth curvature is taken into account by using Airy integral approximations for the relevant spherical wave functions.[3,4] The ionosphere is assumed to be sharply bounded and its lower edge is located at heights of 70 and 90 km. The dashed curves were calculated on the assumption that the upper boundary was equivalent to a perfect magnetical conductor and the lower boundary was a perfect electrical conductor. (Sometimes this is called the "π-model" since the phase shift on reflection at the ionosphere is π radians or 180°.) The solid curves are the result of applying a correction which takes into account the imperfect reflection at the upper boundary. In this example, a parameter B ($\propto$ the reciprocal of the ionospheric conductivity) was taken to be 0.1, which is quite typical.[3] The effects of the earth's magnetic field are neglected in the calculation since its influence on phase velocity is expected to be very small.[4]

It is quite apparent from studying Fig. 1(a) and Fig. 1(b) that the curve for $h=70$ km and $B=0.1$ is quite a good fit for the data in the daytime, whereas the curve for $h=90$ km and $B=0.1$ is more appropriate for the night. In fact, it would be very difficult indeed to find any other smooth curves which could be drawn in a satisfactory manner through the data points. Ideally, one should have coincidences between the dots, circles, and crosses for the correct value of $v/c-1$. Unfortunately, the variability of the relative phase is large enough to destroy any close coincidences. Furthermore, the theoretical curves refer only to the first order or dominant mode. In actuality, there are many modes present and only at very great distances may the higher modes be neglected. The presence of such modes is to render the phase velocity a function of distance. Thus, perfect coincidences would not be expected between the three kinds of points in Figs. 1(a) and 1(b) even if there were not any time variability. Another effect of higher modes is to introduce an undulation in the dispersion curve. There is some suggestion in Fig. 1(b) that the circles would fall on such a curve. The smallness of this undulation is actually strong evidence that only one mode is really significant at these ranges.

It can be concluded that there is fairly good agreement between measured phase velocity and the mode theory as developed for a curved earth-ionosphere waveguide. Furthermore, the heights of the equivalent reflecting layer are fairly consistent with other results.

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[3] J. R. Wait and K. Spies, "Influence of earth curvature and the terrestrial magnetic field on VLF propagation," *J. Geophys. Res.*, vol. 65, pp. 2325–2331; August, 1960.

[4] J. R. Wait, "A new approach to the mode theory of VLF propagation," *J. Res. NBS*, vol. 65D, pp. 37–46; January–February, 1961.

Reflection and Transmission of Conductive Films*

In a recent letter, Koide[1] has discussed the reflectivity of thin conducting films. The author derives a formula for the surface impedance of the film for the case of normal incidence and he obtains the reflection coefficient as a function of the intrinsic impedance, the propagation constant and the thickness of the film. It is interesting to note that the reflection coefficient is completely independent of frequency as long as the complex permittivity of the film remains constant.

The purpose of the present letter is

1) to present a more general relation for the reflection coefficient and transmission coefficient of thin conductive films for an arbitrary angle of incidence;
2) to give the first-order frequency-dependent term for the reflection and transmission coefficient;
3) to present the first-order correction terms in case of non-negligible displacement currents;
4) finally, to derive a formula for thick conducting films.

The formulas derived subsequently will be useful for evaluating the transmission and reflection coefficients for an arbitrary angle of incidence and for estimating the perturbation terms due to the frequency and the displacement current.

The term thin conductive films will only be applied for a film having the following properties:

1) The thickness, d, of the film must be small with respect to the skin depth, s

$$d < s = \sqrt{\frac{2}{\mu_2\mu_0\sigma\omega}} = \frac{1+j}{\beta_2}. \qquad (1)$$

2) The conductivity, σ, of the film must be large compared to the displacement conductivity of the film

$$\sigma > \epsilon_2\epsilon_0\omega. \qquad (2)$$

The following notation will be used:

$\alpha=\alpha_1$ = angle of incidence for a plane electromagnetic wave (Fig. 1)
$\gamma_m=\sqrt{\omega^2\mu_m\epsilon_m\mu_0\epsilon_0-j\omega\mu_m\mu_0\sigma_m}$ = propagation constant
$E\perp$ = electric field perpendicular to the plane of incidence
$H\perp$ = magnetic field perpendicular to the plane of incidence

In addition, we introduce the general impedance

$$g_m = \frac{\gamma_m \cos\alpha_m}{\omega\mu_m\mu_0} \quad \text{for } E\perp \quad m = 1, 2, 3 \qquad (3)$$

$$g_m = \frac{\omega\mu_m\mu_0 \cos\alpha_m}{\gamma_m} \quad \text{for } H\perp \quad m = 1, 2, 3 \qquad (4)$$

where sin α_m satisfies Snell's reflection law

* Received by the IRE, February 16, 1961.

[1] F. T. Koide, "Depth of penetration as a measure of reflectivity of thin conductive films," PROC. IRE (Correspondence), vol. 48, pp. 1654–1655; September, 1960.

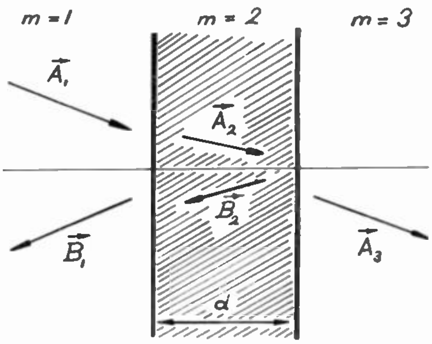

Fig. 1—Schematic representation of a plane electromagnetic wave incident on a plane boundary and transmitted into medium 3.

$$n_m \sin \alpha_m = n_{m+1} \sin \alpha_{m+1} \tag{5}$$

$$n_m = \sqrt{\mu_m \epsilon_m - j\frac{\mu_m \sigma_m}{\epsilon_0 \omega}}. \tag{6}$$

The reflection coefficient, r, and the transmission coefficient, t, may be computed by formulating the boundary conditions on each interface of the layer. Both coefficients are defined as the amplitude ratio of the tangential component of the electric field ($E\perp$) or the magnetic field ($H\perp$) on the interface and the corresponding incident field strength on the interface. The result is

$$r = \frac{1}{p}\left[(g_1 - g_2)(g_2 + g_3)e^{\beta_2 d} + (g_1 + g_2)(g_2 - g_3)e^{-\beta_2 d}\right] \tag{7}$$

$$t = \frac{4g_1 g_2}{p} \tag{8}$$

$$p = (g_1 + g_2)(g_2 + g_3)e^{\beta_2 d} + (g_1 - g_2)(g_2 - g_3)e^{-\beta_2 d}. \tag{9}$$

A general discussion as well as recursion formulas for any discrete number of layers have been published by Wolter.[2]

If we take vacuum for the medium $m=1$ and $m=3$, we have $\sigma_1 = \sigma_3 = 0$, $\mu_1 = \mu_3 = 1$ and $\epsilon_1 = \epsilon_3 = 1$. We put

$$\sigma \equiv \sigma_2 \quad \text{and} \quad \beta \equiv \beta_2 \tag{10}$$

and we obtain by a straightforward calculation from (7)–(9), by neglecting the displacement current and by using only the first three terms of the series expansion,

$$e^{\pm\beta d} \approx 1 \pm \beta d + \frac{(\beta d)^2}{2} \tag{11}$$

$$r = -\frac{1}{1+\sqrt{\frac{\epsilon_0}{\mu_0}}\,\frac{2\cos\alpha}{\sigma d}\left[1+\frac{(\beta d)^2}{2}\right]} \quad E\perp \tag{12}$$

$$t = +\frac{1}{1+\sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{2\sigma d}{\cos\alpha}\,\frac{(\beta d)^2}{2}} \quad E\perp \tag{13}$$

$$r = +\frac{1}{1+\sqrt{\frac{\epsilon_0}{\mu_0}}\,\frac{2}{\sigma d\cos\alpha}\left[1+\frac{(\beta d)^2}{2}\right]} \quad H\perp \tag{14}$$

$$t = +\frac{1}{1+\sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{\sigma d\cos\alpha}{2}+\frac{(\beta d)^2}{2}} \quad H\perp. \tag{15}$$

The perturbation term, $(\beta d)^2$, increases linearly with increasing frequency and may be disregarded if the thickness of the film is extremely small with respect to the skin depth. This will lead finally to the frequency independent reflection and transmission coefficient as mentioned above.

It is important to note that the effective thickness of the film is

$$d_{\text{eff}} = \frac{d}{\cos\alpha} \quad E\perp \tag{16}$$

$$d_{\text{eff}} = d\cos\alpha \quad H\perp. \tag{17}$$

From this it follows that the film has the properties of a polarization filter if $\alpha \approx \pi/2$.

Additional terms are added if the perturbation due to the displacement current is computed. The first-order terms are

$$-\frac{1}{r} = 1 + \sqrt{\frac{\epsilon_0}{\mu_0}}\,\frac{2\cos\alpha}{\sigma d} \cdot \left[1 - j\frac{\epsilon_2\epsilon_0\omega}{2\sigma} + j\frac{\epsilon_0\omega\sin^2\alpha}{2\mu_2\sigma}\right] \quad E\perp \tag{18}$$

$$\frac{1}{t} = 1 + \sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{\sigma d}{2\cos\alpha} \cdot \left[1 + j\frac{\epsilon_2\epsilon_0\omega}{2\sigma} - j\frac{\epsilon_0\omega\sin^2\alpha}{2\mu_2\sigma}\right] \quad E\perp \tag{19}$$

$$\frac{1}{r} = 1 + \sqrt{\frac{\epsilon_0}{\mu_0}}\,\frac{2}{\sigma d\cos\alpha} \cdot \left[1 - j\frac{\epsilon_2\epsilon_0\omega}{2\sigma} - j\frac{\epsilon_0\omega\sin^2\alpha}{2\mu_2\sigma}\right] \quad H\perp \tag{20}$$

$$\frac{1}{t} = 1 + \sqrt{\frac{\mu_0}{\epsilon_0}}\,\frac{\sigma d\cos\alpha}{2} \cdot \left[1 + j\frac{\epsilon_2\epsilon_0\omega}{2\sigma} + j\frac{\epsilon_0\omega\sin^2\alpha}{2\mu_2\sigma}\right] \quad H\perp. \tag{21}$$

The term $(\beta d)^2/2$ may be added, if necessary, according to the formulas (12)–(15).

A thick conducting film is defined by

$$\left|e^{+\beta d}\right| \gg \left|e^{-\beta d}\right| \tag{22}$$

and we obtain directly

$$r = -1 + \cos\alpha\sqrt{\frac{2\epsilon_0\mu_2\omega}{\sigma}}(1+j) \quad E\perp \tag{23}$$

$$t = 2\cos\alpha\sqrt{\frac{2\epsilon_0\mu_2\omega}{\sigma}}(1+j)e^{-\beta d} \quad E\perp \tag{24}$$

$$r = +1 - \frac{1}{\cos\alpha}\sqrt{\frac{2\epsilon_0\mu_2\omega}{\sigma}}(1+j) \quad H\perp \tag{25}$$

$$t = \frac{2}{\cos\alpha}\sqrt{\frac{2\epsilon_0\mu_2\omega}{\sigma}}(1+j)e^{-\beta d} \quad H\perp. \tag{26}$$

As expected, the reflection coefficient is of the order -1 in the electric case and $+1$ in the magnetic case. The perturbation term is always complex and of course frequency-dependent.

The author thanks W. Tschopp for valuable discussions. The aid of the "Schweizerischer Nationalfonds zur Förderung der Wissenschaften" is gratefully acknowledged.

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[2] H. Wolter, "Optics of thin films," in "Encyclopedia of Physics," Springer-Verlag, Berlin, Germany, vol. 24, pp. 469–473; 1956.

A Mechanism for Direct Adjacent Channel Interference*

The type of crosstalk described in the title is likely to occur in systems which have closely packed FM channels such as broadband microwave systems or FM multiplex systems. The phenomenon can be described with the aid of Fig. 1. Assume that the Channel 1 input is an unmodulated carrier and the Channel 2 input is a frequency modulated carrier. In these circumstances, it is possible for the modulation on Channel 2 to appear in a clear, ungarbled form in the baseband output circuit of Channel 1. This is true in spite of the fact that $|\omega_1 - \omega_2|$ is much greater than the highest baseband passed by the output filter in Channel 1. The resulting interference is not, and cannot be, the result of simple beats between the Channel 1 carrier and the spectrum of the signal in Channel 2.

This type of interference was first noted by J. G. Chaffee in 1939, and was observed in the TD-2 Radio System in 1955 by S. D. Hathaway. The problem was brought to the attention of the author by H. E. Curtis of Bell Telephone Laboratories. In view of the continuing interest in this problem and its importance in broad-band FM systems,[1] it was thought that this mechanism may be of general interest.

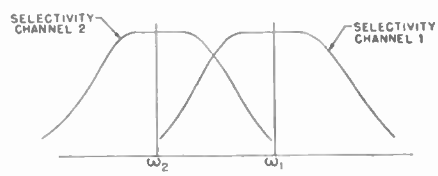

Fig. 1—Channel arrangement resulting in direct adjacent channel interference.

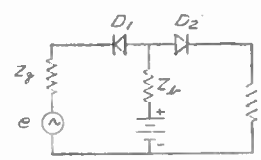

Fig. 2—Limiter circuit.

The purpose of a limiter in an FM receiver is to remove any incidental amplitude modulation on the incoming signal. This AM suppression can be viewed as a combination demodulation and remodulation process. A theory[2] has been worked out along these lines, and the predictions from this theory agree remarkably well with laboratory experiments on practical limiters. Thus, if the limiter input (see Fig. 2) is an amplitude modulated carrier, the combination of D1 and Zb forms a detector which detects the AM signal and causes a voltage, at the modulating frequency, to appear across Zb.

* Received by the IRE, February 27, 1961.

[1] H. E. Curtis, T. R. D. Collins, and B. C. Jamison, "Interstitial channels for doubling TD-2 radio system capacity," *Bell Sys. Tech. J.*, vol. 39, pp. 1505–1527; November, 1960.

[2] C. L. Ruthroff, "Amplitude modulation suppression in FM systems," *Bell Sys. Tech. J.*, vol. 37, pp. 1023–1046; July, 1958.

This voltage remodulates the carrier in D2, creating AM sidebands which tend to cancel the original AM.

With this brief description of limiting, the mechanism of direct adjacent-channel interference can be understood. The FM signal in Channel 2 (Fig. 1) is partly converted to AM on the frequency selectivity skirts of Channel 1. This AM signal is demodulated in the limiter of Channel 1 resulting in the Channel 2 modulation appearing across Zb. This is in turn remodulated on the carrier of Channel 1 as amplitude modulation. The discriminator is, of course, sensitive to AM, so the Channel 2 modulation appears at the output of Channel 1. It remains to examine the behavior of the interference with respect to the amplitude ratio of the desired to undesired signals.

After the signal on Channel 2 is converted to AM, the limiter of the Channel 1 receiver has the following input:

$$\begin{aligned} e &= E[\cos\omega_1 t + I(1 + k\cos pt)\cos\omega_2 t] \\ &= E\Big[\cos\omega_1 t + I\cos\omega_2 t \\ &\quad + I\frac{k}{2}(\cos(\omega_2 + p)t + \cos(\omega_2 - p)t)\Big] \quad (1) \end{aligned}$$

where,

- I is the ratio of the carrier voltage of Channel 2 to that in Channel 1, at the limiter input of Channel 1
- k is the AM index as a result of FM-AM conversion of the FM on Channel 2 on the skirts of Channel 1
- ω_1 is the carrier frequency of Channel 1
- ω_2 is the carrier frequency of Channel 2
- p is the modulating frequency on Channel 2.

In the limiter diode D1, new frequencies will be generated corresponding to the sums, differences, and harmonics of the frequencies in (1). If any new frequency is terminated in a finite impedance, it, too, becomes an input and the frequency generating process continues.[3] The current in the diode can be represented by a power series of the input voltage.

$$i_D = a_1 e + a_2 e^2 + \cdots a_n e^n \cdots \quad (2)$$

Consider the nth term,

$$\begin{aligned} a_n e^n &= a_n E^n[\cos\omega_1 t + I(1 + k\cos pt)\cos\omega_2 t]^n \quad (3) \\ &= a_n E^n\Big[\cos^n\omega_1 t + n\cos^{n-1}\omega_1 t\, I(1+k\cos pt)\cos\omega_2 t \\ &\quad + \frac{n(n-1)}{2!}\cos^{n-2}\omega_1 t I^2 \cdot[(1 + k\cos pt)\cos\omega_2 t]^2 \\ &\quad + \frac{n(n-1)(n-2)}{3!}\cos^{n-3}\omega_1 t I^3 \cdot[(1 + k\cos pt)\cos\omega_2 t]^3 \\ &\quad + \cdots\Big]. \end{aligned}$$

The first term represents harmonics of the Channel 1 carrier. The second term gives distortion predicted in low-index theory which falls outside the output filter of Channel 1 for this case. The third term contributes to direct adjacent channel interference. In addition to terms falling outside the frequency range of interest, the third term includes

$$e_b = \frac{n(n-1)E^n}{2!2^{n-2}} K^{n-2}_{(n-2)/2} I^2 k\cos pt, \quad n \text{ even} \quad (4)$$

where $K^{n-2}_{(n-2)/2}$ is equal to the combination of $n-2$ things taken $(n-2)/2$ at a time.

Therefore, a current at the frequency p is flowing in D1 and a voltage across Zb will appear in proportion. This voltage will remodulate ω_1 in D2, giving the direct adjacent-channel interference effect.

The magnitude of the interference is proportional to I^2k. Therefore, if the interfering Channel 2 is suppressed one db, the interference drops 2 db. If the index of modulation on Channel 2 is increased one db, the interference increases one db since the factor k is proportional to the FM index on Channel 2. (Assuming a constant slope skirt.)

Higher-order distortion terms will result from terms beyond the third in (3), but these will be much smaller in magnitude than the interference due to the third term.

Once the AM is generated in the limiter, the main damage is done. A second limiter would remove the AM generated in the first, but the same mechanism will occur in the second limiter. In general, practical discriminators are very sensitive to AM and it is also known that the AM generated in the limiter can be converted to FM in the limiter. This, of course, would "freeze" the interference in Channel 1 and it would then be impossible to remove it.

The foregoing explanation was based on the limiter, but the mechanism is obviously valid for any nonlinear components, including amplifiers, in compression and discriminators.

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[3] R. S. Caruthers, "Copper oxide modulators in carrier telephone systems," *Bell Sys. Tech. J.*, vol. 18, pp. 315–337; April, 1939.

Tunnel-Diode Binary Counter Circuit*

* Received by the IRE, January 30, 1961; revised manuscript received, February 23, 1961.

Fig. 1 shows a tunnel-diode circuit with a dc load line shown in Fig. 2. This load line establishes two stable dc operating points, A and C. Under ordinary conditions, the load line is determined mainly by R.

Assume that the circuit is at state A (Fig. 2). A positive input pulse of sufficient amplitude would move the operating point beyond the knee and the operating point would switch. The current in the inductance would not change during switching and the operating point would move to B. The current in the inductance would then start to decay and the operating point would finally rest at C.

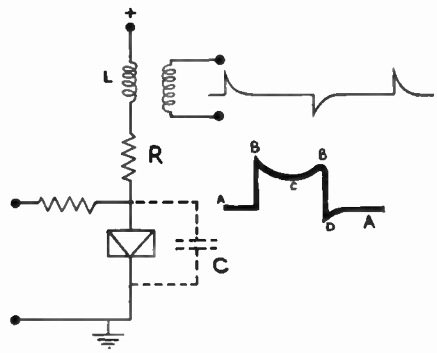

Fig. 1—Single tunnel-diode flip-flop.

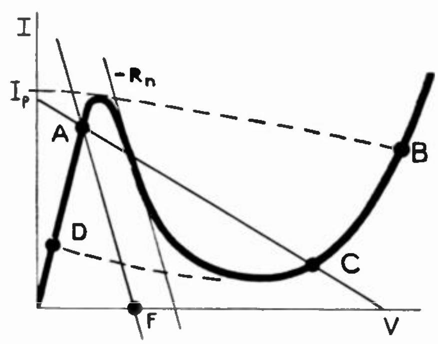

Fig. 2—Operating conditions for single tunnel-diode flip-flop and the monostable circuit.

When a second positive pulse is applied, the voltage on the tunnel-diode will rise (about to B) and start to decay. If the reactive elements of the circuit are such that the circuit is underdamped there will be a negative overshoot. This may bring the operating point below the knee, and hence the diode will switch back to A through D. A typical waveform is shown in Fig. 1. To understand why the negative overshoot occurs only at the second pulse, consider the current initial conditions: on switching from A there is a high initial current I_A in the inductance, while in the other case it is only I_C, with I_C the steady-state value. For transient quasi-linear analysis, the initial current in the inductance could be considered $I_b - I_c$ in the first case and zero in the second. The capacitance initial voltage (after the input pulse is over) is $V_b - V_c$ in both cases. Examination will show that the component of the transient, due to the excited inductance, will cancel the negative overshoot which is present in the component due to the capacitance charge.

Design values for Ge diodes are about

$$R = \frac{400}{I_p}, \qquad \sqrt{L/C}/R = 3.5.$$

R is measured in ohms, and I_p, the diode peak current, in milliamperes.

From the description of the operation, it is clear that the duration of the input pulses has to be short compared to the time constant of the circuits. Hence, if the input pulses are too long, it may be necessary to add capacitance.

The voltage obtained across the inductance, or on a secondary turn, is shown in Fig. 1. Considering only single polarity pulses, a pulse (say positive) is obtained only for every other input pulse. This could be used as a "carry" in counting.

A convenient way to eliminate the negative pulse, and at the same time to amplify and reshape the positive pulse, is to feed it into a monostable amplifier. The monostable circuit is essentially similar to that of Fig. 1, but with a load line AF that cuts the characteristic only at one point. A positive input pulse triggers it, and it produces a square wave while a negative input pulse does not.

Conclusion

The above-described circuit will alternate between two voltage states, high and low, with every input pulse. It will also provide an output pulse for every second input pulse; this would make it eminently suitable as a counter.

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New Ways to Trigger Avalanche Pulse Circuits*

The problem of triggering avalanche pulse stages is considered in this note. A good trigger system should fulfill the following conditions:

1) High sensitivity without responding to noise.
2) Low delay and negligible jitter between trigger and output pulse.
3) No feedback of pulse energy into the trigger circuit.

Base triggering, as shown in Fig. 1, seems most suitable because switching occurs mainly between emitter and collector. In the stand-by condition, the collector voltage V_C is adjusted close to the avalanche breakdown potential where the multiplication factor, M, is high. To prevent regenerative avalanche breakdown, the minority current, I_E, from the emitter junction must be below a critical value. If $V_B > r_B' I_B$, practically no minority carriers are introduced by the reverse-biased emitter junction and $I_E \leqq 0$. Nearly all the collector current $I_C = MI_{CO}$ flows through the base.

$$I_B = -I_C \frac{(M\alpha_0 - 1)}{(M\alpha_0)} \cong -I_C.$$

Avalanche breakdown starts as soon as more than the critical number of minority carriers are generated as a result of increased forward bias across the E-B junction. The trigger sensitivity can be optimized in the following ways:

1) Adjust the collector potential close to the avalanche-breakdown potential.
2) Set the base bias close to the potentential where regenerative multiplication starts.
3) Reduce the base resistance r_B' to the lowest possible value so that most of the trigger pulse energy reaches the emitter junction.

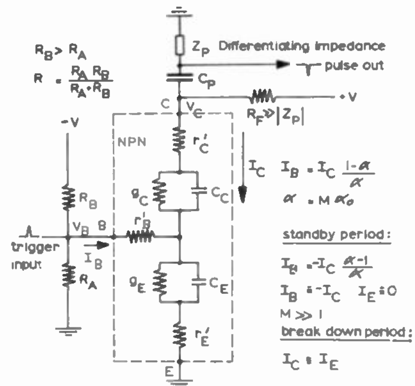

Fig. 1—Base-triggered avalanche stage.

Component variation and temperature changes of transistor parameters force a high safety factor and consequently reduce trigger sensitivity. Particularly damaging is the temperature sensitive collector current MI_{CO}. The voltage drop $MI_{CO}(R + r_B')$ tends to forward bias the emitter junction. The situation can be improved if r_B' and MI_{CO} are kept as low as possible. Reduction of R is only helpful to the point where the gain increase in sensitivity is offset by the loss of trigger energy in R_A.

Better results are possible if the resistor R_A is replaced by a tunnel diode as shown in Fig. 2. The circuit is very sensitive and works reliably if the base resistance r_B', and the current, MI_{CO}, are reasonably low. Reverse bias V_E is applied at the emitter side to compensate for the forward bias V_B across the tunnel diode. A highly capacitative forward-biased diode between emitter and ground supplies the dc bias and produces an ac short at the same time. After breakdown, the tunnel diode must be reset to its original state. Fig. 3 shows an avalanche power device triggered from the emitter side. The resistor R_E between emitter and ground should be large during the stand-by period for high trigger efficiency and small during the breakdown period to prevent power losses and feedback of the output pulse into the trigger circuit. A tunnel diode instead of R_E fulfills these requirements. The circuit has the following features:

1) dc bias can easily be adjusted from the base side. An inhibit circuit may be used.
2) I_B is low during the standby period which allows biasing the tunnel diode very closely to the switching threshold.
3) The avalanche-breakdown current drives the tunnel diode back into the reverse high-conduction state, thus reducing losses and feedback of pulse energy into the trigger reset.
4) The tunnel diode is automatically reset.

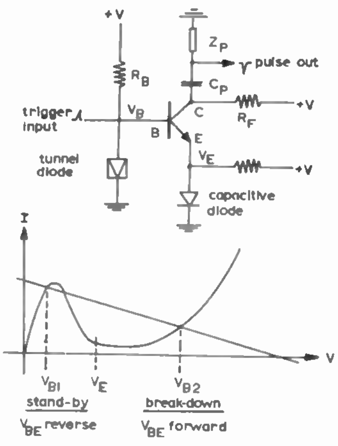

Fig. 2—Base-triggered avalanche stage with tunnel diode.

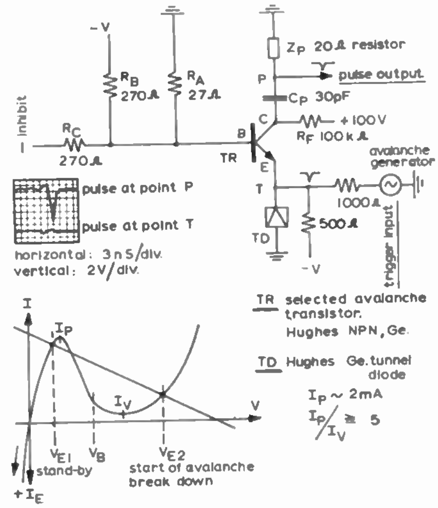

Fig. 3—Emitter-triggered avalanche stage with tunnel diode.

Input and output waveforms in Fig. 3 show that very little energy flows back into the trigger circuit. The time delay of 2–4 mμsec between input and output pulse depends on the lead inductance and stray capacitance. Our test circuit was much easier to control from the emitter than from the base side.

In summary, it can be said that the tunnel diode improves the sensitivity and stability of base-triggered, and especially of emitter-triggered, avalanche circuits. A promising application may be their use as power drivers in microminiaturized form for tunnel diode logic.

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Discussion of "Combined AM and PM for a One-Sided Spectrum"*

In a recent communication,[1] Chakrabarti proposes a simple means of obtaining a one-sided spectrum. His method is to modulate the phase as well as the amplitude of a carrier. The modulating signal, a sine wave, is shifted 90° before being applied to the phase modulator. While this method appears to be an elegant means of obtaining a single-sideband spectrum, it does not, after scrutinization, actually yield the desired result. In what follows, it will be shown that for the case considered by Chakrabarti, *only* the first lower sideband frequency component vanishes.

Consider a sinewave carrier of angular frequency ω_0, modulated by a signal of angular frequency ω_s such that the resultant signal is

$$r(t) = (1 + m \cos \omega_s t) \sin (\omega_0 t + m \sin \omega_s t). \quad (1)$$

This can also be written in complex form as

$$r(t) = \operatorname{Im} \{(1 + m \cos \omega_s t) \cdot \exp [i(\omega_0 t + m \sin \omega_s t)]\}, \quad (2)$$

where Im denotes the imaginary part. Using the relation[2]

$$\exp (im \sin \omega_s t) = \sum_{n=-\infty}^{\infty} J_n(m) \exp (in\omega_s t), \quad (3)$$

one can write

$$r(t) = \operatorname{Im} \left\{\left[1 + \frac{m}{2} [\exp (i\omega_s t) + \exp (-i\omega_s)]\right] \cdot \sum_{n=-\infty}^{\infty} J_n(m) \exp i(\omega_0 + n\omega_s)t \right\},$$

or

$$= \operatorname{Im} \left\{ \sum_{n=-\infty}^{\infty} \left[J_n(m) + \frac{m}{2} [J_{n+1}(m) + J_{n-1}(m)]\right] \exp i(\omega_0 + n\omega_s)t \right\}. \quad (4)$$

Applying the well-known recursion formula[2]

$$nJ_n(m) = \frac{m}{2} [J_{n+1}(m) + J_{n-1}(m)], \quad (5)$$

and taking the imaginary part of (4) yields

$$r(t) = \sum_{n=-\infty}^{\infty} (n + 1)J_n(m) \sin (\omega_0 + n\omega_s)t, \quad (6)$$

the Fourier series expansion of (1). From (6), it is seen that only the term for which $n = -1$ (*i.e.*, the first lower sideband frequency component) vanishes identically for all values of m.

Although the above combination of AM-PM does not produce a spectrum that vanishes identically for all negative values of n, there does exist a combined AM-PM system which will give a one-sided spectrum.

Consider writing (1) in the more general form,

$$r(t) = [1 + a(t)] \cos [\omega_0 t + \phi(t)], \quad (7)$$

where the amplitude modulation function satisfies the inequality

$$1 + a(t) > 0. \quad (8)$$

Now if the phase modulation $\phi(t)$ is related to the amplitude function by

$$\phi(t) = \frac{1}{\pi} \int_{-\infty}^{\infty} \frac{\ln [1 + a(\tau)]}{t - \tau} d\tau, \quad (9)$$

where ln denotes the natural logarithm, then $r(t)$ has no frequency component smaller than the carrier frequency ω_0, provided the integral exists.[3]

For small values of $a(t)$, $a(t) \ll 1$, the logarithm in (9) may be expanded and higher-order terms involving powers of $a(\tau)$ neglected. Hence,

$$\phi(t) \approx \frac{1}{\pi} \int_{-\infty}^{\infty} \frac{a(\tau)}{t - \tau} d\tau \quad \text{for } a(t) \ll 1. \quad (10)$$

The integral on the right is the "Hilbert transform" (quadrature signal) of $a(t)$. In the frequency domain, the Hilbert transform shifts the phases of all the spectral components by $-90°$. Specifically, if

$$a(t) = m \cos \omega_s t, \quad (11)$$

(10) yields

$$\phi(t) \approx m \sin \omega_s t \quad \text{for } m \ll 1, \quad (12)$$

in agreement with the original suggestion of Chakrabarti, but now restricted to small values of m.

The fact that (6) will yield approximately a one-sided spectrum for small values of the modulation index m will now be shown directly from (6).

For small values of m, the Bessel function can be approximated by[2]

$$J_n(m) \approx \frac{(m/2)^n}{n!} \quad \text{for } n \geq 0 \quad (13a)$$

and

$$J_n(m) \approx (-1)^n \frac{(m/2)^{-n}}{(-n)!} \quad \text{for } n \leq 0. \quad (13b)$$

Substituting this relation in (6) and neglecting second- and higher-order terms in m gives a relation involving only the carrier and the first upper-sideband frequency component:

$$r(t) \approx \sin \omega_0 t + m \sin (\omega_0 + \omega_s)t \quad \text{for } m \ll 1. \quad (14)$$

In a later communication, Taylor[4] points out that the "identity" between combined AM-PM and a single-sideband modulation was illustrated and discussed in 1937 by Barkhausen.[5] However, Taylor does not mention that Barkhausen correctly states that this relation is only an approximation for small values of m.

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* Received by the IRE, October 20, 1960.

[1] N. B. Chakrabarti, "Combined AM and PM for a one-sided spectrum," Proc. IRE, vol. 47 (Correspondence), p. 1663; September, 1959.

[2] A. Sommerfeld, "Partial Differential Equations," Academic Press, Inc., New York, N. Y., pp. 84–108; 1949.

[3] For a recent application of this relation and further references, see K. H. Powers, "The compatibility problem in single sideband transmission," Proc. IRE, vol. 48, pp. 1431–1435; August, 1960. See especially p. 1433.

[4] A. H. Taylor, "Combined AM and PM for a one-sided spectrum," Proc. IRE (Correspondence), vol. 48, p. 953; May, 1960.

[5] H. Barkhausen, "Electronon-Röhren," Verlag S. Hirzel, Leipzig, Germany, vol. 4, pp. 159 ff; 1937.

Author's Reply[6]

The author is in full agreement with the remark of Golden and Schroeder that the scheme presented[1] applies to the low-modulation index. He was fully aware that only the first-order sideband will be completely cancelled. One could readily calculate that under the conditions envisaged in the letter under discussion the magnitudes of the second- and third-order lower sidebands are respectively J_2 and $2J_3$. No reference was made to these, as their magnitudes are much smaller than the corresponding higher-order sidebands ($3J_2$ and $4J_3$).

We shall here briefly indicate how the magnitudes of the higher-order sidebands on either side can be reduced. Our consideration will be limited to single-tone modulation. If A_n denotes the magnitude of the nth sideband before the carrier is amplitude modulated by a signal $e_s(t) = 1 + m_a \cos \theta$, then the corresponding amplitude after the modulator is

$$B_n = (A_{n-1} + A_{n+1}) \frac{m_a}{2} + A_n.$$

To ensure better reduction of the second- or third-order sidebands than was possible in the simple scheme, one notes that the magnitude of A_{-2} has to be greater and A_{-2} has to be negative. In the practical setup, the three following schemes were tried:

1) Introducing, by means of a nonlinear shaper, a controlled amount of harmonic distortion in the right phase in the phase-modulating signal prior the 90° phase shifter.
2) Adding an asymmetrical sideband carrier modulated by the harmonics of the phase modulating signal, and limiting.
3) Limiting the output of an asymmetric sideband carrier obtained through modulating two 90° phase-shifted carriers by phase-shifted signals ($C \cos \theta + S \sin \theta$ and $C \cos \theta - S \sin \theta$).

The trouble is that the cancellation of the first-order sideband is no longer automatically complete: the ratio of the magnitude of phase-modulation index to AM index will have to be adjusted for good rejection with variation of modulation depth by means of an amplitude-control arrangement. In case 1), this gain control is very well provided by the shaper itself for certain shaping characteristics. (On the debit side the bandwidth of the phase shifter is considerably greater than that ordinarily required.) The gain control circuit is either a volume expander for the phase modulator or volume compressor for the amplitude modulator.

Let us consider a specific example of case 1). Suppose that the phase-modulated carrier is given by

$$\sin (\omega_0 t + m_p \sin \theta - m_2 \sin 2\theta + m_3 \sin 3\theta).$$

Here m_2 and m_3 are proportional respectively to m_p^2 and m_p^3, the constants of proportionality being adjusted to minimize the magnitude of the undesired components. Taking $m_p = 0.5$, $m_2 = 0.08$, and $m_3 = 0.02$, one gets approximately

[6] Received by the IRE, March 7, 1961.

$A_{+1} = 0.252,\quad A_{-1} = 0.232,\quad A_{+2} = -0.01,$
$A_{+2} = 0.067,\quad A_{+3} \simeq 0,\quad A_{-3} = -0.02$
and
$A_0 = 0.938.$

On adjusting $m_a = 0.475$, one gets

$$B_{-1} = 0.006,\qquad B_{+1} = 0.473,$$
$$B_{-2} = 0.01,\qquad B_{+2} = 0.045.$$

Similarly, for

$$m_p = 0.80,\quad m_2 = 0.22,\quad m_3 = 0.08 \text{ and } m_a = 0.66,$$

one gets

$$B_{+1} = 0.734,\qquad B_{-1} = 0.005,$$
$$B_{+2} = 0.100,\qquad B_{-2} = -0.040.$$

Now to consider cases of scheme 3) in similar conditions, let the modulated carrier prior to the limiter be represented by

$$e(t) = \sin \omega_0 t + A_u \sin (\omega_0 t + \theta) - A_l \sin (\omega_0 t - \theta).$$

For $A_u = 0.375$ and $A_l = 0.125$, the relative magnitudes of the sidebands after the limiter will be given approximately by

$$A_{+1} = 0.267,\qquad A_{-1} = 0.233,$$
$$A_{+2} = -0.02\qquad A_{-2} = 0.034.$$

Taking $m_a = 0.475$, we obtain

$$B_{+1} = 0.505,\quad B_{+2} = 0.043,\quad B_{-2} = -0.021.$$

Similarly, when $A_u = 0.60$ and $A_l = 0.20$, one calculates

$$B_{+1} = 0.720\quad B_{-2} = -0.040,\quad B_{+2} = 0.090.$$

These figures are obviously not the optimum values. We should like only to note that it is quite a practical proposition to keep the total undesired sidebands on the wrong side to within 5 per cent, and the total higher-order sidebands on the same side to within 10 per cent for an equivalent modulation depth not exceeding 80 per cent. It is felt that since the energy distribution of common signals is sparse at higher frequencies, the in-band distortion can be permitted within reasonable limits, in which case it is possible to reduce further the out-of-band undesired sidebands.

In conclusion, the author would like to mention that "combined AM-PM" has been successfully applied to pulse modulation with an almost 50 per cent reduction in bandwidth. The carrier to be pulse modulated is first phase modulated by means of a repetitive pulse sequence. The repetitive wave is to be skew-symmetric, that is, to change sign at the pulse center (corresponding to the requirement of 90° phase shift in the LF case), its time period should bear a certain relationship to the duration of the pulse, depending on the latter's waveform. The resultant spectrum is due to the sum of spectra of modulated carriers centered at the carrier and the combination frequencies of the carrier and the phase-modulating wave, which are of right phases to give cancellation on one side and augmentation on the other.

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Approximate Solution to Semiconductor Noise as a Queuing Problem*

Bell[1] interprets the charge carriers in the conductor band as the waiting line of a queuing operation whose customers (excitation from the base levels) arrive at random, *i.e.*, by a Poisson process:

$$\frac{(\lambda t)^n e^{-\lambda t}}{n!}\qquad (n = 0, 1 \cdots).$$

The service channels of a queue whose total number is c correspond to c vacant base levels to which the charge carriers can return, and the duration of time in which the base level is occupied by a carrier (until it is excited into the conduction band) has an exponential distribution $1-e^{-\mu t}$. The mean $1/\mu$ is fixed by the excitation energy through the thermodynamic relationship between "free" and "bound" times. The number of carriers is assumed governed by the number of excitations (although the two have equal expected values, they will not tally precisely all the time). It is desired to compute the semiconductor noise spectrum.

In his paper, Bell attempted the use of an approximation by Pollaczek[2] which led to considerable analytical difficulty and hence was unable to compute the power spectrum. We shall use an important and useful approximation to carry out the calculation. One first obtains $P(>t)$, the probability that an excitation is delayed for a time greater than t due to the occupation of all base levels.

Riordan,[3] in studying a Poisson input queue with c multiple parallel channels, each of which serves arriving units by an exponential distribution with identical means, and where waiting units are selected for service at random, gives

$$P(>t) = \tfrac{1}{2}(1 - \sqrt{\rho/2})e^{-t(1-\rho)(1-\sqrt{\rho/2})} + \tfrac{1}{2}(1 + \sqrt{\rho/2})e^{-t(1-\rho)(1+\sqrt{\rho/2})}$$

where

$$\rho \equiv \frac{\lambda}{c\mu}.$$

This solution is valid for $0 \leq \rho \leq 0.7$. Curves have been plotted for various values of ρ by Wilkinson.[4] It is well known in the theory of queues, and also pointed out by Bell, that for $\rho \geq 1$ both the queuing problem and the corresponding semiconductor noise problem have divergent solutions. Some reflection will show that the queue size becomes infinite if the input rate equals or exceeds the total service rate.

Thus, the auto-correlation function is given by

$$\psi(\tau) = \int_0^\infty P(>t)P(>t+\tau)dt$$
$$= \frac{1}{8(1-\rho)}\left[(3-\rho-\sqrt{\rho/2})e^{-(1-\rho)(1-\sqrt{\rho/2})} + (3-\rho+\sqrt{\rho/2})e^{-(1-\rho)(1+\sqrt{\rho/2})}\right].$$

The cosine Fourier transform of $\psi(\tau)$ is an even function of τ, and hence the frequency spectrum is given by

$$W(f) = 4\int_0^\infty \psi(\tau)\cos 2\pi f t d\tau$$
$$= \frac{1}{2}\,\frac{(3-\rho-\sqrt{\rho/2})(1-\sqrt{\rho/2})}{(1-\rho)^2(1-\sqrt{\rho/2})^2 + (2\pi f)^2} + \frac{1}{2}\,\frac{(3-\rho+\sqrt{\rho/2})(1+\sqrt{\rho/2})}{(1-\rho)^2(1+\sqrt{\rho/2})^2 + (2\pi f)^2},$$

which is valid for $0 \leq \rho \leq 0.7$.

This is the classical case of the superposition of two low-pass filter characteristics. An approximate behavior of the frequency spectrum as $1/f$ is usually expected. However, it can be seen by examining the components of our solution that the spectrum behaves as $1/f^\alpha$, $1 < \alpha < 2$. There is now the mathematical dilemma that it is impossible for $1/f^\alpha$ spectrum to arise as the transform of a correlation function because the integral would not converge. The way out of this dilemma is that the spectrum is the transform of a correlation function of an infinite series of phenomena.[5] As pointed out by Bell, accumulating evidence shows that a law $1/f^\alpha$, with α a little greater than unity, tends to occur when the excitation energy is large compared with the thermal energy. Evaluation at the break-point, when the frequency term equals the constant term, shows that $\rho = 0.7$ (which yields the minimum value for the smaller constant term), gives $f \sim 0.03$, indicating the possible usefulness of the queuing representation for low frequencies. Semiconductors are used in these low-frequency ranges in switching for computer applications where low noise levels corresponding to frequencies of this order are prohibitive. They are also used in submarine dc battery inverters, in telemetering from space vehicles, in measuring long periods of fluctuation such as measuring noise from actual sea pounding, in the generation of class A amplifiers and more generally in power supplies in which there are thermal fluctuations.

A main use of this representation lies in the possibility of transferring ideas about queuing problems with measures taken to relieve congestion, etc.,[6,7] to the treatment of semiconductor noise problems.

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Taylor-Cauchy Transforms for Analysis of Varying-Parameter Systems*

The method of Taylor-Cauchy transforms[1] can be easily applied to the analysis of varying-parameter systems, as shown by six examples below. Additional transform pairs are given in Table I.

Example 1

Given

$$t\frac{dy}{dt} = kt^k \tag{1}$$

where $y=0$ at $t=0$. According to Ku, *et al.*,[1] we rewrite (1) as

$$\lambda W^{(1)}(\lambda) = k\lambda^k. \tag{2}$$

Taking the Taylor-Cauchy transform of both members of (2) gives

$$w_{n-1} = k\delta_{n-k} \tag{3}$$

where $\delta_{n-k}=1$ for $n=k$ and zero otherwise. So we get $w_{k-1}=k$ and

$$W^{(1)}(\lambda) = k\lambda^{k-1}. \tag{4}$$

Integrating with respect to λ gives

$$W(\lambda) = \lambda^k. \tag{5}$$

This corresponds to $y(t)=t^k$.

Example 2

Given

$$\left[t^2\frac{d^2}{dt^2} + t\frac{d}{dt} + t^2\right]y = 0. \tag{6}$$

The initial conditions are: at $t=0$, $y=1$ and $y'=(dy/dt)=0$. We rewrite (6) as

$$\lambda^2 W^{(2)}(\lambda) + \lambda W^{(1)}(\lambda) + \lambda^2 W(\lambda) = 0. \tag{7}$$

Taking the Taylor-Cauchy transform of (7) gives

$$w_{n-2} + \frac{w_{n-2}}{n-1} + \frac{w_{n-4}}{(n-2)(n-3)} + \delta_{n-2} = 0. \tag{8}$$

Solving recursively gives

$$W^{(2)}(\lambda) = -\frac{1}{2} + \frac{3}{16}\lambda^2 - \frac{5}{384}\lambda^4 + \cdots. \tag{9}$$

Integrating twice with respect to λ and inserting the initial conditions give

$$W(\lambda) = 1 - \frac{1}{4}\lambda^2 + \frac{1}{64}\lambda^4 - \frac{1}{2304}\lambda^6 + \cdots = J_0(\lambda) \tag{10}$$

where $J_0(\lambda)$ denotes the Bessel function of the first kind of order 0 and argument λ. Eq. (10) corresponds to $y(t)=J_0(t)$, which is one of the two solutions of (6).

Example 3

Given the Bessel equation

$$\left[t^2\frac{d^2}{dt^2} + t\frac{d}{dt} + t^2\right]y = b^2 y. \tag{11}$$

TABLE I

TAYLOR-CAUCHY TRANSFORM PAIRS FOR VARYING PARAMETER SYSTEMS*

$\lambda^m W^{(k)}(\lambda)$	w_{n-m}
$\lambda^m W^{(k-1)}(\lambda)$	$\frac{w_{n-m-1}}{n-m} + A_{k-1}\delta_{n-m}$
$\lambda^m W^{(k-r)}(\lambda)$	$\frac{w_{n-m-r}}{(n-m)(n-m-1)\cdots(n-m-r+1)} + \sum_{q=1}^{r}\frac{A_{k-q}}{(r-q)!}\delta_{n-m-r+q}$
$e^{\alpha\lambda}W^{(k)}(\lambda)$	$\sum_{m=0}^{n}\frac{(\alpha)^m}{m!}w_{n-m}$
$e^{\alpha\lambda}W^{(k-1)}(\lambda)$	$\sum_{m=0}^{n-1}\frac{(\alpha)^m}{m!}\left[\frac{w_{n-m-1}}{n-m}\right] + A_{k-1}\frac{(\alpha)^n}{n!}$
$e^{\alpha\lambda}W^{(k-r)}(\lambda)$	$\sum_{m=0}^{n-r}\frac{(\alpha)^m}{m!}\left[\frac{w_{n-m-r}}{(n-m)(n-m-1)\cdots(n-m-r+1)}\right] + \sum_{m=0}^{n-r+q}\frac{(\alpha)^m}{m!}\sum_{q=1}^{r}\frac{A_{k-q}}{(r-q)!}\delta_{n-m-r+q}$
$\lambda^m\left[W^{(k)}(\lambda)\right]^2$	$\sum_{\beta=0}^{n-m}w_\beta w_{n-m-\beta}$
$\lambda^m\left[W^{(k-1)}(\lambda)\right]^2$	$\sum_{\beta=0}^{n-m-2}\frac{w_\beta}{\beta+1}\left[\frac{w_{n-m-\beta-2}}{n-m-\beta-1}\right] + 2A_{k-1}\frac{w_{n-m-1}}{n-m} + A_{k-1}^2\delta_{n-m}$
$\lambda^m\left[W^{(k-r)}(\lambda)\right]^2$	$C^{(2)}_{n-m-2r,k-r} + \sum_{q=1}^{r}\binom{2}{1}C^{(1)}_{n-m-2r+q}A^{(1)}_{r-q} + \sum_{q=1}^{r}A^{(2)}_{r-q}\delta_{n-m-r+q}$
$\lambda^m\left[W^{(k-r)}(\lambda)\right]^\nu$	$C^{(\nu)}_{n-m-\nu r,k-r} + \sum_{i=1}^{\nu-1}\sum_{q=1}^{r}\binom{\nu}{i}C^{(\nu-i)}_{n-m-\mu r+q}A^{(i)}_{r-q} + \sum_{q=1}^{r}A^{(\nu)}_{r-q}\delta_{n-m-r+q}$ $(\mu = \nu - i + 1)$
$e^{\alpha\lambda}\left[W^{(k)}(\lambda)\right]^2$	$\sum_{m=0}^{n-\beta}\frac{(\alpha)^m}{m!}\sum_{\beta=0}^{n-m}w_\beta w_{n-m-\beta}$ (β may be omitted in the upper limit of the first summation.)
$e^{\alpha\lambda}\left[W^{(k-1)}(\lambda)\right]^2$	$\sum_{m=0}^{n-\beta-2}\frac{(\alpha)^m}{m!}\sum_{\beta=0}^{n-m-2}\frac{w_\beta}{\beta+1}\left[\frac{w_{n-m-\beta-2}}{n-m-\beta-1}\right] + 2A_{k-1}\sum_{m=0}^{n-1}\frac{(\alpha)^m}{m!}\left[\frac{w_{n-m-1}}{n-m}\right] + A_{k-1}^2\frac{(\alpha)^n}{n!}$
$e^{\alpha\lambda}\left[W^{(k-2)}(\lambda)\right]^2$	$\sum_{m=0}^{n-\beta-4}\frac{(\alpha)^m}{m!}\sum_{\beta=0}^{n-m-4}\frac{w_\beta}{(\beta+1)(\beta+2)}\left[\frac{w_{n-m-\beta-4}}{(n-m-\beta-3)(n-m-\beta-2)}\right] + 2A_{k-1}\sum_{m=0}^{n-3}\frac{(\alpha)^m}{m!}\left[\frac{w_{n-m-3}}{(n-m-2)(n-m-1)}\right] + 2A_{k-2}\sum_{m=0}^{n-2}\frac{(\alpha)^m}{m!}\left[\frac{w_{n-m-2}}{(n-m)(n-m-1)}\right] + A_{k-1}^2\sum_{m=0}^{n-2}\frac{(\alpha)^m}{m!}\delta_{n-m-2} + 2A_{k-1}A_{k-2}\sum_{m=0}^{n-2}\frac{(\alpha)^m}{m!}\delta_{n-m-1} + A_{k-2}^2\frac{(\alpha)^n}{n!}$
$e^{\alpha\lambda}\left[W^{(k-r)}(\lambda)\right]^2$	$\sum_{m=0}^{n-2r}\frac{(\alpha)^m}{m!}C^{(2)}_{n-m-2r,k-r} + \sum_{m=0}^{n-2r+q}\frac{(\alpha)^m}{m!}\sum_{q=1}^{r}\binom{2}{1}C^{(1)}_{n-m-2r+q}A^{(1)}_{r-q} + \sum_{m=0}^{n-r+q}\frac{(\alpha)^m}{m!}\sum_{q=1}^{r}A^{(2)}_{r-q}\delta_{n-m-r+q}$
$e^{\alpha\lambda}\left[W^{(k-r)}(\lambda)\right]^\nu$	$\sum_{m=0}^{n-\nu r}\frac{(\alpha)^m}{m!}C^{(\nu)}_{n-m-\nu r,k-r} + \sum_{m=0}^{n-\mu r+q}\frac{(\alpha)^m}{m!}\sum_{i=0}^{\nu-1}\sum_{q=1}^{r}\binom{\nu}{i}C^{(\nu-i)}_{n-m-\mu r+q}A^{(i)}_{r-q} + \sum_{m=0}^{n-r+q}\frac{(\alpha)^m}{m!}\sum_{q=1}^{r}A^{(\nu)}_{r-q}\delta_{n-m-r+q}$ $(\mu = \nu - i + 1)$

* The symbols $C_{n-\nu r}$ and $A^{(i)}_{r-q}$ are defined in footnote 1.

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[1] Y. H. Ku, A. A. Wolf, and J. H. Dietz, "Taylor-Cauchy transforms for analysis of a class of nonlinear systems," PROC. IRE, vol. 48, pp. 912–922; May, 1960.

Eq. (11) is rewritten as

$$\lambda^2 W^{(2)}(\lambda) + \lambda W^{(1)}(\lambda) + \lambda^2 W(\lambda) = b^2 W(\lambda). \quad (12)$$

Let $b=+1$. Taking the Taylor-Cauchy transform gives

$$w_{n-2} + \frac{w_{n-2}}{n-1} + A_1\delta_{n-1} + \frac{u_{n-4}}{(n-2)(n-3)} + A_1\delta_{n-3} + A_0\delta_{n-2} = \frac{w_{n-2}}{n(n-1)} + A_1\delta_{n-1} + A_0\delta_n. \quad (13)$$

The initial conditions are: $t=0$, $y=0$ and $y'=\frac{1}{2}$. Hence, $A_0=0$ and $A_1=\frac{1}{2}$. Substituting the values of A_0 and A_1 in (13) and solving recursively give

$$w_0 = 0, \quad w_1 = -\frac{3}{8}, \quad w_2 = 0 \quad w_3 = \frac{5}{96}, \quad w_4 = 0, \quad w_5 = -\frac{7}{3072}, \cdots. \quad (14)$$

Therefore,

$$W^{(2)}(\lambda) = \sum_{n=0}^{\infty} w_n\lambda^n = \sum_{r=0}^{\infty} \frac{(-1)^r(2r)(2r+1)\lambda^{(2r-1)}}{r!2^{(2r+1)}(r+1)!}. \quad (15)$$

Integrating twice with respect to λ and inserting the initial conditions give

$$W(\lambda) = \sum_{r=0}^{\infty} \frac{(-1)^r(\lambda/2)^{(2r+1)}}{r!(r+1)!} = J_1(\lambda), \quad (16)$$

where $J_1(\lambda)$ is the Bessel function of the first kind of order 1 and argument λ. The corresponding expression is $y(t)=J_1(t)$. The other solution is given by $Y_1(t)$, which is the Bessel function of the second kind.

Example 4

Given

$$y'' + (2 - e^{-t})y' + y = e^{-2t}. \quad (17)$$

The initial conditions are: $t=0$, $y=1$ and $y'=-1$. Eq. (17) is rewritten as

$$W^{(2)}(\lambda) + 2W^{(1)}(\lambda) + W(\lambda) = e^{-2\lambda} + e^{-\lambda}W^{(1)}(\lambda). \quad (18)$$

Taking the Taylor-Cauchy transform of (18) gives (see Table I):

$$w_n + \frac{2w_{n-1}}{n} - 2\delta_n + \frac{w_{n-2}}{n(n-1)} - \delta_{n-1} + \delta_n = \frac{(-2)^n}{n!} - \frac{(-1)^n}{n!} + \sum_{m=0}^{n-1} \frac{(-1)^m}{m!}\left[\frac{w_{n-m-1}}{n-m}\right] \quad (19)$$

Solving recursively gives

$$w_0 = 1, \; w_1 = -1, \; w_2 = \frac{1}{2}, \; w_3 = -\frac{1}{3!}, \; w_4 = \frac{1}{4!}, \; w_5 = -\frac{1}{5!}, \cdots. \quad (20)$$

Therefore,

$$W^{(2)}(\lambda) = e^{-\lambda}, \quad W^{(1)}(\lambda) = -e^{-\lambda}, \quad W(\lambda) = e^{-\lambda}. \quad (21)$$

This gives $y(t)=e^{-t}$.

Example 5

Given the nonlinear varying-parameter equation

$$e^{-t}y'' + (1 + 2e^{-t})y' + y + y^2 = 0, \quad (22)$$

where $y''=d^2y/dt^2$ and $y'=dy/dt$. The initial conditions are: $t=0$, $y=1$, and $y'=-1$. Eq. (22) is rewritten as

$$e^{-\lambda}W^{(2)}(\lambda) + W^{(1)}(\lambda) + 2e^{-\lambda}W^{(1)}(\lambda) + W(\lambda) + [W(\lambda)]^2 = 0. \quad (23)$$

Taking the Taylor-Cauchy transform of (23) gives

$$\sum_{m=0}^{n} \frac{(-1)^m}{m!} w_{n-m} + \frac{w_{n-1}}{n} + A_1\delta_n + 2\sum_{m=0}^{n-1} \frac{(-1)^m}{m!}\left[\frac{w_{n-m-1}}{n-m}\right] + 2A_1\frac{(-1)^n}{n!} + \frac{w_{n-2}}{n(n-1)} + A_1\delta_{n-1} + A_0\delta_n + \sum_{k=0}^{n-4} \frac{w_k}{(k+1)(k+2)}\left[\frac{w_{n-k-4}}{(n-k-3)(n-k-2)}\right] + 2A_1\frac{w_{n-3}}{(n-1)(n-2)} + 2A_0\frac{w_{n-2}}{n(n-1)} + A_1^2\delta_{n-2} + 2A_0A_1\delta_{n-1} + A_0^2\delta_n = 0, \quad (24)$$

where $A_0=1$ and $A_1=-1$. Solving recursively gives

$$W^{(2)}(\lambda) = e^{-\lambda}, \quad W^{(1)}(\lambda) = -e^{-\lambda}, \quad W(\lambda) = e^{-\lambda}. \quad (25)$$

This corresponds to $y(t)=e^{-t}$.

Example 6

Given the nonlinear varying-parameter equation

$$y'' + (1 - e^{-t}y')y' + e^{-2t}y = 0. \quad (26)$$

The initial conditions are: $t=0$, $y=1$, and $y'=-1$. Eq. (26) is rewritten as

$$W^{(2)}(\lambda) + W^{(1)}(\lambda) - e^{-\lambda}[W^{(1)}(\lambda)]^2 + e^{-2\lambda}W(\lambda) = 0. \quad (27)$$

Taking the Taylor-Cauchy transform of (27) gives

$$w_n + \frac{w_{n-1}}{n} + A_1\delta_n + \sum_{m=0}^{n-k-2} \frac{(-1)^m}{m!} \sum_{k=0}^{n-m-2} \frac{w_k}{k+1}\left[\frac{w_{n-m-k-2}}{n-m-k-1}\right] - 2A_1\sum_{m=0}^{n-1} \frac{(-1)^m}{m!}\left[\frac{w_{n-m-1}}{n-m}\right] - A_1^2\frac{(-1)^n}{n!} + \sum_{m=0}^{n-2} \frac{(-2)^m}{m!}\left[\frac{w_{n-m-2}}{(n-m)(n-m-1)}\right] + \sum_{m=0}^{n-1} \frac{(-2)^m}{m!} A_1\delta_{n-m-1} + A_0\frac{(-2)^m}{n!} = 0, \quad (28)$$

where $A_0=1$ and $A_1=-1$. Solving recursively gives

$$w_n = \frac{(-1)^n}{n!}. \quad (29)$$

This gives $W^{(2)}(\lambda)=e^{-\lambda}$. The corresponding expression is $y''=e^{-t}$. Integrating twice with respect to t and noting the initial conditions give $y=e^{-t}$.

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Correction to "Laurent-Cauchy Transforms for Analysis of Linear Systems Described by Differential-Difference and Sum Equations"*

In the above paper,[1] the authors wish to make the following corrections.

Eq. (78) should be

$$\sum_{n=0}^{k} n = \frac{1}{2!}\left[\frac{d^2}{d\rho^2}(\rho^{k+1} - 1)\right]_{\rho=1} \quad (78)$$

In Table I, Section 1, Operational Forms, item 18 should be

$$\sum_{k=0}^{n} kh_k g_{n-k}.$$

In item 26, the three primes should be replaced by three dots.

In Table I, Section 2, Functional Forms, item 10, the Laurent-Cauchy transform of $(-1/n)$, should be

$$\log \frac{\rho - 1}{\rho}.$$

In item 20, the transform of $(-n^3h_n)$ should be

$$\rho^3H'''(\rho) + 3\rho^2H''(\rho) + \rho H'(\rho).$$

In item 21, the denominator of the transform should be

$$\rho^2 - 2e^{-\alpha}\rho\cos a + e^{-2\alpha}.$$

We can add a similar pair:

$$e^{-\alpha n}\cos an \qquad \frac{\rho(\rho - e^{-\alpha}\cos a)}{\rho^2 - 2e^{-\alpha}\rho\cos a + e^{-2\alpha}}.$$

Above item 6, we may add the pair:

$$a^n \qquad \frac{\rho}{\rho - a}.$$

In item 24, we should define

$$n^{[k]} = n(n+1)\cdots(n+k-1).$$

The authors would like to thank Prof. E. I. Jury of the University of California for suggesting the corrections in item 18 of operational forms and items 10 and 20 of functional forms.[2]

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[1] Y. H. Ku and A. A. Wolf, Proc. IRE, vol. 48, pp. 923–931; May, 1960.

[2] Y. H. Ku and A. A. Wolf, "Laurent-Cauchy transform for analysis of linear systems described by differential-difference and sum equations," Proc. IRE (Correspondence), vol. 48, pp. 2026–2027; December, 1960.

Some Recent Papers in Threshold Logic*

Recent papers in the area of threshold logic point up a great need for some sort of drawing together of the various workers not only in the matter of basic notation, but also in the area of what *facts* are known about the subject. There are already available the many papers listed below, and the author knows of at least half a dozen other independent studies in threshold logic so far unpublished. The following remarks are intended to bring these papers to the attention of interested readers and perhaps to suggest some order in this fast developing field. No attempt at comparison or relative evaluation is made, but emphasis certainly is affected by the author's bias and state of information.

The basic ideas of single-element threshold logic have been discovered independently by many people. These ideas include, first, a number of simple properties,[1] the most important of which is that the realizability of a given function is equivalent to the consistency of a certain set of linear inequalities—the solutions of this set give all realizing weightings. Going beyond these simple properties, most workers evolve at least parts of a doubly infinite chain of conditions necessary for realizability. Since no one else has suggested names for these properties, we will use the author's terms. The first chain consists of *1-monotonicity, 2-monotonicity*, and *k-monotonicity* in general.[2] Each of these conditions entails its predecessors, and is stricter than them; the union of all these conditions is called *complete monotonicity*.[2] Complete monotonicity in turn is equivalent to *2-acyclicity*, and then *3-acyclicity, 4-acyclicity*, and *k-acyclicity* in general[3] form successively stricter conditions yet. All of these conditions, together with their union—*complete acyclicity*, are easily seen to be necessary for realizability. Whether complete acyclicity is sufficient for realizability remains an open question. The importance of these conditions in testing for realizability, completing partially specified functions, and deriving compound realizations is discussed in the References.

Now we shall trace these and other ideas in the various studies available. Around 1956–1957, several men at Bell Telephone Laboratories were interested in magnetic core switching, and were familiar with many of the basic properties of threshold logic. Unfortunately, their work was not published then and is only recently coming to light. The Paull and McCluskey paper [10], for instance, establishes the necessity of 1-monotonicity (unateness), 2-monotonicity, and complete monotonicity. A function devised by E. F. Moore, now widely known, establishes that complete monotonicity is not sufficient for realizability (see [12]). The report of McNaughton [6] derives from the Bell Labs. body of knowledge and discusses 1-monotonicity and the equivalence of realizability to the consistency of a set of linear equalities.

An independent effort was generated by Minnick [7], who applies the theory of linear programming to obtain solutions of this system of inequalities. He takes a step in the important direction of *compound* realization (realization using more than one threshold element) by treating the "residue" of an unrealizable function as a new function to be realized, and, after a series of such steps, producing a collection of threshold devices whose outputs are then "or-ed" together to realize the original function. Stram [11] streamlines this procedure by the use of 2-monotonicity (equivalent to the passing of all "sieve" tests). His suggestion that 2-monotonicity is sufficient for realizability (consistency) is erroneous.

The independently derived papers of Muroga, *et al.*, Elgot, and the author give extensive treatments of the various simple properties already discussed, and further consider: chains of functions realized by fixed weights, varying threshold [2], [9]; bounds on the number or realizable functions of n arguments [9], [12]; an "enumeration lemma" which facilitates the listing of all 5-argument realizable functions [9], [12] and has been used by Muroga, *et al.*, and by the author, to list all 6-argument realizable functions. These lists demonstrate that a 4- or 5-argument function is realizable if and only if it is 2-monotonic [2], a 6-argument function if and only if 3-monotonic. Muroga, *et al.*, [9], also discuss the linear programming approach to synthesis, relate a completely monotonic function and its dual, study the "convolution" of two functions obtained by adding weights termwise, and describe several families of realizable functions. Elgot [2] identifies complete monotonicity and 2-acyclicity, shows the necessity of complete acyclicity for realizability, further analyzes 3-acyclicity, characterizes realizable functions of one, two, three, and four prime implicants, discusses the system of linear inequalities which characterizes a function's realizability, and gives an equivalent characterization in terms of a classical property of such systems. (Chow [1] also states this characterization, and provides a short and elegant treatment of a condition closely related to k-acyclicity—also discussed by Elgot.) The emphasis in the writer's paper [12] is on a facilitation of the algebraic test-synthesis procedure. Aids in the checking of complete monotonicity are given (for instance, it's shown that at most $n-1$ sieves are needed, instead of Stram's $n(n-1)/2$, to check 2-monotonicity); and a process of deleting redundant inequalities from the original set of 2^n characterizing realizability is studied in detail.

Two further efforts in threshold logic, somewhat out of the general stream of the above discussed work, should be mentioned. Lindaman's treatment of compound design using three-argument majority elements [3], [4], and Mattson's experiments with self-organizing, many-input threshold devices [5]. There is much other work in the area of self-organizing threshold devices, at present little involved in the specific properties of threshold logic; yet, such efforts will very likely eventually contribute to the growing theory of threshold logic.

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(For papers with incidental importance to threshold logic, see the references of these papers.)

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* Received by the IRE, February 17, 1961; revised manuscript received, March 3, 1961.

[1] [2], [9], and [12] provide the most detailed discussions.

[2] These ideas are defined in these terms in [12].

[3] In Elgot's notation [2], f is k-acyclic when there is no cycle of length $r \leq k$:

$$X_1 \xrightarrow{f} X_2 \xrightarrow{f} \cdots \xrightarrow{f} X_r \xrightarrow{f} X_1.$$

WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table. The corrections reported have been improved by means of improved measurement methods based on LF and VLF transmissions.

The characteristics of the USFS, and its relation to time scales such as ET and UT2,

* Received by the IRE, April 14, 1961.

have been described in a previous issue,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. Also, they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960. A retardation time adjustment of 20 msec was made on December 16, 1959; another retardation adjustment of 5 msec was made at 0000 UT on January 1, 1961.

WWV FREQUENCY WITH RESPECT TO U. S. FREQUENCY STANDARD

1961 March	Parts in 10^{10}†
1	−150.2
2	−150.3
3	−150.7
4	−150.6
5	−150.8
6	−150.5
7	−150.2
8	−150.0
9	−150.0
10	−150.1
11	−150.3
12	−150.5
13	−150.7
14	−150.4
15	−149.8
16	−149.8
17	−149.5
18	−149.4
19	−149.3
20	−149.6
21	−149.6
22	−149.5
23	−149.4
24	−149.4
26	−149.2
26	−148.9
27	−148.7
28	−148.6
29	−148.8
30‡	−149.1
31	−150.3

† A minus sign indicates that the broadcast frequency was low. The uncertainty associated with these values is $\pm 5 \times 10^{-11}$.
‡ The frequency was decreased $10 \pm 3 \times 10^{-11}$ on March 30, 1961.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

[1] "National Standards of Time and Frequency in the United States," PROC. IRE, vol. 48, pp. 105–106; January, 1960.

Gallium Antimonide Esaki Diodes for High-Frequency Applications*

Gallium antimonide (GaSb) has not aroused wide interest as a semiconductor for diode fabrication, probably because its energy gap (0.7 ev) closely approximates that of germanium. However, Esaki diode voltage-current characteristics available in the literature[1] suggested that this material should make Esaki diodes with relatively low noise potentialities for negative resistance amplifier applications, and therefore that it warranted investigation as a possible semiconductor for the fabrication of high-frequency Esaki diodes.

Such diodes, capab e of operation into the millimeter wave region, have now been made from both p- and n-type GaSb. The techniques and processes differed only in detail from those previously employed in the fabrication of gallium arsenide (GaAs) Esaki diodes for use at very high frequencies.[2]

Briefly, small-area alloyed junctions on p-type GaSb were produced by electrically "forming" a light contact between the GaSb surface and a metal alloy point containing an arbitrarily large percentage of tellurium or selenium. Heavily doped polycrystalline p-type GaSb was obtained by remelting p-type material from commercial sources[3] with any of several acceptors (germanium or zinc, for example). A back contact was formed most simply by soldering directly to the semiconductor. The junction-forming operation consisted of a low voltage current pulse in the forward-diode direction. The exact composition of the alloy, not critical, was dictated by the limiting amount of tellurium or selenium which could be incorporated into tin, lead or other convenient carrier without rendering the alloy too brittle to handle.

The resulting Esaki diodes, using p-type GaSb with a resistivity of about 0.0011 ohm-cm, normally exhibited a peak current at 0.07–0.09 volt with the valley at 0.2–0.3 volt. The forward current returned to the peak-current value at 0.45–0.6 volt. Peak currents ranged from several microamperes to a few tens of milliamperes, depending upon the exact forming conditions, with peak-to-valley ratios commonly 6:1 to 12:1, but frequently exceeding 15:1. Current densities well in excess of 50,000 a/cm², estimated from a microscopic examination of the apparent junction areas, have been obtained. Fig. 1 (a)–(c) shows three rather typical 60-cps forward characteristics for diodes of p-type GaSb with formed Sn-Te alloy point contacts. The different curves were obtained by variations in the forming operation.

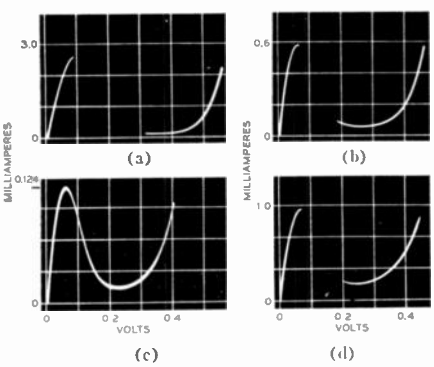

Fig. 1. Oscilloscope tracings of forward 60-cps V-I characteristics of high-frequency GaSb Esaki diodes: (a)–(c) diodes of p-type GaSb with formed point contacts of a Sn-Te alloy; (d) diode of n-type GaSb with formed point contact of zinc.

Use of slightly higher semiconductor resistivities yielded diodes with peak and valley voltages reduced to about 0.05 and 0.17 volt minimum, respectively, but at the expense of reduced current densities and therefore lower expected high-frequency cutoff frequencies. Conversely, use of lower-resistivity material shifted these voltages to considerably higher values and increased the current dens ties.

Diodes have been made in an analogous way from n-type GaSb doped with tellurium or selenium. The maximum peak-to-valley ratios were considerably lower than those of comparable p-type units, and the peak and valley points in many cases tended to shift to somewhat higher voltages. Zinc, cadmium or copper was employed as the point material. Fig. 1 (d) shows the 60-cps characteristic of an n-type diode with a formed point contact of zinc.

A number of authors[4] have pointed out that a useful noise figure of merit for Esaki diodes is the quantity I_0R_n, where I_0 is the diode current at the operating point and R_n is the diode negative resistance evaluated at I_0. The product should be small for low noise amplification. High-frequency diodes of moderately doped p-type GaSb have exhibited room temperature I_0R_n values as low as 0.043 volt. This value compares favorably with that previously reported[5] for lower-frequency germanium (0.06 volt) and gallium arsenide (0.12 volt) Esaki diodes. The I_0R_n values for Ge and GaAs diodes with demonstrated higher-frequency capabilities have been somewhat larger[6] than those listed by Dacey.[5] We feel, therefore, that when compared to materials previously used in fabricating Esaki diodes, GaSb Esaki diodes can provide at least some noise advantage in high-frequency microwave amplifiers operated at room temperature. This advantage also should be realized in diodes fabricated for lower-frequency applications.

Fundamental oscillations to frequencies in excess of 50 kMc (6 millimeters wavelength) have been obtained with 0.3–0.6 ma p-type GaSb diodes in the simple waveguide circuits previously described.[2] Somewhat higher peak-current units of n-type GaSb oscillating in the same circuits have provided fundamental power to 62.5 kMc. The microwave output in all cases was readily detectable with the simplest video receiver.

We should like to thank R. F. Trambarulo for helpful suggestions and discussions concerning this work.

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* Received by the IRE, April 20, 1961.

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[5] G. C. Dacey, "Materials Requirements for Esaki Diodes," presented at the AIME Symp. on Elemental and Compound Semiconductors, Boston, Mass.; August 29, 1960.
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Contributors

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A. A. ASHKIN

From 1942 to 1946 he worked at the Columbia Radiation Laboratory on microwave magnetrons. From 1945 to 1946 he served in the U. S. Army Signal Corps. At Cornell in the Laboratory of Nuclear Studies, he performed experiments on high-energy gamma-ray absorption and high-energy electron and positron scattering. In 1952 he joined the Bell Telephone Laboratories, Murray Hill, N. J., where he has been working in the field of microwave electronics research, studying electron-beam dynamics, electron-beam noise, and electron-beam parametric amplifiers.

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❖

Bruce B. Barrow, for a photograph and biography, please see page 1655 of the September, 1960, issue of PROCEEDINGS.

❖

Charles J. Bickart was born in Union City, N. J., on September 29, 1923. He attended signal school while serving with the U. S. Navy during World War II. Upon completion of his term of duty he was employed by Bendix Aviation Corporation, Teterboro, N. J., in the field of instrument design and fabrication.

C. J. BICKART

In 1951 he joined the USASRDL, Fort Monmouth, N. J., and participated at first in extensive experiments in the fluid dynamic field as applied to servo-mechanisms. In 1957 he transferred to the Atomic Resonance Branch of USASRDL, and has worked ever since on NH_3 Masers and the application of C_s beam standards to high-precision clock synchronization.

❖

G. Dermit (M'54) was born in Istanbul, Turkey, on February 9, 1925. He received the B.S.E.E. degree from Robert College, Istanbul, Turkey, and the M.E.E. degree from Cornell University, Ithaca, N. Y., in 1947 and 1949, respectively. He continued work at The Ohio State University, Columbus, from 1949 to 1951 in physics, and presently is completing his thesis for the Ph.D. degree in physics at the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. His dissertation is on theoretical work in crystal binding.

G. DERMIT

From 1952 to 1956 he was employed in the field of receiver design and development by Sylvania Electric Company, Buffalo, N. Y., and in the field of transistor circuit applications by Radio Receptor Company, Brooklyn, N. Y., and by Link Aviation, Inc., Binghamton, N. Y. In 1956 he joined the General Transistor Company, Jamaica, N. Y., where he carried out and directed semiconductor device work including the drift transistor, four-layer switch, high-voltage transistor and other device work. In 1959 he joined General Telephone and Electronics Laboratories, Inc., Bayside, N. Y., where he is presently Head of the Advanced Device Research Section in the Solid-State Laboratory.

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❖

W. A. Edson (M'41–SM'43–F'37), for a photograph and biography, please see page 1505 of the August, 1960, issue of PROCEEDINGS.

❖

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A. B. FONTAINE

He worked as a research assistant from 1950 to 1953, and as a research fellow from 1953 to 1954 at the Ohio State Research Foundation. From 1954 to 1959 he was employed by the IBM Product Development and Research Laboratories, Poughkeepsie, N. Y., where he worked on the design and development of the IBM 705 computer and in the Information Research Development. In 1959 he joined the faculty of the University of Wisconsin, where he is now an associate professor.

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Robert G. Gallager was born in Philadelphia, Pa., on May 29, 1931. He received the B.S. degree from the University of Pennsylvania, Philadelphia, in 1953.

R. G. GALLAGER

From 1953 to 1954 he was employed by the Bell Telephone Laboratories, and from 1954 to 1956 he served with the U. S. Army. From 1956 to 1960 he was associated with the Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, receiving the M.S. and D.Sc. degrees in 1957 and 1960, respectively. Since September, 1960, he has been an assistant professor at M.I.T., working in the field of information theory.

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Robert C. Hansen (S'47–A'49–M'55–SM'56) was born in St. Louis, Mo., on August 3, 1926. He received the B.S.E.E. degree from the Missouri School of Mines and Metallurgy in 1949, where he was an honor student, and the M.S. and Ph.D. degrees from the University of Illinois, Urbana, in 1950 and 1955, respectively.

R. C. HANSEN

In 1945 and 1946 he was in the U. S. Navy Service as a radio technician and electronics instructor. At the University of Illinois he worked in the Antenna Laboratory on ferrite loops, streamlined airborne antennas, direction finding and homing systems. Subsequently he was Senior Staff Engineer in the Microwave Laboratory of Hughes Aircraft Company, Culver City, Calif., working on surface-wave antennas, slot arrays, near fields and radio power transfer, electronic scanning and steerable arrays, and dynamic antennas. In 1960 he became a Senior Staff Engineer in the Telecommunications Laboratory of Space Technology Laboratories, Inc., engaged in communications satellite, telemetry, tracking, and command studies. After the formation of Aerospace Corporation and transfer of projects from STL, he transferred to Aerospace Corporation, Los Angeles, Calif., where he is Senior Staff Engineer in the Electronics Laboratory.

He is Vice Chairman of the Los Angeles Section of the IRE, Chairman of the Los Angeles Chapter of PGMTT, and Associate Editor of the *Microwave Journal*.

Dr. Hansen is a member of the American Physical Society, Tau Beta Pi, Sigma Xi, Eta Kappa Nu, Phi Kappa Phi, Commission VI of URSI, and is a registered Professional Engineer in Missouri.

❖

Ernest S. Kuh (S'49–A'52–M'57) was born in China on October 2, 1928. He received the B.S.E.E. degree from the University of Michigan, Ann Arbor, in 1949, the M.S. degree from Massachusetts Institute of Technology, Cambridge, in 1950, and the Ph.D. degree from Stanford University, Stanford, Calif., in 1952.

E. S. Kuh

He then joined Bell Telephone Laboratories, Murray Hill, N. J., where he worked in the transmission systems group, principally in the field of network theory. In 1956 he joined the Electrical Engineering Department of the University of California, Berkeley, where he is now an Associate Professor.

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❖

Maurice W. Long (S'47–A'51–SM'55) was born on April 20, 1925, in Madisonville, Ky. He received the B.E.E. degree from the Georgia Institute of Technology, Atlanta, in 1946, and the M.E.E. degree from the University of Kentucky, Lexington, in 1948, did a year of graduate work at Columbia University, New York, N. Y., and received the Ph.D. degree in physics from the Georgia Institute of Technology in 1959.

M. W. Long

After spending 1943 to 1946 in the U. S. Navy, he was employed as a research assistant for parts of 1946, 1947, and 1948 at the Georgia Institute of Technology. From 1947 to 1949 he was an Instructor of electrical engineering at the University of Kentucky. Since 1950 he has been at the Engineering Experiment Station of the Georgia Institute of Technology, where he is presently Chief of the Electronics Division.

❖

John D. Patterson was born in Santa Rosa, Calif., on April 24, 1933. He received the B.S.E.E. degree in 1959, and the M.S. degree in 1960, both from the University of California, Berkeley, where he is presently working toward the Ph.D. degree.

J. D. Patterson

He has been employed part time by E. A. Hosmer and Company, where he was engaged in development and construction of telephone and carrier systems, and by the Applied Radiation Corporation, where he worked on electron linear accelerators.

Mr. Patterson received the National Science Foundation Cooperative Graduate Fellowships for 1959–1960 and 1960–1961. He is a member of Tau Beta Pi, Sigma Xi, and Phi Beta Kappa.

❖

Friedrich H. Reder (A'54) was born in Garsten, Austria, on December 9, 1919. He received the Cand. Ing. degree in technical physics from the Institute of Technology, Graz, Austria, in 1947, and the Ph.D. degree in physics from the University of Graz in 1949.

F. H. Reder

From 1948 to 1950 he was scientific assistant at the Physics Institute of the University of Graz, and worked on microwave cavity methods for determining the dielectric constant of solids and liquids, and in interference microscopy. From 1950 to 1951 he held a research fellowship at Massachusetts Institute of Technology, Cambridge, Mass., where he was engaged in studying microwave gas discharge phenomena in pure helium gas. From 1951 to 1953 he worked on gas discharge projects and high-vacuum techniques at the University of Graz and at the Shell Research Laboratory, Amsterdam, The Netherlands. Since 1953 he has been with the Frequency Control Division of the U. S. Army Signal Research and Development Laboratory, Fort Monmouth, N. J., where he is now chief of the Atomic Resonance Branch, and in charge of research and development work in atomic and molecular frequency control.

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

❖

Wayne K. Rivers, Jr., was born in Atlanta, Ga., on May 18, 1930. He received the B.S.E.E. degree in 1951, and the M.S. degree in physics in 1958, both from the Georgia Institute of Technology, Atlanta.

From 1951 to 1954 he served with the U. S. Navy as degaussing officer and instructor. From 1956 to 1958 he was engaged in the production of quartz crystals with Pan Electronics Corporation, Griffin, Ga. In 1957 he joined the Engineering Experiment Station, Georgia Institute of Technology, where he is presently a research physicist with the Electronics Division.

W. K. Rivers, Jr.

Mr. Rivers is a member of the American Physical Society and Sigma Xi.

❖

Ronald G. Smart was born in Sydney, Australia, on October 26, 1927. He received the B.E.E.E. degree with first class honours and University medal from the University of New South Wales, Sydney, Australia, in 1954.

R. G. Smart

After one year in telecommunications with the Australian Post Office, he spent a year in England studying digital computers for the University of New South Wales. In 1956 he set up the UTECOM digital computing laboratory in the School of Electrical Engineering and since then has been in charge of the laboratory, providing a teaching and computing service for Eastern Australia. Since 1953 he has been working on varying reactance devices, making extensive use of digital computation. Since April, 1961, he has been Director of Computing Services for Remington Rand Chartres in Australia.

Mr. Smart is an associate member of the Institution of Engineers, Australia, and a member of the British Computer Society, Ltd. and the Statistical Society of New South Wales.

❖

Gernot M. R. Winkler (M'60) was born in Bruck, Austria, on October 17, 1922. He received the Ph.D. degree from the University of Graz, Austria, in 1952.

G. M. R. Winkler

He was on the staff of the Astronomical Observatory in Graz, as Assistant to the Director, from 1949 to 1953. He was employed in industry from 1953 to 1956. He came to this country in 1956 and joined the Frequency Control Division, USASRDL, Fort Monmouth, N. J., as a consulting physicist. He has been engaged with various developments in the field of microwave electronics, precision frequency and time measurements. He is now Deputy Director of the Exploratory Research Division C.

Report of the Secretary—1960

To the Board of Directors
The Institute of Radio Engineers, Inc.

Gentlemen:

The Secretary's Report for the year 1960 is transmitted herewith

The continued enlargement of your Institute's size and importance, as particularly evidenced by an 11.8% increase in membership (see Fig. 1 and Tables I and II), and a 16.1% increase in attendance at the International Convention during 1960, clearly indicates the growing importance of the fields of endeavor that IRE represents. A mere perusal of these statistics discloses how rapidly our ramified activities are moving forward geographically, technically, professionally, and in education. Also can be seen how the IRE structure serves the membership.

The acquisition of your new building on Fifth Avenue adjoining the main building at 1 East 79th Street was completed and its renovation well under way at the end of the year so that space will soon be ready for the heavier work load carried by the Headquarters Staff.

Respectfully submitted,

Haraden Pratt

Haraden Pratt
Secretary

January 31, 1961

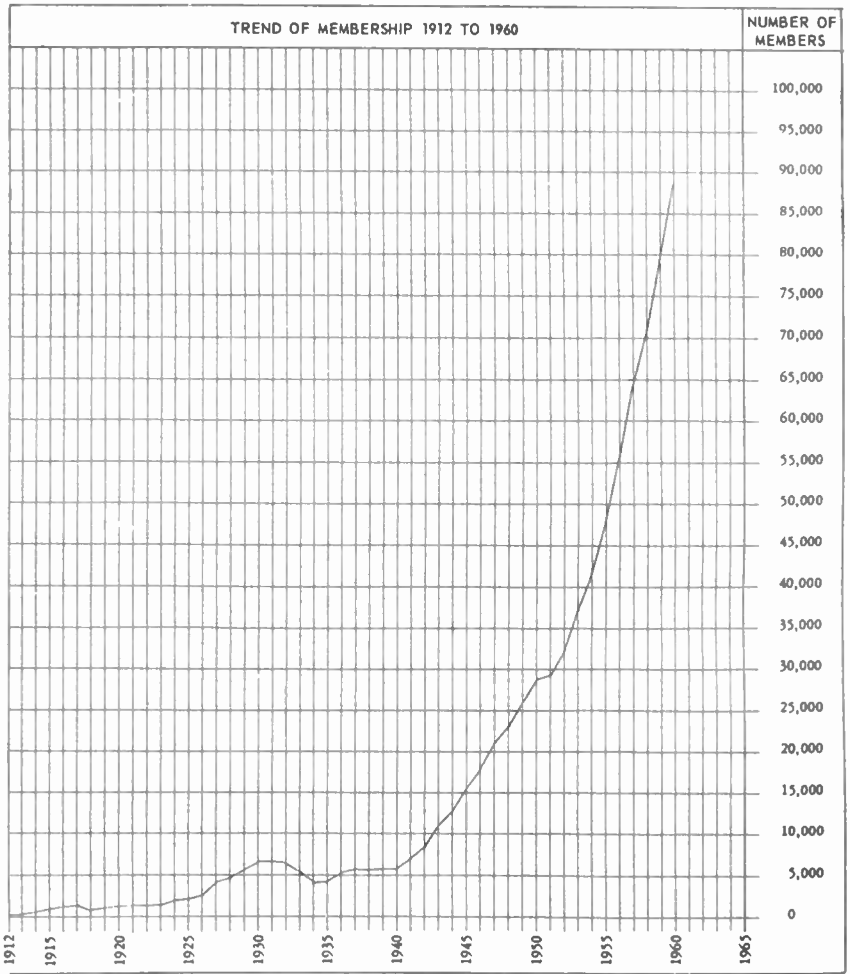

Fig. 1

Fiscal

A condensed summary of income and expenses for 1960 is shown in Table III, and a balance sheet in Table IV (page 1106).

TABLE I

Comparison of Total Membership by Grades, 1958–1960

Grade	As of Dec. 31, 1960		As of Dec. 31, 1959		As of Dec. 31, 1958	
	Number	% of Total	Number	% of Total	Number	% of Total
Fellow	896	1	823	1	770	1.1
Senior Member	10,001	11	9,463	12	8,536	12.0
Member	47,114	53	38,977	49	32,373	45.4
Associate	12,806	15	13,165	17	14,721	20.6
Student	17,662	20	16,738	21	14,961	20.9
Totals	88,479		79,166		71,361	

Editorial Department

The year 1960 saw the IRE publication program continue its steady growth. During the year the IRE published 135 issues totaling 18,610 pages, of which 15,388 pages were devoted to technical and editorial material, a 6% increase over 1959. The rise in publication output underscored the fact that the Professional Group Transactions are growing with continued vigor and now account for more than half the articles published by the IRE.

Proceedings of the IRE

The year was highlighted by the appearance of two special issues of the Proceedings: "Space Electronics" in April and "Report of the Television Allocations Study Organization" in June. The total number of papers published, 174, was less than the previous year's total of 208 due to the heavy

TABLE II

Five-Year Analysis of Membership in U. S. and Other Countries

	1960	1959	1958	1957	1956
Total	88,479	79,166	71,361	64,773	55,494
U. S. and Possessions	81,634	73,044	65,786	59,961	51,551
Other Countries	6,845	6,122	5,575	4,812	3,943
Per Cent Other Countries	7.7	7.7	7.8	7.4	7.1

TABLE V

VOLUME OF PROCEEDINGS PAGES

	1960	1959	1958	1957
Editorial	2325	2370	2199	1868
Advertising	2485	2760	2169	2700
TOTAL	4810	5130	4368	4568

schedule of special issues in 1959. However, the number of contributed papers in regular issues rose substantially in 1960. This increase was accompanied by an expansion of the Correspondence section from 209 letters to 280 letters. The continued growth of the Correspondence section, which has now doubled in size in the last three years, made it necessary for the Editorial Board to place a limit on the length of letters to the editor. The resulting total of PROCEEDINGS papers for 1960 is shown in Table V and in Fig. 2.

The number of papers reviewed for the PROCEEDINGS remained relatively constant: 259 papers totaling 2156 pages. Of these 26% were accepted, 39% were referred to the TRANSACTIONS for publication consideration, and 35% were rejected. An additional 153 papers were reviewed for special issues. Seven IRE Standards also appeared during the year.

TRANSACTIONS

The continued good health of the Professional Groups was evidenced by an 11% increase in TRANSACTIONS output during 1960. As shown in Fig. 3, total pages increased from 7,778 in 1959 to 8,616. The total number of papers and letters published, 1,213, accounted for more than half the total IRE output (2,085). It is noteworthy that the number of TRANSACTIONS on a bi-monthly publication schedule rose from 1 to 3.

IRE CONVENTION RECORDS

The 1960 IRE INTERNATIONAL CONVENTION RECORD, published in 10 parts, contained 240 papers and 18 abstracts totaling 2,208 pages, while the 8-part IRE WESCON CONVENTION RECORD contained 135 papers and 23 abstracts totaling 1,188 pages. The IRE WESCON CONVENTION RECORD was published in time for distribution at WESCON.

IRE STUDENT QUARTERLY

Four issues, totaling 224 pages, were sent free to IRE Student members during the year. In addition, approximately 24,000 free copies of the September issue were distributed to all non-IRE junior and senior electrical engineering students.

IRE DIRECTORY

The 1961 IRE DIRECTORY, which was published in November, 1960, contained 1,432 pages including covers, of which 645 were membership listings and information and 787 were advertisements and listings of manufacturers and products.

NEW PUBLICATIONS

A five-year cumulative index to all IRE publications, covering the years 1954–1958, was issued early in 1960. Meanwhile, work

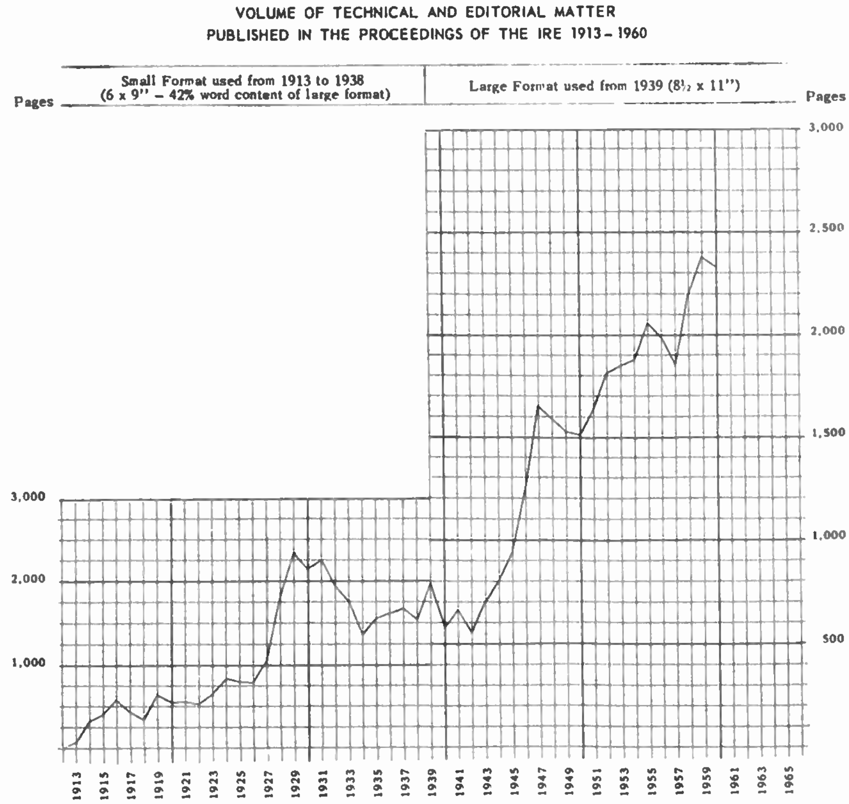

Fig. 2

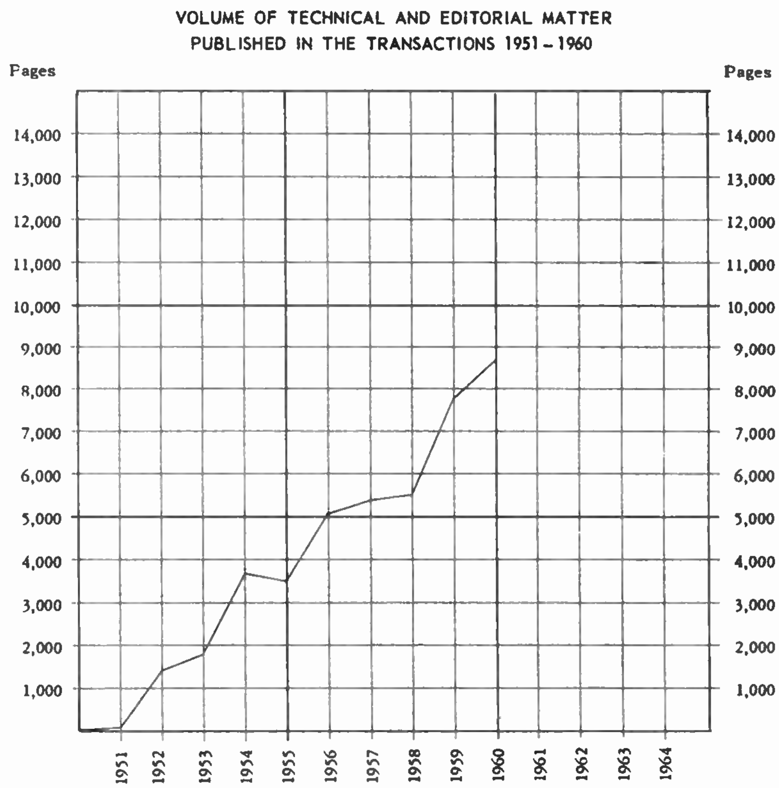

Fig. 3

TABLE III

SUMMARY OF INCOME AND EXPENSE, 1960

Income		
Advertising	$1,710,134	
Members Dues and Convention	1,769,532	
Subscriptions	217,707	
Sales Items, Binders, Emblems, etc.	187,043	
Investments	52,969	
Miscellaneous	1,607	
TOTAL INCOME		$3,938,992
Expense		
Proceedings Editorial Pages	$ 585,458	
Advertising Pages	855,441	
Directory	318,199	
Section Rebates	88,469	
Student Program	142,624	
Professional Groups	237,539	
Sales Items	108,923	
General Operations	587,283	
Convention	545,757	
Publications	117,500	
TOTAL EXPENSE		$3,587,193
Reserve for Future Operations - Gross		$ 351,799
Depreciation		28,033
Reserve for Future Operations—Net		$ 323,766

TABLE IV

BALANCE SHEET, DECEMBER 31, 1960

Assets		
Cash and Accounts Receivable	$1,087,667	
Inventory	20,379	
TOTAL CURRENT ASSETS		$1,108,046
Investments at Cost	1,769,701	
Buildings and Land at Cost	1,261,361	
Furniture and Fixtures at Cost	288,390	
Other Assets	104,609	
TOTAL		$3,424,061
TOTAL ASSETS		$4,532,107
Liabilities and Surplus		
Accounts Payable	$ 94,660	
TOTAL CURRENT LIABILITIES		$ 94,660
Deferred Income	1,282,265	
Professional Groups Funds on Deposit	187,389	
		1,469,654
TOTAL LIABILITIES		$1,564,314
Reserve for Publications	165,000	
Reserve for Depreciation	113,798	
TOTAL RESERVES		278,798
Surplus Donated	595,287	
Surplus	2,093,708	
TOTAL SURPLUS		2,688,995
TOTAL LIABILITIES AND SURPLUS		$4,532,107

on the *IRE Dictionary of Electronics Terms and Symbols*, containing all definitions of terms and graphical and letter symbols which have appeared in IRE Standards over the past 18 years, was completed. The Dictionary was issued in January, 1961.

Technical Activities

During 1960, 26 Technical Committees and their 120 subcommittees and Task Groups held 233 meetings, of which 224 were held at IRE Headquarters and 9 throughout the nation.

Seven IRE Standards, having been approved by the IRE Standards Committee and the Executive Committee, were published in the PROCEEDINGS in 1960. Reprints are now available to the public.

IRE is directly represented on 34 Sectional Committees of the American Standards Association, of which IRE sponsors three: The ASA Sectional Committee on Radio and Electronic Equipment, C16; the ASA Sectional Committee on Sound Recording, S4; and the ASA Sectional Committee on Nuclear Instrumentation, N3. One IRE Standard received approval by the American Standards Association in 1960, and is now available overseas through its international agency.

IRE Technical Committees participated in international standardization in 1960 by reviewing and preparing comments on documents for the United States National Committee of the International Electrotechnical Commission.

The International Electrotechnical Commission (IEC)

The annual meeting of the International Electrotechnical Commission was held in New Delhi, India from October 31 to November 7, 1960. The following meetings, of particular significance to the IRE, took place: Technical Committee 45 on Electrical Measuring Instruments used in connection with Ionizing Radiation; Technical Committee 12 on Radio Communication; and Subcommittee 12-1 on Radio Receiving Equipment. IRE secured the services of two delegates to represent the United States, and the IRE Technical Committees prepared these delegates on all items to be discussed at this meeting.

In addition, IEC Technical Committee 25 on Symbols, met in Paris, France, from April 20 to 22, 1960. At this meeting, the United States was represented by a delegate appointed by the IRE.

Appointed IRE Delegates to Outside Organizations

IRE appointed Delegates to a number of organizations for the one-year period May 1, 1960, to April 30, 1961 (as listed on page 60A of the October, 1960 PROCEEDINGS).

The Joint Annual Spring Meeting of the International Scientific Radio Union (URSI) and IRE was held in Washington, D.C.—May 2–5, 1960. The Fall Meeting was held at the Boulder Laboratories, National Bureau of Standards, Boulder, Colo., December 12–14, 1960. The IRE Professional Groups on Antennas and Propagation; Circuit Theory; Information Theory; Instrumentation; and Microwave Theory and Techniques co-sponsored these meetings. Several IRE Board members attended the XIIIth General Assembly of URSI which was held at the University of London, London, England, from September 5 through 15, 1960.

During 1960, the Executive Committee of the U. S. National CCIR (International Radio Consultative Committee) Organization held three meetings during which the representatives of the fourteen Study Groups summarized and reported on their activities. Lists of all material received at IRE during 1960 were distributed quarterly to the Chairmen of the IRE Technical Committees and Professional Groups, as well as to the members of the Joint Technical Advisory Committee.

The Joint IRE-EIA-SMPTE-NAB (JCIC) Committee for Inter-Society Coordination held four meetings during 1960 to discuss the problems facing intersociety coordination. The scope of JCIC has been expanded.

Armed Forces National Research Council Committee: Panel on Bio-Instrumentation held several meetings of its various Panels during the year. Among these were one in Los Angeles, April 19–20, 1960, and another at Woods Hole, Mass., August 31 to September 2, 1960.

The Joint Technical Advisory Committee (JTAC)

The Joint Technical Advisory Committee held a total of ten meetings for the period July 1, 1959, through June 30, 1960. The Twelfth Anniversary dinner was held in May, 1960 at the University Club, New York.

Volume XVII, the cumulative Annual Report of the JTAC Proceedings was published in September, 1960. Section I of the Report included official correspondence between the Federal Communications Commission and the Joint Technical Advisory Committee (IRE-EIA) and correspondence pertinent to the activities of the JTAC: Section II contained approved Minutes of Meetings of the Joint Technical Advisory Committee from July 1, 1959, through June 30, 1960.

The JTAC ad hoc Subcommittee 60.1 was formed in January, 1960 to study Frequency Diversity problems posed by the FCC. The Committee submitted its report to the FCC in November, 1960.

In July, 1960, the JTAC, recognizing the technical and policy problems concerning the United States, formed an ad hoc Subcommittee on Frequency Allocations for Space Communications to study problems posed by the FCC's inquiry into the allocation of frequency bands for space communication of an industry-wide nature. The Subcommittee held three meetings during 1960.

Professional Group System

General: There are currently 28 Professional Groups operating actively within the IRE. The majority of IRE members have taken advantage of the Professional Group System which now has a total membership of 89,114. Included are 6,729 Student members of the IRE who have joined the Groups at the special Student member rate of $1.00 annually. Under the newly instituted Affiliate Plan, 503 scientists and medical doctors, whose major interests lie in fields other than electronics, have affiliated with a number of the Professional Groups.

All of the Groups have levied publications fees and their members are receiving the pertinent Group TRANSACTIONS regularly. In addition, large a number of company, university and public libraries have subscribed to the TRANSACTIONS of all the Groups. There is also a demand for individual Group subscriptions and individual copies of the TRANSACTIONS from outside sources.

Financial and editorial assistance were among the many services rendered by Headquarters to the Groups during 1960. The Office of the Technical Secretary provided administrative services for Group operations, the planning of meetings, advance publicity and the recording and mailing for all activities, including 800 mailings to Group members during the year.

Symposia: The procurement of papers and actual management of national symposia are entirely in the hands of the Professional Groups. Each of the Groups sponsored one or more technical meetings during this year, in addition to technical sessions at the IRE International Convention, the WESCON, the National Electronics Conference and other jointly sponsored meetings, for a total of 59 meetings of national and international import in 1960 (see list attached). The Groups participated in the following meetings held overseas:

- International Conference on Automatic Control, June, 1960, Moscow, USSR
- Third International Conference on Medical Electronics, July, 1960, London, England
- International Information Theory Symposium, August 29–September 3, London, England
- International Symposium on Data Transmission, September, 1960, Delft, Netherlands

Publications: During the year 27 Groups published TRANSACTIONS. Since publication began in 1951 the number of issues and pages published have expanded each year. Full details in Group TRANSACTIONS are included in the Report of the Editorial Department.

Professional Group Chapters: 279 Professional Group Chapters have been organized by Group members in 60 IRE Sections. Chapter growth is continuing at a healthy rate. Most of the Chapters are meeting regularly and sponsoring meetings covering the fields of interest of their associated Groups in the various Sections of the IRE.

Section Activities

We were glad to welcome four new Sections into the IRE during the past year. They are as follows: Geneva, Kitchener-Waterloo, Las Vegas, and Mobile.

The total number of Sections is now 109.

The Subsections of Sections now total 28, the following having been formed in 1960: Catskill (New York) and Pikes Peak (Denver).

A growing major activity of many Sections and the larger Subsections in recent years is the publication of a local monthly Bulletin to fulfill the need for announcing to the Section member the increasing activities of the Section, including 1) Section Meetings 2) Professional Group Chapter Meetings, and 3) Information on the local and national level of interest to the Section member.

Fifty-four of the Sections and Subsections are now issuing these monthly publications.

Student Branches

The number of Student Branches formed during 1960 was 13. The total number of Student Branches is now 196, 124 of which operate as joint IRE-AIEE Branches, 24 as Student Associate Branches, and 1 as a joint IRE-AIEE Student Associate Branch.

Following is a list of the Student Branches formed during the year: Arlington State College, University of Buenos Aires (Argentina), Merrimack College, Municipal University of Omaha, State University of New York, New York Trade School, Oklahoma City University, Port Arthur College, New York Trade School; Oklahoma City University, Port Arthur College, Rochester Institute of Technology, Saint Joseph's College (Pennsylvania), Southern Technical Institute, Tri-State College (Indiana), and Valparaiso Technical Institute (Indiana).

Books

Frequency-Power Formulas, by Paul Penfield, Jr.

Published (1960) by The Technology Press, Mass. Inst. Tech., Cambridge, and John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 138 pages+2 index pages+viii pages+7 bibliography pages+20 appendix pages. Illus. 6×9¼. $4.00.

This is a research monograph that contains a systematic and general treatment of frequency-power formulas. It discusses four such general formulas. Of these, the Manley-Rowe formulas, first developed in the study of power flow at various frequencies in a nonlinear capacitor, are the best known, and possess the greatest applicability. The three other frequency-power formulas are less well known, and presently have less practical uses. It is established that any system with an energy state function, and in particular that distributed systems that obey Hamilton's principle, obey the Manley-Rowe formulas.

The book is divided into three parts: Part I discusses and proves the four frequency-power formulas; Part II describes systems that obey formulas of each type; Part III discusses uses, applying them to rotating machines and communication systems. An appendix lists devices known to obey each formula. The author anticipates that the results of this book will be useful in future work on energy and frequency conversion, parametric excitation, hydrodynamic and magnetodynamic stability.

Because of its specialized nature, the book will appeal principally to those who are engaged in some related field of endeavor. But to these research workers, the book will be of real interest.

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Plasma Physics, by James E. Drummond

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 372 pages+13 index pages+xiv pages+bibliography by chapter. Illus. 6¼×9¼. $12.50.

This book has its origins in a series of seminars in plasma physics held in 1957–1958. Consequently the material is at seminar level and is definitely for readers well advanced in physics. The listing of very comprehensive references provides a means, although not an easy one, for those who must acquire more background for the subject.

The organization of the book is in three main sections: "Basic Plasma Theory," "Magnetohydrodynamics," and "Microwave Plasma Physics." It has eleven different contributors, so that choice, emphasis, and treatment of topics are inevitably heavily influenced by the interests of the individual authors. These contributors are: Y. Klimontovich and V. P. Silin on collective excitations in plasmas; E. Meeron on the statistical mechanics of plasmas; P. A. Sturrock on amplifying and evanescent waves; Y. A. Yoler on magnetohydrodynamics; O. Buneman on relativistic plasma waves; D. Finkelstein on relativistic self-focusing streams; P. A. Goldberg on high-altitude ionized shock waves; J. E. Drummond on two specific microwave plasma problems; C. B. Wharton on microwave diagnostics; and S. C. Brown on microwave measurements of gas discharges. In some of these chapters it is pleasing to find the fundamental aspects of the topics laid down in both precise and concise terms, which could only result from mastery of the subject and an appreciation of its relation to the general framework of physics. In particular, Drummond's introduction is especially valuable. Meeron's discussion of the Debye length and Sturrock's discussion of dispersion relations are also good.

There is still a need, which this book does not fulfill, to unify and consolidate the physical approach, terminology, and mathematical description of plasma physics for the student. Such unity, of course, only comes as a subject approaches greater maturity than plasma physics now enjoys. Indeed, the subject is currently suffering from an excess of theory over critical experiment, so that the usefulness of all the manifold theories has not been adequately verified. Wisely, the book avoids attention to the complicated and imperfectly understood empirical devices of large scale plasma experimentation. Lack of attention to production and transport of optical radiation will appear as a deficiency to the astrophysically trained.

In general, however, this is as useful a collection of topics in plasma physics as is now available, and is pointed toward fundamental principles and illustrated with well-developed examples.

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High Frequency Applications of Ferrites, by J. Roberts

Published (1961) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 159 pages+5 index pages+x pages+2 bibliography pages. Illus. 5×7½. $4.85.

This book is, in effect, an extended review article on the subject of high frequency applications of ferrites. It is very readable and deals effectively with several basic processes in magnetic materials, although in some cases the author has presented only one of several possible interpretations. The book is intended for advanced students in electrical engineering and physics, and is written at a level of easy understanding for this audience. While it is not complete enough to serve as an editorial reference or as a text book, it would nevertheless be informative to the audience it was designed to reach.

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Design for a Brain, 2nd Edition, by W. Ross Ashby

Published (1960) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 280 pages+4 index pages+2 appendix pages+ix pages. Illus. 5¾×8¾. $6.50.

When the first edition of "Design for a Brain" appeared in 1952, any attempt to provide a systematic framework for the logical study of adaptive processes in biological systems was sure to be welcomed by those of us who were concerned with related theoretical models—a field for which this reviewer has proposed the term "neurodynamics." The simple, lucid exposition which is characteristic of Ashby's writing did much to win converts among biological and behavioral scientists, for whom a more formal treatment would have proven excessively forbidding. After eight years, however, it seems reasonable to reappraise the new edition of this book in terms of its contribution to the enduring structure of neurodynamic theory, which, it is hoped, is beginning to emerge. From this standpoint Ashby's work, though still provocative, is apt to be somewhat disappointing for a reader who seeks some understanding of brain functioning in terms of its physical structure.

"Design for a Brain" is concerned primarily with the problem of adaptive behavior, defined in terms of abstract systems of interacting variables. The recognition of the importance of studying closed systems is a major asset to this work. The concepts of stability, ultrastability, and habituation are defined, and a number of basic theorems are demonstrated, with a more rigorous treatment in a mathematical appendix. The homeostat, a physical model for an adaptive system consisting of linked servomechanisms, is described in detail, and there is some discussion of the problems of minimizing convergence time in adaptive systems. This is particularly critical in devices like the homeostat, which, in effect, execute a random walk through the phase space of the system until they happen to achieve a "successful" configuration in which stability is possible. There is little reference, however, to the physical structure or dynamics of either a natural or an artificial nervous system; the application of Ashby's principles to a network consisting of *neurons* rather than "variables" is left largely to the imagination. Indeed, the book might equally well have been entitled "Design for an Organism," as it is concerned with adaptation at so abstract a level that its principles could apply equally well to the processes of evolution, maturation, or social organization.

This is at once valuable and frustrating; valuable inasmuch as it suggests a general mode of approach to all adaptive processes, but frustrating in the obscurity of its relationship to the brain problem. The question is whether the brain can best be understood by an analysis of the abstract "logic of systems," or whether it is best approached as a problem in the analysis of a particular physical organization of unusual complexity. While these two strategies are likely to provide complementary insights into the problem, this reviewer tends to favor the second attack as the more promising one at this time. If this view proves correct, then Ashby's work will find its place more appropriately in the metamathematics of biological systems than as a building block in the mathematical development of a physical theory.

Whichever way this may turn out, "Design for a Brain" is assured a place as one of the landmarks in the awakening interest in theories of brain mechanisms which has occurred during the last two decades.

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Fundamentals of Signal Theory, by John L. Stewart

Published (1960) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 335 pages+10 index pages+xiii pages. Illus. 6¼×9¼. $9.00.

The title of this book might mislead some readers because it is primarily concerned with the mathematical concepts and techniques useful in the analysis of linear time-invariant systems. Its level of presentation is that of an average senior course. The only part of the book that considers random signals is the second half of the last chapter and this discussion requires 26 pages!

The first part of the book (166 pages) reviews the mathematics and develops the methods and concepts of steady-state analysis; in order of presentation: phasors, complex numbers, transfer functions (25 pages); differential equations, linear algebraic equations, determinants, matrices, flow graphs (25 pages); poles and zeros, partial fraction and continued fraction expansions, root loci, Nyquist criterion (65 pages); maximally flat, Tchebycheff approximation (40 pages).

The second part of the book is concerned primarily with the mathematics of transient analysis. First, transfer functions are considered as operators; the unit impulse response is written $H(p)\delta(t)$, (26 pages). Next, the Laplace transform is introduced (23 pages), followed by some complex function theory (49 pages). The latter includes Jordan's lemma and Hilbert transforms. Then come Fourier series and integrals (33 pages), followed finally by 26 pages of random signal analysis.

Frankly, I must report a slight lack of enthusiasm for the book. First, there is a problem of organization. For example, does it make sense to present Nyquist's criterion before the two chapters on complex variables? What is gained by introducing both the operational and the Laplace transform interpretation of transfer functions? Second, there is the difficulty created by the spread in the level of the material presented. For example, the first few chapters might be adequate for sophomores but as review chapters for seniors they are, in the reviewer's opinion, too vague and intuitive. For example, the requirements placed on systems to be called linear time-invariant include systems that do not obey the superposition theorem. On the other hand, the book is successful in mentioning at appropriate places interesting applications and uses of the techniques presented.

This book might be used in a senior course entitled "Analysis of Linear Systems." It will, however, have to displace well-established texts in this field.

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The McGraw-Hill Encyclopedia of Science and Technology

Published (1960) by the McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 8500 pages +536 index pages. Illus. 7¾×10. 15 vols. $175.00.

The reviewer found his examination of this work to be a fascinating and informative occupation. The scientific investigator of today's space age cannot practice intellectual myopia by pursuing his narrow specialty and disregarding the wealth of information developed by other disciplines. Rarely is the scope of a technical problem so limited that one cannot benefit from the experience garnered by investigators in other technical fields of endeavor. Acquiring an acquaintanceship with other fields is a goal not easily achieved through the medium of handbooks and similar reference works. An entirely different type of pedagogical medium is required, one that takes the reader by the hand and builds complex technical theories from "concepts of one syllable." The periodical, *Scientific American*, does this very successfully and is used by many of this reviewer's scientific colleagues to "stretch" their minds. This suffices for the leisurely reader who does not have an immediate and pressing goal, but for the investigator with an immediate specific need, a more methodical compendium of broad technical scope is required. "The McGraw-Hill Encyclopedia of Science and Technology" fills this latter requirement quite amply.

The entries in the Encyclopedia fall within the broad areas of Life Sciences, Earth Sciences, Physical Sciences and Engineering. These broad areas are divided into more than 60 subdivisions, each organized by a consulting editor. A typical subdivision, Electronics, for which Donald G. Fink is the Consulting Editor, is comprised of more than 200 articles, each written by an expert. The 7224 individual articles of the Encyclopedia, ranging from concise definitions of only 100 words to lengthy, 24,000-word dissertations, were written by more than 2000 expert contributors.

A reviewer should generally avoid a cataloging of a book's contents; however, in the case of this review, it is almost mandatory that it be done in a modified fashion if only to delineate the scope of the Encyclopedia. The major subdivisions are: Acoustics, Aeronautical Airframes, Agriculture and Soils, Analytical Chemistry, Inorganic Chemistry, Organic Chemistry, Electrical Engineering, Electricity, Electronics, Graphic Arts, Growth and Morphogenesis, Heat, Microbiology, Medical Microbiology, Mineralogy and Petrology, Petroleum Chemistry, Petroleum Engineering, General Physiology, Animal and Plant Anatomy, Animal Systematics, Anthropology, Physical Chemistry, Civil Engineering, Communications, Flight Science, Food Engineering, Forestry, Industrial and Production Engineering, Low Temperature Physics, Machine Design, Mining Engineering, Naval Architecture and Marine Engineering, Nuclear Engineering, Plant Physiology, Solid State Physics, Space Technology, Archeology, Astronomy, Atomic, Molecular and Nuclear Physics, Conservation, Control Systems, Cytology, Genetics and Evolution, Geochemistry, Physical Geography, Mathematics, Mechanical Power, Classical Mechanics, Oceanography, Optics, Paleontology, Theoretical Physics, Plant Taxonomy, Propulsion, Biochemistry, Biophysics, Chemical Engineering, Animal Ecology, Surficial and Historical Geology, Geophysics, Metallurgical Engineering, Meteorology and Climatology, Animal Pathology, Physiological and Experimental Psychology, and Invertebrate Zoology.

The usefulness of this encyclopedia rests not alone on the excellence of material choice and presentation but also on the extensive cross-reference system employed. Each article contains numerous references to companion articles in the same subdivision; thus, by following the leads from one article to others, the entire subdivision can be systematically explored. Most of the articles contain bibliographical references, permitting the serious reader to pursue a particular topic further. At the close of each article is a two- to four-letter code through which the author and his affiliation may be identified. The articles written by each author are listed so that one may completely explore a writer's point of view on a subject. The Index Volume has been carefully organized to provide rapid and efficient access to the Encyclopedia.

The publishers plan to issue annually a two-part year book. Part I will feature articles on the important scientific breakthroughs of the previous year. It is intended that this material will be rewritten and integrated into future printings of the Encyclopedia. Part II will contain shorter articles of less earthshaking importance but required to update the Encyclopedia. These latter articles will probably be added verbatim to future printings.

These volumes can serve two groups: those that require an authoritative reference work having broad coverage and those who wish to broaden their technical knowledge in an interesting painless manner. The authors impart their information with varying degrees of effectiveness but on the whole the overwhelming majority have done their job well. Leisurely browsing through these volumes can be an interesting and rewarding experience for the casual reader.

This reviewer can find fault with the Encyclopedia only by being picayunish, and he has no intention of stooping to do so just to make this a convincing "critical" review. With no reservations, this work can be commended as much needed and well done.

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Boolean Algebra and Its Applications, by J. Eldon Whitesitt

Published (1961) by Addison-Wesley Publishing Co., Inc., Reading, Mass. 167 pages+4 index pages +x pages+bibliography by chapter+10 pages answers to selected problems. Illus. 6¼×9¼. $6.75.

This book is an introductory treatment of Boolean algebra with applications to Symbolic Logic, Switching Theory and Probability in Finite Sample Spaces. It was designed to be used as a text in a one-semester course for students of mathematics and engineering. As stated by the author, "No particular subject matter is prerequisite to the study of this text, although any maturity gained in other mathematics courses will be helpful."

The book contains seven chapters. Chapter 1 is a treatment of the algebra of sets from an intuitive point of view, *i.e.*, a treatment in which illustrations and examples facilitate understanding. Chapter 2 presents Boolean algebra formally, without any reference to applications. Chapter 3 is an introduction to Symbolic Logic, utilizing especially the algebra of propositions, a Boolean algebra. Chapters 4-6 deal with the application of Boolean algebra to Switching Theory, *i.e.*, the analysis and synthesis of control and of computer circuits. Chapter 4 deals with switching circuits, the design of circuits possessing given two-state properties. Chapter 5 deals with relay circuits and control problems, including sequential circuits. Chapter 6 deals with the logical design of circuits for arithmetic computation, particularly binary addition and subtraction. Finally, Chapter 7 is a brief introduction to Probability in Finite Sample Spaces, utilizing parts of the algebra of sets as fundamental concepts. Well-chosen exercises and a short but adequate list of references for further study are given in each chapter. Answers to selected problems and an index are included.

As an introductory text, this book is excellent. It is definitely recommended, particularly to engineering students and others who wish to study the increasingly important applications of Boolean algebra in the computer field.

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Modern Mathematics for the Engineer, Second Series, Edwin F. Beckenbach, Ed.

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 441 pages+14 index pages+xviii pages+bibliography by chapter. Illus. 6×9¼. $9.50.

This book, like the earlier work of the same title, is a series of lectures presented in the well-known University of California Extension Series, "Modern Science for the Engineer." The list of contributors to this book, as that of its predecessor, is a roster of mathematicians who have made significant original contributions toward the solution of contemporary problems in science and technology.

The three major divisions of the book—"Mathematical Methods," "Statistical and Scheduling Studies," and "Physical Phenomena"—each contain five or six chapters of about 25 pages in length. As might be expected of a book of this nature, the level and style of presentation varies considerably from chapter to chapter. Some of the contributors assume little mathematical sophistication of the reader and proceed by the way of familiar examples; others use a more general approach. Yet, throughout the work, one can detect a coherent objective: to give the reader an intuitive feeling for the subject matter instead of a dry recital of the mathematical formalism. Virtually every chapter exudes the enthusiasm of the author, and is likely to capture the interest and imagination of the reader.

Since these lectures were intended not only for electrical engineers, one might suspect that much of the material is not relevant. However, even the chapters on such topics as theory of inventory processes, Monte Carlo calculations in mathematical physics, and fluid mechanics are likely to be of interest—not only to workers in these specific areas, but also to others who might find the *techniques*, if not the results, applicable to their own problems.

Except for the bibliographies at the end of each chapter, the book should not be treated as a reference handbook. It is unlikely that the reader will find that his problem becomes completely transparent as a result of his reading the appropriate chapter. Rather he should expect only to find a clue to a possible solution, and this perhaps in an unlikely place. Hence, instead of putting it on the reference shelf for emergency use, the reader should read the book leisurely and reflectively. In so doing he will be introduced to the theory of distributions, modern operator calculus, random processes and information theory, optimum control theory, linear programming, and more; and he will be treated to a delightful essay on the intuitive method by its most famous exponent, George Pólya.

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Scanning the Transactions

A 300-year-old antenna design, despite the fact it predates radio itself, still qualifies as one of the most interesting "new" antenna designs in recent years. The new antenna takes its shape from the Cassegrain telescope, which originated in the 17th century. A Cassegrain telescope consists of two mirrors instead of one. A large concave primary mirror collects the incoming light and reflects it to a small convex mirror directly in front, which in turn reflects the light back again through a small hole in the center of the primary mirror to an observer or camera behind. It combines the advantages of a short focal length and permitting the observer to remain behind, instead of in front of, the main mirror. This same geometry has now been adapted as a microwave antenna, with the transmitter (or receiver) and its feed replacing the observer behind the center of the primary dish. The arrangement leads to a number of interesting applications, such as very-low-noise reception and novel two-in-one antennas. (P. W. Hannan, "Microwave antennas derived from the Cassegrain telescope," IRE TRANS. ON ANTENNAS AND PROPAGATION, March, 1961.)

The Fabry-Perot interferometer is becoming an increasingly familiar and important figure on the electronics scene. In essence, it is a reflector system in which multiple reflections of plane waves occur between two parallel reflecting plates, forming in effect a resonant cavity. The Fabry-Perot interferometer is usually associated with optical phenomena. It has been mentioned with increasing frequency in the electronics literature of late because it forms an important part of the optical maser. However, there is also growing interest in this technique in connection with the upper end of the radio spectrum, in the millimeter and submillimeter region. An interferometer of this type has now been successfully operated at 8 mm. This development may well lead to the extension of klystron techniques to shorter wavelengths, and perhaps may eventually play a role in closing the gap between the millimeter and infrared regions of the spectrum. (W. Culshaw, "Resonators for millimeter and submillimeter wavelengths," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, March, 1961.)

Maiden names have a marked propensity of being changed after the second decade in festive ceremonies marked by champagne and rice. Although no rice and champagne are at hand and only one decade has passed since their birth, we wish to announce that the maiden names of three of the family of IRE Transactions have been changed this year: Medical Electronics has become Bio-Medical Electronics; Production Techniques has changed to Product Engineering and Production; and Aeronautical and Navigational Electronics has dropped "Aeronautical" in favor of "Aerospace." These are by no means the first such changes to be recorded. The last named Group at one time was not "Navigational"; Reliability and Quality Control once had no "Reliability"; and Space Electronics and Telemetry started out life as Radio Telemetry and Remote Control. As other eligible young subjects come upon the scene in the future we shall, no doubt, see more marriages in the family.

A novel three-in-one waveguide has been proposed for transmitting three different messages at one time inside a single closed waveguide. The configuration being explored consists of a double ridged waveguide having two center conductors. In cross section it looks much like a large block letter H, with a conductor in the center of each vertical portion. Recent theoretical and experimental studies show that by launching a TEM mode on each conductor and by also using the dominant waveguide mode, it seems likely that a three-in-one transmission scheme could be made to work using one mode for each message. (J. E. Storer and T. W. Thompson, "TEM impedance and cross coupling for small circular center conductors in a double ridged waveguide," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, March, 1961.)

Definition of a Committee: A group of the unwilling chosen from the unfit to perform the unnecessary. (H. A. Zahl, "Looking backward toward tomorrow," IRE TRANS. ON AEROSPACE AND NAVIGATIONAL ELECTRONICS, March, 1961.)

Parametric feedback is a new concept which describes a way of looking at nonlinear filters that sheds interesting new light on them. In a recent study it was found that quasi-linear forced oscillations can be regarded in filter theory as a form of parametric feedback whereby the system parameters are functions of the output. The idea of parametric feedback leads naturally to the inclusion phase, as well as amplitude, feedback. This new perspective opens up a broad class of filters with some rather startling characteristics. One example that can be cited is a filter with phase feedback which can discriminate between amplitude and phase modulation by passing the former and rejecting the latter. In addition to phase or amplitude filters, the unusual properties of parametric feedback can be used to construct modulation amplifiers, modulation function generators, and phase-to-amplitude modulation converters. (D. A. Robinson, "Modulation in nonlinear filters with parametric feedback," IRE TRANS. ON CIRCUIT THEORY, March, 1961.)

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. 1	$2.25	$3.25	$4.50
Circuit Theory	CT-8, No. 1	2.25	3.25	4.50
Communications Systems	CS-9, No. 1	2.25	3.25	4.50
Component Parts	CP-8, No. 1	2.25	3.25	4.50
Information Theory	IT-7, No. 2	2.25	3.25	4.50
Microwave Theory and Techniques	MTT-9, No. 2	2.25	3.25	4.50
Nuclear Science	NS-8, No. 1	2.25	3.25	4.50
Reliability and Quality Control	RQC-10, No. 1	2.25	3.25	4.50

Aerospace and Navigational Electronics

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The Editor Reports (p. 2)

Looking Back Toward Tomorrow—Harold A. Zahl (p. 3)

Post-War Developments in Continuous-Wave and Frequency-Modulated Radar—W. K. Saunders (p. 7)

This paper is a survey of trends in the post-war development of continuous-wave and frequency-modulated radar with emphasis on systems designed to operate to short range. The larger part of the material is concerned with systems which use a single microwave generator as both a transmitting tube and as a source of local oscillator power. The discussion is organized around the various methods of analysis which may be used to study these systems. Hitherto unpublished work on a statistical method, developed at the Diamond Ordnance

Fuze Laboratories, is summarized. Some information of a practical nature on microphonism, nonlinearities, feed-thru, and saturation is included. A bibliography of work done in the post-war period is also included.

An Analysis of Bistatic Radar—Merrill I. Skolnik (p. 19)

This paper briefly describes some of the characteristics, capabilities, and limitations of the bistatic radar when used for the detection and location of targets such as aircraft or satellites. (A bistatic radar is one in which the receiver is physically separated from the transmitter so that the echo signal does not travel over the same path as the transmitted signal.) Among the topics discussed are the type of information available from the received signal, methods for extracting information from the bistatic radar signal, bistatic radar equation, target cross sections, and the properties of the bistatic radar fence. Throughout the paper the bistatic radar is continually compared to the conventional monostatic radar. It is concluded that although the bistatic radar has several interesting attributes, it is not as generally applicable as the more versatile monostatic radar because of the limited fence-type coverage and the difficulty of extracting target location information. An appendix is included which describes the multiple-frequency CW method for measuring distance as applied to either the bistatic or the monostatic radar.

A Self-Contained Interplanetary Navigator—H. B. Haake and J. D. Welch (p. 28)

This paper is concerned primarily with various methods that can be used to mechanize a self-contained space navigation system, using a positional perturbation guidance equation to compute velocity corrections. The key components involved are those used to determine present position, a control system to implement velocity corrections, and a computer to interrelate their functional operations.

In the area of determining present position, various schemes are described for obtaining a celestial fix. These include purely-optical and inertial-optical schemes, which are compared. In addition, a least-squares data smoothing criterion is shown, whereby position is computed using redundant fix data.

The implementation of velocity corrections is largely dependent upon the ability to control accurately vehicle attitude and to measure the magnitude of the velocity increment. Therefore, various schemes are proposed for controlling the direction and magnitude of a velocity correction.

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Circuit Theory

Vol. CT-8, No. 1, March, 1961

A Method of Tree Expansion in Network Topology—Hitoshi Watanabe (p. 4)

This paper presents a systematic method of obtaining the total sum of all possible tree admittance products of a very complicated network. First, such a network is partitioned into several simple subnetworks by a set of independent cut sets. Then, the relationship between trees or tree admittance products of the network and multitrees or multitree admittance products of each subnetwork is found and presented as the tree expansion theorem. Finally, an effective method for decomposition of the large network into simple subnetworks is discussed. The methods described in this paper are very suitable for use by automatic computation.

On the Realizability of a Set of Trees—S. L. Hakimi (p. 11)

A matrix with entries 0's and 1's whose rows correspond to the trees of a graph G and whose columns correspond to the elements of G is introduced. Many interesting properties of such a matrix are derived. The concept of the rank (modulus 2) of the tree matrix is found to be very useful in determining the number of separable parts of the corresponding graph. A simple algebraic way is presented by which one can find, from a given set of trees (or tree matrix), the fundamental circuit (or cut-set) matrix with respect to a prespecified tree. It is also shown that one can easily find, from the tree matrix, the set of all paths between the vertices of an element of a graph. Some interesting conjectures are stated concerning graphs with a given rank and nullity which have a minimum number of trees.

The Seg: A New Class of Subgraphs—Myril B. Reed (p. 17)

The linear graph, after a delay of about a hundred years from its conception by Kirchhoff, is today rapidly assuming a dominant position in the foundation of electric network theory. This paper is a presentation, by definition and properties deducible therefrom, of a new class of subgraphs (segs) which includes stars and cut sets as special cases. The matrix associated with the seg can be shown to be the general coefficient matrix of the Kirchhoff current equations. Hence, in addition to adding to the structure of linear graph theory, the results in this paper broaden the base of knowledge of currents in electric networks. This broadened base has already permitted the opening of a new phase of network theory and can certainly be expected to open others.

On the Synthesis of Resistive *N*-Port Networks—G. Biorci and P. P. Civalleri (p. 22)

A new approach to the problem of the synthesis of resistive n-port networks with $(n+1)$ nodes is given. First it is shown that, given a conductance matrix G of order n, the signs of its elements define the complete tree of the n port uniquely (if it exists). The case of two (or more) ports in "series" is exceptional: the order of such ports in the complete tree can be found by comparing the absolute values of a couple (or more couples) of elements of G. Once the complete tree has been determined, the synthesis becomes very simple, since the G matrix can be transformed into another matrix G_0 referred to a set of node-to-datum independent voltages, whose conditions of realizability are well known.

Paramount Matrices and Synthesis of Resistive *N* Ports—I. Cederbaum (p. 28)

A method of reducing the main diagonal elements of a paramount matrix and its inverse without destroying their paramount character is described. It is shown that in not more than $2n$ steps the irreducible remainder of a given matrix can be found. Also, the conditions are formulated for a paramount matrix with non-positive off-diagonal elements to be the impedance matrix of an n-port network with the minimum number of independent circuits.

On the Synthesis of *R*-Networks—D. P. Brown and Y. Tokad (p. 31)

A study is made of a class of real symmetric matrices, and a new set of necessary and sufficient conditions is defined on the entries of these matrices such that they can be synthesized, without using ideal transformers, as an R-network with a minimum number of terminals. The conditions are stated in terms of the terminal graph used in representing the terminal characteristics of multiterminal components. The new class of matrices then represents "terminal equations" corresponding to a path-tree terminal graph.

It is also shown that one of the well-known classes of matrices, referred to as dominant matrices, represent the terminal equations for a Lagrangian-tree terminal graph. It is further indicated that any such class of real symmetric matrices is distinguishable by a particular terminal graph.

For the cases when a real symmetric matrix of order n cannot be synthesized by an R-network with $(n+1)$ terminal vertices, an "enlarged" matrix is formed, and the necessary and sufficient conditions for realizability of these matrices are given.

Analysis and Synthesis Techniques of Oriented Communication Nets—D. T. Tang and R. T. Chen (p. 39)

An oriented communication net is a communication network in which channel capacities between pairs of terminals are not symmetrical. Such a system can be represented by an oriented graph. The concept of a minimum-valued cut is used to determine the terminal-capacity matrix which gives the maximum possible communication between any ordered pair of terminals.

Necessary conditions for a terminal capacity matrix to be realizable as an oriented communication network are obtained. These conditions are shown to be sufficient when the order of the terminal capacity matrix is three or less. They are not sufficient for higher-order cases in general. Synthesis techniques based on matrix partition and matrix addition are used to realize any three-by-three terminal capacity matrix with extensions to some higher-order cases.

Optimal Synthesis of a Communication Net—O. Wing and R. T. Chen (p. 44)

This paper gives solutions to the problem of realizing a communication network at minimum cost. The network is composed of a set of nodes connected by a set of branches. Every branch has associated with it a capacity. The required amount of flow between every pair of nodes is specified. The unit costs of the branch capacities are given. The problem is to find the network and the branch capacities such that the total cost is minimum. The set of branch capacities and the set of terminal demands are shown to satisfy a set of linear inequalities. Linear programming is used to obtain the optimal solution. In the case of identical unit costs, several realizations are given which require fewer branches than previously reported.

Flowgraphs for Nonlinear Systems—T. A. Bickart (p. 49)

The development of a nonlinear flowgraph notation for the representation of a large class of nonlinear systems is the central topic of this paper. By the definition of power operator branches, exponential operator branches, switching operator branches, and limiting operator branches, it becomes possible to represent a variety of nonlinear systems. In particular, it becomes possible to represent in a very simple manner a number of different types of hysteretic systems. A study of the rules for manipulation of nonlinear flowgraphs is made, so as to bring out the procedures that might be used in reducing the complexity of such a system as that represented by the flowgraphs. Generalized flowgraph inversion procedures are developed as a consequence of the defined nonlinear branch transmittances. A variety of nonlinear systems are illustrated to indicate the broad applicability of the notation in representing such systems.

Darlington's Synthesis and RMS Error Evaluation—A. Papoulis (p. 58)

A theorem relating the integrated square of the impulse and step response of a two terminal-pair network to its input impedance is developed; this theorem is in a sense the time domain equivalent to the gain-bandwidth relationship in steady state. As an application, a simple method is presented for evaluating the integrated square,

$$\int_0^\infty [r(t) - r_{ss}]^2 dt.$$

of the difference between a signal $r(t)$ and its steady-state value r_{ss}; the method is based on Darlington's procedure for determining a reactive network from its transfer function.

Modulation in Nonlinear Filters with Parametric Feedback—D. A. Robinson (p. 60)

A class of nonlinear second-order filters is considered which may be described and analysed by the theory of quasi-linear differential equations. It is shown that the effect of nonlinearity in these equations is equivalent to that obtained by parametric feedback where the system parameters are functions of the output amplitude. This leads to an extension of the filter class by admitting phase parametric feedback. Filters of this class exhibit startling characteristics in comparison with the behavior of LLFPB circuits. The methods of Mandelstam and Papalexi, and Andronow and Witt in complex notation underlie the derivations. After examining the solution under sinusoidal inputs, the more interesting problem of modulation response is considered. Under small signal conditions a second-order linear differential equation with complex coefficients is derived to describe the modulation response. One example is offered of a filter with phase feedback that discriminates between amplitude and phase modulation by passing the former and rejecting the latter.

Synthesis Using Tunnel Diodes and Masers—Louis Weinberg (p. 66)

Methods are suggested for the realization of multistage amplifier networks containing negative resistances. These methods are exact synthesis procedures which are based on ***reverse predistortion***, both uniform and nonuniform. They thus realize a specified function of frequency—its magnitude, its phase, and its constant multiplier. The methods apply to the realization of tunnel-diode amplifiers if the equivalent circuit of a tunnel diode is considered to be a parallel connection of the junction transition capacitance and a negative resistance. They can be used to realize cascaded maser amplifiers if the equivalent circuit of a maser amplifier is taken to be a parallel connection of an inductance, a capacitance, and a negative resistance, or a series connection of these three elements. The methods used automatically provide for the stability of the multistage active networks.

The straightforward technique of predistorting the complete system function may for some applications require too many active elements. Simple modifications of this technique are given which allow a control over the position in the network, the time constants, and the required number of the active elements.

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Communications Systems

Vol. CS-9, No. 1, March, 1961

Why in the World in Europe?—Bruce B. Barrow (p. 3)

The International Symposium on Data Transmission—H. C. A. Van Duuren, F. L. Stumpers, and B. B. Barrow (p. 4)

Notes on the Transmission of Data at 750 Bauds over Practical Circuits—P. A. Chittenden (p. 7)

This paper describes experience over some 12 months of using a 750-baud phase-modulated data-transmission equipment over British Post Office trunk-line circuits, with a brief description of the equipment and test gear and of the lines used.

It is concluded that line attenuation and general circuit noise are not limiting transmission characteristics in the magnitudes to be expected. Delay distortion and impulsive noise can be troublesome, but the most serious disability is short interruptions of the transmission path lasting from a few milliseconds to a few seconds.

Because of this, the view is expressed that the assessment of the suitability of circuits for data transmission by quoting average digit error rates is unrealistic, and a figure for the number of minute periods per day when error-free operation can be expected is to be preferred; also, that with the duration of the interruptions encountered, automatic error correction would be complicated, and have to be supported by a system of repetition of messages in error. In the interests of simplicity, this could be the preferred solution even for single errors, since they appear to occur infrequently.

The view is expressed also that users who cannot rely on a network provided by fixed plant in a civil area, *e.g.*, a military user in the field, may well have a worse interruption problem due to the inclusion of numerous cable couplers in relatively short line circuits, and could better be served by a system of telegraph channels, each working at a slow digit speed, in order to provide a long digit length to combat the interruptions.

Error Rates and Error Distributions on Data Transmitted Over Switched Telephone Connections—E. P. G. Wright (p. 12)

This paper describes the results of tests which have been carried out in order to collect material with respect to errors caused by interference during the transmission of data. The tests have been carried out on a variety of local and toll switched telephone connections at 250, 500, and 1000 bauds. The signal has been set at various levels between -6 dbmo and -46 dbmo. Frequency modulation using only a portion of the available bandwidth has been employed so that a return path can be provided for signals in the reverse direction. Statistics are included in respect to data blocks of 50, 100, 200, and 500 bits. The attenuation-frequency characteristics of a number of test connections are given as both mean and extreme measurements.

Many of the tests described were made before the issue of the Test Program drawn up by Working Party 43 of the CCITT in March, 1960, but in most important respects all the tests are in agreement with the procedure recommended.

Errors in Data Transmission Systems—R. G. Enticknap (p. 15)

The aims and methods of a program of work being conducted at the Lincoln Laboratory of the Massachusetts Institute of Technology in data-transmission research are discussed. The work, which has been concentrated primarily on telephone circuits of a type generally available, is aimed at a better understanding of these transmission media. Machines for the measurement of error and error distributions occurring in data-transmission channels and machines for recording impulsive-type noise on telephone circuits are described. Some results from both types of measurement are presented. One of the most significant results of this program has been the demonstration that errors occur on telephone circuits in bursts. Among the major sources of errors on telephone circuits is the burst of additive noise. With the machine described above, some success has been achieved in classifying several of the types of burst which occur and in deducing their probable causes.

HF Radio Data Transmission—Bernard Goldberg (p. 21)

An exposition of some of the past, present and future activity of the U. S. Army Signal Research and Development Laboratory in the field of HF radio data transmission is treated in broad terms. A limited base for use in comparison with HF channel attributes is set up by relating the data-handling characteristics of switched long-distance telephone lines.

Generalized criteria of performance for digital data and TTY transmission is reviewed, together with a discussion of what we presently know about the behavior of the HF ionospheric channel wirh regard to data transmission.

A broad and detailed coverage is given to the determination of the required channel characteristics and to the methodology associated with obtaining this information. A description of the USASRDL field test program and some of its results and implications are discussed. Curves of performance of multichannel digitally-modulated FSK and PSK systems are presented. Data-collection and -reduction techniques are described.

The over-all program is reviewed in terms of applications of the medium's statistics in a manner which points the way toward future activity in the field of transmission of digital data by HF ionospheric channels. In this regard both channel matching and coding receive their proper recognition and are discussed in terms of providing the promise of significant improvements in the HF radio transmission of digital data.

Error Rates and Error Detection on Telegraph Circuits—A. C. Croisdale (p. 28)

This paper reviews error rates obtained in tests on the full range of telegraph facilities provided by an administration at signalling speeds not exceeding 200 bauds. The telegraph facilities tested include: an inland private circuit, automatic switched circuits, international circuits, intercontinental circuits in a submarine telephone cable, a speech plus telegraphy facility on inland telephone circuits.

For transmission of data with a character error rate of better than $1/10^6$ on cable circuits, an improvement in performance will be necessary, and as this is unlikely to be obtained economically by attention to the cable or telegraph networks, the improvement will have to be achieved by error detection and retransmission. Where a short renter-to-renter circuit is involved, it may be satisfactory to use automatic error detection and manual retransmission, but over other telegraph facilities automatic retransmission may be necessary.

Tests have also been made on both unprotected radio telegraph circuits and circuits employing the Van Duuren automatic error correction (ARQ) system. An appraisal is made of the over-all performance of this system, with suggestions for improvements necessary for the transmission of data.

Where a duplex telegraph circuit is available, a simple method of error detection is to feed the received signal back to the sender where it can be checked against the original.

Discussion on Experimental Results (p. 37)

Error Probability and Transmission Speed on Circuits Using Error Detection and Automatic Repetition of Signals—H. C. A. Van Duuren (p. 38)

The performance of a telegraph system using error detection and automatic repetition of signals is analyzed for several fading and noise conditions.

At first the effect of the decision feedback is calculated—the number of errors and the transmission speed as a function of the character error probabilities. The properties of the code give the character error probabilities expressed in element error rate. The error probabilities are calculated for several detectors and fading models as a function of the signal-to-noise ratio. Using these results the behavior of the whole system is computed and graphs are given of the transmission speed and the number of undetected errors as a function of the signal-to-noise ratio. The analysis shows that with the system considered, the number of errors can be made as low as desired, for all circumstances, including interruptions. This cannot be obtained with a system using error correction.

Finally some simple equations are given for the most important points of the curves.

Some Recent Developments in Digital Feedback Communication Systems—Leonard S. Schwartz (p. 51)

This paper summarizes the results of investigations into two methods for increasing the reliability of digital communication systems: 1) by improvements in the decision system of the receiver, and 2) by the introduction of a feedback link for error correction between the receiver and the transmitter. The improved decision system provides for the withholding of decision in doubtful cases which are marked by the received signals falling in an ambiguous region called the null zone. The use of a null-zone decision region in conjunction with feedback is called decision feedback. The characteristics and performance of this system are discussed. In addition, the results of combining integration and coding with decision feedback, called, respectively, cumulative and coded decision feedback, are presented. Finally, the behavior of decision feedback under conditions of multipath reception is noted. The performance of decision feedback systems is compared with that of unidirectional systems.

Discussion on "Some Recent Developments in Digital Feedback Communication Systems" (p. 57)

Some Results on the Effectiveness of Error-Control Procedures in Digital Data Transmission—W. R. Bennett and F. E. Froehlich (p. 58)

In recent years considerable attention has been directed toward digital data communication among machines utilizing the telephone network. Such communication raises many questions concerning the accuracy of transmission attainable. This paper discusses the effectiveness of several error-control techniques.

First, the transmission channel is described by three parameters which permit correlation among the errors. A number of error-correcting methods are evaluated by a computer simulation technique using the parameters of several hypothetical transmission channels which might be representative of telephone circuits. Emphasis has been placed on the recurrent burst-correcting codes. The performance of these "recurrent" codes, when subjected to actual errors collected from Data-Phone test calls, is compared to the performance of these codes using the assumed channel parameters. Graphs relating the average final error rate after correction to the transmission channel error statistics are shown.

The difficulty of extracting channel parameters from real error data is discussed and a method is presented which might permit such calculations.

It is concluded that some of the codes considered are capable of reducing the error rates in digital communication considerably. However, most error-control methods are highly sensitive to the particular error statistics found on the transmission channel. An analytical description of the errors found on a transmission channel is needed for the proper evaluation of error-control methods.

Discussion on "Some Results on the Effectiveness of Error Control Procedures in Digital Data Transmission" (p. 65)

High-Speed Data Transmission Over a Group Link Telephone Channel—J. Labeyrie (p. 66)

The studies carried out by the Centre National d'Etudes des Télécommunications had for its object the determination of the maximum modulation rate which is possible over a group link telephone channel of a modern carrier system.

The data modulator-demodulator which has been realized for a modulation rate of about 3700 bauds makes use of vestigial sideband amplitude modulation. One particular point is the direct translation of the signals from the baseband into the 4-kc to 8-kc band at the transmitting end and inversely from the 4-kc to 8-kc band into the baseband at the receiving end. For this reason it is necessary to modify the first stages of the modulating chain and the last stages of the demodulating chain of the carrier system.

Measurements of binary error rates have been made for a speed of 3700 bauds with addition of white noise at different levels (from 20,000 pw up to 180,000 pw, referred to zero transmission level). A typical result is an error rate of 10^{-6} for a noise power of 50,000 pw.

Two applications are considered: providing special private lines for high-speed data transmission and designing time-division multichannel telegraphy.

Detection and Information Rate of Telegraphic Signals—D. A. Bell (p. 70)

"Detection" in the title of this paper is used in the sense of decision-making, rather than demodulation. The ideal communication rate for a physical channel is provided by Shannon's formula $C \leq W \log(1+P/N)$, and for a binary symmetric channel the theoretical amount of information per digit is $I = 1 + p \log p + q \log q$, where p is probability of error per digit and $p+q=1$. Various error-correcting codes are compared with the second formula. It is shown that correlation-detection is equivalent to minimum-distance detection provided all code groups are of equal mean power, and that a binary-coded signal can always be adjusted to the constant-power state. Experimental results are reported on error reduction in a two-out-of-five code by the use of correlation detection. It is noted that square-law detection would be inferior to linear detection in such a system.

Discussion on "Detection and Information Rate of Telegraphic Signals" (p. 77)

Synchronous Demodulation of Phase-Reversing Binary Signals, and the Effect of Limiting Action—D. G. Tucker (p. 77)

It is shown that a considerable improvement of SNR in the output is produced in a synchronous demodulator (when the SNR in the input is greater than unity) by interchanging the input terminals with the local oscillator terminals of the modulator circuit when this is of the switching type, or by using a balanced limiter followed by a linear multiplier. Owing to the nonlinear relationship introduced, however, the consequent improvement in error rate in a binary system is much less than the improved SNR might suggest; it is likely to be fairly small and may be nil when the decision device is efficient, but may be fairly large if the decision regarding polarity has to be made at an amplitude which is a substantial fraction of the pulse amplitude. As the performance of the new arrangement, when different, is always better than that of the usual circuit, there seems to be a good case for using it when convenient. Moreover, when it is used, there is little loss of performance with respect to error rate if the binary decision regarding polarity of output pulse is made, not at zero amplitude, but at a finite amplitude which may be as great as half-amplitude. Consequently, it is feasible to make a single receiver which will operate almost indiscriminately on ON-OFF or phase-reversing signals.

A Comparison of Frequency-Shift Anti-Multipath Signaling Techniques for Digital Communications—W. B. Jones, Jr. (p. 83)

An important result of multipath propagation is the lengthening of the received signal due to a single transmitted pulse. The received pulse duration may be increased by the difference in propagation times between the longest and shortest of the active paths. This time difference represents the minimum pulse repetition period which can be used if errors due to inter-symbol interference are to be avoided.

This minimum pulse repetition period places a limit on the available data rate per channel. The total data rate can be increased by frequency-division multiplexing. Several recent papers have described a technique for synchronous frequency-division multiplexing for use under the multipath conditions described above. This synchronous system is analyzed to show its bandwidth, data rate, and signal-to-noise properties, and these are compared with those of a conventional FSK system.

A second refinement of conventional FSK is the use of more than two possible frequencies for each pulse. The bandwidth, data rate, and signal-to-noise properties of multilevel synchronous FSK are presented, and some circumstances under which it may be used to advantage are described.

Digital Data Transmission Systems of the Future—R. J. Filipowsky and E. H. Scherer (p. 88)

General predictions are made about the development of the state of the art by shortly reviewing the latest findings of theoretical investigations as well as hardware developments. A few examples are given to illustrate recent attempts to reorganize, standardize and integrate present-day digital data processing and transmission systems for future use.

Concepts of "data moving" and "dynamic matching" within a transmission network are introduced. It is believed that tele-data processing (TDP) systems, *i.e.*, electronic data processing (EDP) systems with input/output devices at remote locations, will experience a growth that will warrant the specialization of a number of manufacturers and organizations on data moving only. Dynamic matching may be considered as the basic principle of a new organization of digital data-transmission networks. It will ultimately load every available communication channel to its practical peak capacity and will therefore optimize the efficiency of systems. This principle is discussed in detail at four different levels of a network. Demand matching of input devices, branch matching, link matching, and finally network matching will be a common feature of many systems in the 1970's.

For the user of TDP facilities, data movers will incorporate these principles in "general services" networks, which will operate on a message switched basis and in completely digital form. "Private facilities" will also be provided on a circuit switched basis for special purposes.

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Component Parts

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Thermistors, Their Theory, Manufacture and Application—R. W. A. Scarr and R. A. Setterington (p. 6)

Thermistors are components of comparatively recent development, although the phenomenon of conduction in metallic salts was recorded by Faraday over a century ago.

The paper, in a broad survey of the subject, describes the theory of their operation, explains why they, in common with other semiconductor devices, possess a negative temperature coefficient of resistance, and develops the expressions which govern their parameters.

Methods of manufacture of the various forms of thermistor are outlined in general terms, and numerous applications of the device are described and discussed. Guidance is given on the approach to circuit problems, and the possibilities of exploiting some of the device's interesting and unusual properties are indicated. The point is made that the thermistor can now be used in many fields as an adequate alternative to components of more specialized application.

A Distributed-Parameter Approach to the High-Frequency Network Representation of Wide-Band Transformers—T. R. O'Meara (p. 23)

An approach to the high-frequency representation is developed from the point of view of distributed-parameter networks. This viewpoint may be directly employed to compute insertion or transmission functions. Alternatively, it provides a method for determining the element values for lumped-parameter equivalent circuits. This method is most easily applied to transformers having two winding layers with turns ratios which are either very nearly unity or which depart very greatly from unity, and the paper largely confines itself to these cases.

Distributed-parameter insertion-voltage ratios for a number of common two-winding transformer connections are presented and compared with experimental results.

The Long-Term Stability of Fixed Resistors—H. F. Church (p. 31)

The causes of long-term failure under practical conditions of use or storage of different types of fixed resistors commonly used in electronic equipment have been investigated. Some reported life tests have proceeded without interruption for almost four years. Carbon-composition (grade 2) resistors under load fail by slow thermal degradation of the resistive material. Drift of value may also occur if unloaded resistors of this type are stored in a damp atmosphere. Vitreous-enamelled wire-wound resistors made with fine wire may fail during tropical exposure both unloaded and especially when lightly loaded with direct current. This is owing to electrochemical corrosion taking place at faults in the vitreous coating. High-stability cracked-carbon (grade 1) resistors may fail rapidly under light dc load by electrochemical action if moisture condensation occurs and the protective paint or varnish coating is inadequate.

Tests for long-term resistor stability are critically discussed.

Contributors (p. 41)

Information Theory

Vol. IT-7, No. 2, April, 1961

Effect of Hard Limiting on the Probabilities of Incorrect Dismissal and False Alarm at the Output of an Envelope Detector—P. Bello and W. Higgins (p. 60)

This paper is concerned with the effect of hard limiting on the signal detectability of a system consisting of a limiter, narrow-band filter, and envelope detector in cascade. The input to the system is a pulsed IF signal immersed in noise whose power spectrum is uniform over a band of width W cycles.

Assuming that the noise bandwidth W is much larger than the bandwidth of the narrow-band filter, the probability distribution of the output of the filter will approach Gaussian. A bivariate Edgeworth series approximation is necessary to handle the narrow-band filter output since the "in-phase" and "quadrature" components of the narrow-band-filter output are statistically dependent random variables. An expression is derived for the probability of incorrect dismissal that requires the numerical evaluation of single integrals only. From the same bivariate Edgeworth series, an expression is derived for the probability-density function of the output of the envelope detector for the zero-input-signal case. Subsequent integration leads to the probability of false alarm.

On the Asymptotic Efficiency of Locally Optimum Detectors—Jack Capon (p. 67)

A detector examines an unknown waveform to determine whether it is a mixture of signal and noise, or noise alone. The Neyman-Pearson detector is optimum in the sense that for given false alarm probability, signal-to-noise ratio, and number of observations, it minimizes the false dismissal probability. This detector is optimum for all values of the signal-to-noise ratio, and its implementation is usually quite complicated.

In many situations it is desired to detect signals which are very weak compared to the noise. The locally optimum detector is defined as one which has optimum properties only for small signal-to-noise ratios. It is proposed as an alternative to the Neyman-Pearson detector, since in practice it is usually only necessary to have a near-optimum detector for weak signals, since strong signals will be detected with reasonable accuracy even if the detector is well below optimum.

In order to evaluate the performance of the locally optimum detector, it is compared to the Neyman-Pearson detector. This comparison is based on the concept of asymptotic relative efficiency introduced by Pitman for comparing hypothesis testing procedures. On the basis of this comparison, it is shown that the locally optimum detector is asymptotically as efficient as the Neyman-Pearson detector.

A number of applications to several detection problems are considered. It is found that the implementation of the locally optimum detector is less, or at most as complicated as that of the Neyman-Pearson detector.

Frequency Differences Between Two Partially Correlated Noise Channels—Janis Galejs (p. 72)

Approximate probability distributions of the difference frequency between two noise channels which contain dissimilar Gaussian, rectangular or triple-tuned RLC band-pass filters are calculated. For noise channels that differ only in time delay, a proportionality between rate of change of instantaneous frequency and the difference frequency is assumed. For dissimilar filters, an approximately equivalent single filter-time delay process is defined. The single filter is determined from the moment averages of the two dissimiliar filters, while the equivalent time delay is computed by equating the magnitude of the correlation function in the two processes.

Complementary Series—Marcel J. E. Golay (p. 82)

A set of complementary series is defined as a pair of equally long, finite sequences of two kinds of elements which have the property that the number of pairs of like elements with any one given separation in one series is equal to the number of pairs of unlike elements with the same given separation in the other series.

(For instance the two series 1001010001 and 1000000110 have, respectively, three pairs of like and three pairs of unlike adjacent elements, four pairs of like and four pairs of unlike alternate elements, and so forth for all possible separations.)

These series, which were originally conceived in connection with the optical problem of multislit spectrometry, also have possible applications in communication engineering, for when the two kinds of elements of these series are taken to be $+1$ and -1, it follows immediately from their definition that the sum of their two respective autocorrelation series is zero everywhere, except for the center term. Several propositions relative to these series, to their permissible number of elements, and to their synthesis are demonstrated.

Signal Detection by Adaptive Filters—E. M. Glaser (p. 87)

Communication engineers are now giving increased attention to detection systems which are able to adjust their own structure so as to be optimum for the particular detection problem of the moment. This paper describes a system which is capable of adapting and optimizing its response to the class of pulse signals whose individual pulses are less than T seconds in duration.

The analysis and synthesis of the adaptive system is facilitated by the use of an orthogonal function decomposition of the received signal. The use of the orthogonal decomposition permits synthesis of optimum linear filters by various circuit techniques, several of which have been reported elsewhere. The structure of the system utilizing such a decomposition is described in detail.

Since the operation of the adaptive filter is based upon signal detection and estimation in noise backgrounds, considerable attention is devoted to the relationship between optimum signal detection and estimation. The methods of statistical decision theory are used.

A program to test the validity of the approximations and assess the over-all system performance was carried out by simulation of the system on both analog and digital computers. The results of these experimental runs are described.

The Coding of Pictorial Data—Joseph S. Wholey (p. 99)

We are concerned with the problem of designing an efficient general method of coding two-level pictorial data. Both exact and approximate coding techniques are illustrated.

A pilot experiment is presented, in which a digital computer was used to realize two-dimensional "predictive coding." Although the resulting compression was not great, there are reasons for believing that this procedure would be more successful with realistic pictorial data.

Further experiments, which made use of approximation methods, are described. These methods arose from the application of pattern recognition theory to the present problem. Their use, either independently or prior to predictive coding, yielded compression significantly greater than that attained by predictive coding alone.

On Singular and Nonsingular Optimum (Bayes) Tests for the Detection of Normal Stochastic Signals in Normal Noise—David Middleton (p. 105)

The necessary and sufficient (n. and s.) conditions for the nonsingularity, *i.e.*, regularity, and for the singularity of optimum tests for the presence of one Gaussian process vs another on a finite sample are established, for both nonstationary and stationary processes, including those with nonrational spectra. In the stationary cases, the condition may be expressed alternatively in terms of an integral of suitable spectral ratios when the random processes possess rational spectra and for certain classes of nonrational spectra as well. Equivalently, for rational spectra the n. and s. condition for nonsingularity is that the spectral ratio approach unity as frequency becomes infinite, while for singularity the n. and s. condition requires that this ratio differ from unity in the limit. Some of the implications of these results in applications to signal detection are considered, and a method of solution of an associated class of integral equations, of the type

$$\int_{0-}^{T+} L(\tau, u)K(|u - t|)du = G(t, \tau), \qquad 0- < t, \tau < T+$$

where K is a rational kernel and G is suitably specified, is briefly outlined. Specific results in the case of RC and LRC noise kernels, with G correspondingly the difference of two (different)

RC or LRC covariance functions, are also given.

Microwave Theory and Techniques

Vol. MTT-9, No. 2, March, 1961

Dynamic Interaction Fields in a Two-Dimensional Lattice—R. E. Collin and W. H. Eggimann (p. 110)

In the theory of artificial dielectrics and aperture coupling in rectangular waveguides, a knowledge of the dynamic interaction fields is required in order to evaluate the polarizing fields. This paper presents suitable methods for evaluating the dynamic interaction fields in a two-dimensional lattice. Both electric and magnetic dipoles are considered. The results are presented in closed form apart from correction terms involving rapidly converging series. Cross-polarization interaction constants are also evaluated.

TEM Impedance and Cross Coupling for Small Circular Conductors in a Double Ridged Waveguide—J. E. Storer and T. W. Thompson (p. 116)

The even and odd mode TEM impedances and cross-coupling coefficient were found for two small circular center conductors in a double ridge waveguide structure. Expresssions were found by the use of a variational approximation for the case where the centers of the circular conductors lie on the horizontal center line of the guide; the conductors were placed symmetrically about the vertical plane of symmetry of the guide, and the conductors were placed a reasonable distance from the guide and from the region between the ridges. Results calculated from these expressions agree reasonably well with experimental data.

The experimental and theoretical results tend to indicate that proper placement of the two conductors in a double ridge guide could be used as a method of transmitting three different messages inside a single closed waveguide.

Design of a Coaxial Hybrid Junction—L. Stark (p. 124)

The design of a coaxial hybrid junction is discussed. The hybrid consists of a shunt junction and a series junction. The shunt junction is a broad-band stub compensated tee, and the series junction is basically a balun of the type used to excite a slotted dipole. There is inherent isolation between the shunt and series terminals. The useful bandwidth of the hybrid is at least 10 per cent, while the bandwidth of the shunt junction alone exceeds this by a factor of four.

Design data are presented for frequency bands centered at 425 Mc and 220 Mc. Many of these hybrids have been manufactured for application, and the performance repeats very well. Performance data are given for VSWR, isolation, and peak power capacity.

Step-Twist-Junction Waveguide Filters—B. C. De Loach, Jr., (p. 130)

The properties of step-twist-junction discontinuities in rectangular waveguide are considered. Methods are presented whereby these step-twist junctions may be used in filter design and, in particular, in the design of variable bandwidth constant-resonant frequency filters.

Resonators for Millimeter and Submillimeter Wavelengths—William Culshaw (p. 135)

Further considerations on the mm-wave Fabry-Perot interferometer are presented. Computed Q values for parallel metal plate resonators indicate that at spacings around 2.5 cm, values ranging from 60,000 at 3 mm to 300,000 at 0.1 mm wavelengths, are possible. The plates must, however, be quite flat. These results are important for many investigations, and in particular for mm and sub-mm wave maser research. For the aperture per wavelength ratios possible here, diffraction effects should be small. Consideration is given to using curved reflectors or focused radiation in applications where the fields must be concentrated. For this purpose, re-entrant conical spherical resonators are treated in detail, as regards operation in the TEM mode at high orders of interference. Expressions for the Q and shunt impedance are given, and high values are possible at mm and sub-mm wavelengths. Quasi-optical methods of coupling into and out of such a resonator are proposed, and the higher modes possible in such a resonator are considered. Results indicate that it could have application to the mm-wave-generation problem, and that it represents a good resonant cavity for solid-state research at mm and sub-mm wavelengths, and for maser applications in particular.

A Recording Microwave Spectrograph—D. Ilias and G. Boudouris (p. 144)

The principle of operation and the fundamentals of realization of a recording microwave spectrograph designed for use in the study of the absorption and the index of refraction of gases under medium pressures (1 mm Hg to 1 atm) are presented. The apparatus results from a similar spectrograph with synchroscope, in which the responses of the cavity resonators are interpreted by means of a pulse method. The high performances of the apparatus render its use advantageous, not only as a spectrograph, but also as an accurate recording refractometer, as well as a direct-reading Q meter.

A Cavity-Type Parametric Circuit as a Phase-Distortionless Limiter—F. A. Olson and G. Wade (p. 153)

This paper is a study of the properties of a diode parametric frequency converter (negative-conductance type) when used to perform microwave limiting. Unlike the parametric amplifier, the output power of a converter cannot exceed a certain level, regardless of the amplitude of the input signal. Thus, the ability to limit is a fundamental property of regenerative parametric frequency converters. An experimental limiter circuit, consisting of two stages of parametric frequency conversion, provided an output which was constant to within ± 1 db over a range of input of 50 db, and had 10-db small-signal gain. The phase variation was less than seven degrees over the entire range of input power.

A Stripline Frequency Translator—Elisabeth M. Rutz (p. 158)

A frequency translator is discussed which operates at C-band frequencies. The modulators in the frequency translator are crystal diodes, and modulation is obtained by periodic variation of the reflection characteristic of the crystal modulators. The conversion loss of the frequency translator is 6.5 db at 8-mw input power. The unwanted sidebands are at least 25 db below the translated signal.

An Approximate Solution to Some Ferrite Filled Waveguide Problems with Longitudinal Magnetization—Sheldon S. Sandler (p. 162)

An approximate solution for the field structure and propagating modes in parallel plane, circular, and coaxial ferrite-filled waveguide is presented. Bundles of plane waves are assumed to propagate in these structures which bounce back and forth along the guide. The solutions are classified into two types depending on the negative or positive equality of the incident and reflected waves. In the case of the circular guide the waves form a cone, and in the coaxial guide they form a frustum of a cone about the axis. The elemental plane waves are also assumed to satisfy Polder's relation and the boundary conditions at the guide walls. Simple relations are obtained with this equivalence for the propagation constant and the field. Comparison to rigorous theory is made in the case of the parallel plane and circular guide. Some experimental verification is presented for the completely-filled coaxial waveguide.

Synthesis of Low-Reflection Waveguide Joint Systems—P. Foldes and N. Gothard (p. 169)

Some characteristics of flat-flange type joints are analyzed. Experimental evidence is given proving that it is possible to reproduce, in practice, waveguide joints which have identical complex-reflection coefficients. Such joints can be combined to form large joint systems by means of a synthesis method, which keeps the over-all reflection coefficient to a minimum. Both theory and experimental data are presented.

A Channel-Dropping Filter in the Millimeter Region Using Circular Electric Modes—E. A. J. Marcatili (p. 176)

A channel-dropping filter in the millimeter region that transfers $TE_{01}○$ to $TE_{10}□$ is described and analyzed. The important features are the use of $TE_{011}○$ mode in the resonant cavities combined with a mode-selective coupling between circular symmetric and rectangular waveguides which make both heat loss and mode conversion low. Design formulas and experimental results on a model filter centered at 56 kMc are included. Finally, several possible mode transducers and filters based on the idea of mode-selective coupling are described.

Coupled-Mode Description of Crossed-Field Interaction—J. E. Rowe and R. Y. Lee (p. 182)

The coupled-mode theory is developed for two-dimensional M-type flow, and a system of five coupled-mode equations is obtained. A fifth degree secular equation is found for the perturbed propagation constants of the system. Under weak space-charge field conditions, both the forward-wave and backward-wave interactions may be described in terms of only two coupled modes. The two-mode theory is applied to the calculation of starting conditions for the M-BWO, and to the M-FWA. The conditions for beating-wave amplification are determined, and the variation of the mode amplitudes with distance is given.

Broad-Band Cavity-Type Parametric Amplifier Design—Kenneth M. Johnson (p. 187)

This paper tells how maximum bandwidth can be obtained from a nondegenerate parametric amplifier which utilizes a circulator. Expressions are derived for the gain bandwidth product and maximum possible gain bandwidth product. It is then shown how the Q of the cavities used for the signal and idler circuits may be kept at a minimum without degrading the noise performance of the amplifier. It is shown that best performance results when the TEM mode is used in coax, or, if waveguide is used, when the operating frequency is far away from the waveguide cutoff frequency. The diode used should have as high a self-resonant frequency as possible and the line admittance should be approximately the diode susceptance. Using a diode with a self-resonant frequency at the idler frequency will be seen to give optimum performance.

This paper also discusses double tuning the signal circuit to achieve broader bandwidths. In this case, the addition of the second tuned circuit will be seen to give much broader bandwidths than one would expect from conventional filter theory.

Two sample amplifiers are considered and their bandwidths calculated. The effect of double tuning one of the amplifiers is then considered.

Nuclear Science

Vol. NS-8, No. 2, April, 1961

Undesired Background Subtraction in a Logarithmic Count Rate Meter Circuit—Harold E. DeBolt (p. 1)

Background compensation can be accomplished by simple subtraction of a background signal in the linear count rate meter. However, in the logarithmic count rate meter circuit (Cooke-Yarborough type), compensation must be accomplished by subtraction from each section of the logarithmic circuit before summing the outputs of the sections. The method is described and mathematically justified.

A Mathematical Method of Analysis for Predicting the Performance of an Anti-Compton Gamma Ray Spectrometer—J. S. Rosen, J. C. Whiton, and C. W. Hill (p. 4)

A mathematical model has been developed for estimating the performance of an Anti-Compton gamma ray spectrometer. This spectrometer consists of a small cylindrical crystal surrounded by a larger, hollow cylindrical crystal, both composed of NaI(Tl). The results of a two-parameter study—incident gamma ray energy and relative crystal positions—are given.

An Electrostatic-Magnetic Separator for Obtaining Isotopes of High Purity—F. A. White, F. M. Rourke, J. C. Sheffield, and L. A. Dietz (p. 13)

The design of a 30-inch-radius combined velocity and direction focusing mass analyzer is described. Application of this instrument as a high purity isotopic separator for research in nuclear physics, metallurgy, and semiconductors is also discussed. Brief mention is made of the development of *p-n* junctions for detecting singly-charged ions of low kinetic energy.

Administrative Committee Meeting (p. 22)

International Electrotechnical Commission. Report of Meeting of Technical Committee TC/45 (p. 23)

Eighth Annual PNGS Meeting—International Symposium on Aerospace Nuclear Propulsion (p. 24)

Contributors (p. 25)

Reliability and Quality Control

Vol. RQC-10, No. 1, March, 1961

Safety Margins Established by Combined Environmental Tests Increase Atlas Missile Component Reliability—C. C. Campbell (p. 1)

A reliability test program known as a "Search for Critical Weaknesses" has been put into operation by the Convair-Astronautics Reliability Organization. The function of this program is to direct critical component weaknesses so that corrective action can be taken before operational failure occurs. Components selected for testing are subjected to combined environmental severities at and beyond the design requirements. This is done in order to establish a margin of safety and to be able to reassess component reliability in the event that actual flight environments differ from those expected. After completing the first three phases of this program, 203 component types consisting of 980 individual specimens have been evaluated. A fourth phase in which 105 component types are being tested is nearing completion, and a fifth phase has been initiated.

This paper covers the testing procedures and the techniques used in determining the relative component reliability during the first three phases of the testing program. In addition, an explanation of representative components' weaknesses was revealed during the tests, and the corrective action taken has been included.

The Electronic Component Reliability Center—An Evaluation of the First Year's Operation—J. R. Funk (p. 7)

Effect of Circuit Design on System Reliability—J. J. Suran (p. 12)

It is shown that circuit drift failures may be eliminated by worst-case design procedures but that a considerable price is paid for this immunity in the form of increased system complexity, increased component stresses and increased power demand. Consideration of the entire problem leads to the conclusion that decreasing the probability of circuit drift failures (by increasing the tolerance margin of the circuit) tends to increase the probability of component catastrophic failures and that consequently an optimum component tolerance design point exists for maximum system reliability. The optimum tolerance margin depends upon the specific system and generally varies inversely with the number of components comprising the system. Thus, to maintain a specified system reliability in the face of increasing system complexity, it is necessary to assume a decreasing component parameter spread (tighter tolerances) and a decreasing component catastrophic failure rate. Both of these requirements may be relaxed if some form of redundancy is introduced to overcome the inevitable occurrence of catastrophic failures.

Measuring Missile Reliability in Prelaunch Environments—David S. Stoller (p. 19)

This paper examines some of the problems involved in measuring missile reliability during its prelaunch phase. This is an important problem for three reasons: 1) a missile typically is exposed to many kinds of operating environments over long periods of time before it is launched; 2) a missile's reliability history in its prelaunch operating environments is an index of its in-flight reliability; and 3) the major cost of a missile weapon system derives from the consequences of its prelaunch reliability behavior.

Most of the emphasis to date in reliability measurement efforts has been focused on the in-flight environment, to the near-exclusion of the important prelaunch environment. Attempts to infer prelaunch reliability on the basis of data collected for other purposes, such as supply transactions, have shown that this is not an adequate substitute for direct measurements. This document provides an approach to the problem specifically oriented to the collection and analysis of reliability data.

Reliability measurement requirements stem directly from the basic definition of reliability: the probability that an aggregate will successfully perform a specified task in a specified environment over a specified period of time. Several of the most important prelaunch operating environments are: turn-on (-off), check-outs, countdowns, operating alert, standby alert, handling, transportation, periodic maintenance and storage.

Monitoring the reliability history of small aggregates, such as pumps, is most important during the research, development, and early operational phase of weapon-system growth when extensive redesign can be achieved, but it becomes less important and even a burden later on when design effort is concentrated on the next-generation weapon. When the system is in full operation it is much more important to monitor the large aggregates, such as ground guidance stations, so as to gain knowledge of operational effectiveness. However, since it is still important to have some knowledge of the reliability behavior of the small aggregates in the later phase, the reliability measurement policy given below merits consideration:

1) define standard operating environments for large aggregates,
2) relate the operating environments of smaller aggregates to the standards of the large aggregates,
3) monitor only the large aggregates,
4) synthesize the reliability history of small aggregates from 3),
5) relate standard and synthesized environments to design environments by bench test.

A reliability measurement system, then, needs flexibility in form so that environments and criteria can be changed readily. It should also interact with the operations, maintenance, and supply data system so as to avoid redundancy of data collection, since many of the data elements required for reliability measurements are also required for other operations and materiel purposes. A missile status log, based on the principles of weapon-centered event recording, can be used to obtain reliability data in accordance with the considerations above.

The log obtains data in the following five categories:

1) time reference of event causing entry,
2) impact of event on weapon-system effectiveness,
3) consequences of event on operations-maintenance status,
4) identification of aggregates,
5) remarks for local use.

Finally, some of the problems arising in the classification of malfunctions, configuration control, data discipline, and data volume are discussed briefly.

The Effect of the NOL and Battelle Data Interchange Programs on Librascope Reliability Test Efforts—Leonard G. Rado (p. 26)

Due to the emphasis placed on the interchange of test data and also on the attempt to give the defense effort the best possible test information, two particular programs have been promulgated: The NOL Component Reliability History Program (FBMWS) and the efforts of the Battelle Memorial Institute's Electronic Component Research Center which have been in existence for several years now.

This paper describes the different aspects of each program and the mechanics of each. The NOL Program is for the use of Fleet Ballistic Missile Weapons System (FBMWS) contractors and the Battelle Program is funded by contritions to the members.

This paper also indicates how Librascope, as a participating member with a small evaluation group, gleans the most results from membership in these two programs.

A Measure of Reliability and Information Quality in Redundant Systems—S. A. Rosenthal, H. Jaffe, and M. D. Katz (p. 29)

In the development of complex military electronic equipment, it is necessary to consider the effects of many interacting operational and design factors. For both the over-all weapon system, and for the electronic subsystems with which we are concerned, optimum trade-offs for such interdependent parameters as reliability, weight, accuracy, maintainability, cost, mission time, and target vulnerability must be determined. Considered individually, these parameters do not adequately describe the operational worth of equipment incorporating multiple redundancies and operating modes. Operational worth is the probability that a system will achieve success in performing its required functions, and is obviously a prime consideration in the planning and selection of systems for military application.

The inadequacy in describing operational worth in terms of its individual parameters was encountered in the synthesis of an optimum redundant configuration for the B-58 bombing-navigation system. To resolve this difficulty, operational worth models describing the interaction between parameters were developed and utilized by Sperry Gyroscope Company. These techniques enabled Sperry to design this bombing-navigation system to have a high operational worth and an inherent mission reliability in excess of 95 per cent. In this paper, the model for "system worth," the interaction between reliability and information quality is discussed.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*), London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The 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* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) and included in the March, 1961 issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. 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.

ACOUSTICS AND AUDIO FREQUENCIES

534.2-14 **1382**
Transient Sound Propagation in a Layered Liquid Medium—J. W. C. Sherwood. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1673–1684; December, 1960.) A theoretical analysis is given of transient sound propagation between source and receiver located at arbitrary points in a horizontally stratified liquid.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears at the end of the Abstracts and References section.

534.21 **1383**
Propagation of Sonic and Ultrasonic Waves in Natural Waveguides—L. Brekhovskikh. (*Usepekhi. Fiz. Nauk*, vol. 70, pp. 351–360; February, 1960.) A discussion is given of sound propagation to very great distances in the sea and in the atmosphere by a natural waveguide action.

534.24-14:534.88 **1384**
Fluctuations of Sound Reflected from the Sea Surface—C. S. Clay. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1547–1551; December, 1960.) Data obtained from an experiment in which both source and receiver are at a depth of about 3000 feet (1385 below) are compared with calculations based on the theory of Eckart (2541 of 1953).

534.24-14:534.88 **1385**
Fluctuations in Surface-Reflected Pulsed C.W. Arrivals—M. V. Brown and J. Ricard. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1551–1554; December, 1960.) A 168-cps signal was transmitted at a depth of 2880 feet, and both direct and surface-reflected components of the signal received at a depth of 3000 feet were analyzed for fluctuations as a function of range and angle of incidence. A $(\cos\phi)^{2.5}$ relation was found between the relative standard deviation of the energy and the incident angle, as measured from the normal.

534.286.2 **1386**
The Influence of Edges on the Sound Absorption of Porous Materials—W. Kuhl. (*Akust. Beih.*, no. 1, pp. 264–276; 1960.) Measurements of sound absorption were made on various arrangements of absorber plates in a reverberation chamber. The results are discussed with reference to the work of other authors to assess the magnitude of the edge effect under different conditions.

534.512.2:538.566:535.31 **1387**
Electromagnetic Reflection from Sound Waves—H. J. Schmitt and T. T. Wu. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1660–1667; December, 1960.) The reflection of EM waves normally incident on the wavefronts of a semi-infinite standing sound wave in a liquid is discussed theoretically.

534.6:534.322.3 **1388**
Development of a Noise Generator for Producing Noise of Constant Spectral Bandwidth with any Desired Centre Frequency and Constant-Ampli ude T:me Functions—H. Niese. (*Nachrtech.*, vol. 10, pp. 196–204; May, 1960.) The noise generator described provides bands of FM noise of bandwidth 40, 100, 200 and 400 cps with the center frequency adjustable over the range 20 cps-17 kc. See also 8 of January.

534.7 **1389**
Electro-acoustics for Human Listeners—C. Cherry. (*J. Brit. IRE*, vol. 21, pp. 5–15; January, 1961. Discussion.) A review of existing knowledge of electro-acoustics in relation to the listener is given. The present theories of stereophonic hearing are shown to be incomplete and a plea is made for further study.

534.78 **1390**
Speech Processing by the Selective Amplitude System—L. R. Spogen, H. N. Shaver, D. E. Baker, and B. V. Blom. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1621–1625; December, 1960.) A system is described in which sampling times are determined from the speech waveform; the average rate is considerably less than that required for periodic sampling.

534.79 **1391**
A Method for the Calculation of Loudness—E. Zwicker. (*Akust. Beih.*, no. 1, pp. 304–308; 1960.) A simplified graphical method is described for evaluating loudness levels from third-interval level diagrams. See 3349 of 1960.

534.79 **1392**
The Difference between the Curves of Equal Loudness for a Plane Wave and a Diffuse Sound Field—G. Jahn. (*Hochfrequenz. und Elektroak.*, vol. 69, pp. 75–81; April, 1960.) Objective measurements were made to determine the influence of the shape of the sound field on loudness sensation; these measurements are based on the results given in 740 of 1960. A comparison is made with the loudness curves obtained by subjective tests (1844 of 1960) and relatively good agreement is found.

534.844.1+[621.317.2:538.566.08 **1393**
A New Reverberation Chamber for Sound and Electromagnetic Waves—E. Meyer, G. Kurtze, H. Kuttruff, and K. Tamm. (*Akust. Beih.*, no. 1, pp. 253–264; 1960.) The acoustic reverberation time of the room described is 33 sec at 100 cps and 13 sec at 1 kc; the sound insulation results in about 80-db attenuation. The chamber is internally lined with copper foil and the EM wave reverberation time is 400 μsec at 10 Gc without scattering objects; this corresponds to a Q-factor of 1.7×10^6.

534.845 **1394**
Design of Acoustic Resonator Panels—C. R. Nagaraja Rao. (*J. Sci. Ind. Res.*, vol. 19A, pp. 506–509; October, 1960.) A design

procedure based on the Helmholtz formula has been verified experimentally for sound waves normally incident on the panel.

534.845.1 1395

Determination of Sound Absorption Coefficients using a Pulse Technique—C. L. Rogers and R. B. Watson. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1555–1558; December, 1960.) A laboratory method is described by means of which sound absorption coefficients can be determined as a function of angle of incidence using a loudspeaker at the focus of a paraboloidal sound mirror.

621.395.6:621.372.44 1396

Exact Solution of a Time-Varying Capacitance Problem—Macdonald and Edmondson. (See 1432.)

621.395.61 1397

Absolute Determination of the Response of Microphones in a Diffuse Sound Field—H. G. Diestel. (*Akust. Beih.*, no. 1, pp. 277–280; 1960.) An application of the reciprocity method in a diffuse sound field is given. The calibration of a condenser microphone for the range 500–16000 cps in a reverberation chamber is described.

621.395.61 1398

Determination of the Pressure Transmission Factor of Microphones—H. G. Diestel and H. Mrass. (*Arch. tech. Messen*, no. 291, pp. 65–68; April, 1960.) A review of methods and apparatus used for the pressure calibration of microphones is given. About 40 references.

621.395.623.8 1399

'Loudspeaker with Increased Degree of Presence'—H. Harz. (*Elektron. Rundschau*, vol. 14, pp. 193–197; May, 1960.) A description is given of an array of loudspeakers for use in broadcast studios. Facilities are provided for the adjustment of the ratio of direct to diffuse sound in order to achieve a greater degree of realism.

621.395.625.3 1400

Investigation of the Fluctuations of Synchronism in a Magnetic Tape Recorder with the Aid of Electromechanical Analogues—W. Wolf. (*Hochfrequenz. und Elektroak.*, vol. 69, pp. 41–52; April, 1960.) The equivalent circuit of a mechanical tape transport system is derived. The electrical analogs of the various mechanical parameters are evaluated and their application in improving the design of tape transport mechanisms is discussed.

621.395.625.3:538.221 1401

Particle Interaction in Magnetic Record Tapes—Woodward and Della Torre (See 1590.)

ANTENNAS AND TRANSMISSION LINES

621.372.8:621.317.39:531.71 1402

Measurement Method for Determining the Internal Dimensions of Waveguides—J. Bachel. (*Frequenz*, vol. 14, pp. 131–134; April, 1960.) The method described is suitable for long waveguide, of metal or internally metallized at least at the points of measurement; dimensions can be measured to within $\pm 1\,\mu$ sec. The underlying principle is that of a measurement of capacitance between one inner wall and a metal plate accurately positioned on and insulated from the opposite inner wall. The method is also applicable to circular waveguide. Test procedure and results are discussed.

621.372.8.09 1403

Transport of Angular Momentum in a Waveguide—G. Toraldo di Francia. (*Alta Frequenza*, vol. 29, pp. 148–153; April, 1960.) In a waveguide where the field consists of two or more superposed degenerate modes, the angular momentum travels at the same velocity as the energy. In the case of two nondegenerate modes, the momentum has a velocity equal to the arithmetic mean of the group velocities of both modes.

621.372.81 1404

The Theory of Waveguides and Cavities: Part 1—A General Approach to the Exact Theory of Waveguides and Cavities—R. A. Waldron. (*Electronic Tech.*, vol. 38, pp. 98–105; March, 1961.)

621.372.822 1405

The Rotation of Polarization with the Twist in a Rectangular Waveguide—K. Schnetzler. (*Frequenz.*, vol. 14, pp. 123–126; April, 1960.) The plane of polarization is rotated only through about $\frac{1}{3}$ the angle of twist of a rectangular waveguide. In a waveguide carrying two waves of differing polarization, the coupling from one to the other depends only on the relative twist between input and output cross sections; for 1° twist this amounts to 37 db.

621.372.823:621.372.832.6 1406

A Circular-Electric Hybrid Junction and some Channel-Dropping Filters—E. A. Marcatili. (*Bell. Sys. Tech. J.*, vol. 40, pp. 185–196; January, 1961.) A description of a telescopically mounted Riblet short-slot coupler adjusted for 3-db power division at 55 Gc with a bandwidth of 20 per cent, a balance of 0.5 db and isolation better than 23 db is given. Application to a circular waveguide communication system in conjunction with mode conversion filters (1417 below) is considered.

621.372.823:621.372.852.1 1407

Band-Splitting Filter—E. A. Marcatili and D. L. Disbee. (*Bell Sys. Tech. J.*, vol. 40, pp. 197–212; January, 1961.) A splitting transition between two adjacent sub-bands covering one octave at mm λ is brought about within 160 Mc by a constant-resistance filter containing an elbow and a hybrid junction operating with 2-inch-diameter waveguide.

621.372.823:621.372.852.5 1408

Mode Exciters in Circular Waveguides—K. Noda. (*Rev. Elec. Commun. Lab., Japan*, vol. 8, pp. 465–476; September/October, 1960.) Resonant-iris-type exciters have been developed for several kinds of modes. They have the advantage of being small in size and weight, but have a restricted bandwidth.

621.372.823:621.372.853.1 1409

Circular Electric Wave Transmission through Hybrid-Mode Waveguide—K. Noda. (*Rev. Elec. Commun. Lab., Japan*, vol. 8, pp. 426–464; September/October, 1960.) "General formulae for mode generation from distributed sources are derived for hybrid-mode waveguides. Using the formulae, the mode conversion by irregular parts of the dielectric layer in a dielectric-coated waveguide is calculated. Coupling between modes at geometrical imperfections and in intentional bends are described as the relations between the normal hybrid modes. The coupling coefficients between the TE_{01} and HY_{1m} modes in the curved helix waveguide are obtained as functions of the arbitrary wall impedance. Several other transmission characteristics in the hybrid-mode waveguide are also analysed."

621.372.823.2 1410

The Transmission Characteristics of H_{01} Waveguide with Statistically Distributed Irregularities—H. Larsen. (*Frequenz*, vol. 14, pp. 135–142; April, 1960.) The influence of random distribution of axial curvature and cross-sectional deformation in circular waveguide on propagation of desired and undesired modes is investigated using autocorrelation functions. Examples are given of the calculation of dimensional tolerances using the Gaussian distribution as a basis for the autocorrelation function.

621.372.829 1411

Noncylindrical Helix Waveguide—H. G. Unger. (*Bell Sys. Tech. J.*, vol. 40, pp. 233–254; January, 1961.) Two cases are analyzed: a) random imperfections in the cross section defined in terms of correlation distance, and b) uniform imperfections. Random imperfections lead to mode conversion and the resulting loss is similar to that obtained without the helix, but uniform imperfections give considerably greater loss with the helix than without.

621.372.829 1412

Normal Modes and Mode Conversion in Helix Waveguide—H. G. Unger. (*Bell Sys. Tech. J.*, vol. 40, pp. 255–280; January, 1961.) An analysis of unwanted modes and their coupling coefficients due to guide curvature and deformation leads to consideration of mode-absorbing filters consisting of lengths of circular helical guide, and the effects of random curvature and random ellipticity.

621.372.831.4 1413

The 3-dB Coupler—W. Stösser. (*Frequenz*, vol. 14, pp. 117–121; April, 1960.) The operation of a short-slot hybrid junction is explained and its characteristics are derived. A method of reducing the physical length of such couplers is outlined and design features of a 3-db coupler are discussed.

621.372.831.6 1414

The Design of Two-Step Transforming Sections for Waveguide—W. Haken. (*Frequenz*, vol. 14, pp. 126–131; April, 1960.) A method is given for the design calculation of two-step sections coupling rectangular waveguides of different height but equal width with minimum reflection in a given frequency band. A two-stage transformer for the range 3.8–4.2 Gc was designed by this method and, by means of certain mechanical corrections, a section was obtained whose measured reflection coefficient agrees closely with the theoretical optimum.

621.372.852.323 1415

Experimental Investigations of Ferrite Resonance Isolators—R. Steinhart. (*Nachrtech. Z.*, vol. 13, pp. 183–191; April, 1960.) Measurements of directional attenuation and phase shift have been made with reference to the theory given earlier (411 of February), and the influence of ferrite dimensions on isolator characteristics has been investigated systematically. A method of self-compensation of temperature effects is described which is adequate up to 70°C.

621.372.852.4 1416

A Polarization Filter with Symmetric Excitation of H_{11} Waves—E. Schuegraf. (*Frequenz*, vol. 14, pp. 121–123; April, 1960.) In the polarizer described, the two H_{11} modes are symmetrically excited in a circular waveguide; this method extends the usable range of frequencies, and does not give rise to undesirable modes such as E_{01}.

621.372.852.5 1417

Mode-Conversion Filters—E. A. Marcatili. (*Bell Sys. Tech. J.*, vol. 40, pp. 149–184; January, 1961.) The theory and construction of low-loss filters that simultaneously drop channels and transfer from circular to rectangular waveguide modes is described. The advantages over other methods for the mm λ range are discussed.

621.396.67 1418

Wide-Band Omnidirectional Radiators of

Cylindrical Symmetry with High-Pass Matching—H. Meinke. (*Nachrtech. Z.*, vol. 13, pp. 161–168; April, 1960.) Measurements of input impedance and reflection coefficient were made in the frequency range 200–4000 Mc on omnidirection radiators of differing shape. The input impedance was calculated for lower frequencies and the lower cutoff frequency is determined. The optimum shape of a wide-band radiator is a cone with smoothly rounded edges using a reflection-free feed arrangement (see 2999 of 1960).

621.396.676:621.317.332:621.398 1419
Measuring Antenna Impedance in the Ionosphere—O. C. Haycock and K. D. Baker. (*Electronics*, vol. 34, pp. 88–92; January 13, 1961.) Both resistive and capacitive components are measured continuously from the standing-wave pattern set up on an artificial transmission line terminated by the antenna.

621.396.677:621.396.62 1420
Designing Low-Noise Antennas—R. Caldecott and W. H. Peake. (*Electronics*, vol. 34, pp. 60–63; January 20, 1961.) With the advent of very-low-noise receivers, the noise emphasis must be shifted to the receiving antenna. The effects of mismatch, attenuation, sidelobes and spill-over are considered. The optimum antenna is not necessarily the one with maximum gain.

621.396.677.32 1421
Endfire Antennas—G. Broussaud and E. Spitz. (Proc. IRE, vol. 49, pp. 515–516; February, 1961.) The antenna, consisting of a two-wire line surrounded by a helix, has a gain of 18–20 db between 2 and 5 Gc with secondary lobes below 30 db. See also 2236 of 1960 (Wickersham).

621.396.679.4 1422
Feeders for Decimetre-Wave Aerials of High Power Rating—H. Laub and S. Stöhr. (*Frequenz*, vol. 14, pp. 144–155; April, 1960.) The transmission characteristics of long feeders consisting of flexible or rigid coaxial lines, waveguide or surface-wave transmission line are discussed. The results are given of theoretical and experimental investigations of attenuation and reflection coefficient on long runs of rectangular waveguide constructed from numerous identical elements. Waveguide components for use in antenna feeders are described. Results are also included of measurements on feeders of the Goubau-line type. 36 references.

AUTOMATIC COMPUTERS

681.142 1423
A Multipurpose Computer Element—C. B. Taylor. (*Electronic Engrg.*, vol. 33, pp. 96–99; February, 1961.) A "long-tailed pair" waveform reshaper is described which, in conjunction with a number of logical gates, can be used as a basic element in computer design.

681.142:621.382.323 1424
Microcircuit Binary Full Adder uses Unipolar Transistors—M. E. Szekely, J. T. Wallmark, and S. M. Marcus. (*Electronics*, vol. 33, pp. 48–49; December 23, 1960.) Unipolar transistors are used for both active and passive elements.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.7 1425
Circuit Considerations relating to Microelectronics—J. J. Suran. (Proc. IRE, vol. 49, pp. 420–426; February, 1961.) A discussion of the problems associated with microminiaturization, especially power dissipation and its relation to circuit functions and packing density, is given.

621.314.2:621.391.822 1426
Barkhausen Noise in Transformer Cores—K. G. Warren. (*Electronic Tech.*, vol. 38, pp. 89–94; March, 1961.) The following are discussed: a) isolation and measurement of the noise, b) its relation to the slope of the hysteresis loop and the cycling rate, c) an equivalent circuit which includes the noise source, and d) the frequency spectrum of the noise when sinusoidally polarized.

621.318.57:621.382.23 1427
Margin Considerations for an Esaki-Diode:Resistor OR Gate—H. K. Gummel and F. M. Smits. (*Bell Sys. Tech. J.*, vol. 40, pp. 213–232; January, 1961.) Worst-case margin considerations imply that applications are restricted to storage systems, flip-flops, shift registers and the like.

621.319.43 1428
A Helix-Type Variable Capacitor—Y. Yamamoto. (Proc. IRE, vol. 49, pp. 522–523; February, 1961.) Maximum-to-minimum capacitance ratios of 10,000 or more are obtainable.

621.372.41 1429
Two-Terminal-Network Functions through Prescribed Points. Additions to the Interpolation Problem of Pick and Nevanlinna—H. Frank. (*Nachrtech. Z.*, vol. 13, pp. 241–243; May, 1960.) The synthesis of two-terminal networks consisting of resistive, inductive and capacitive elements and having a given impedance function is considered.

621.372.412:621.372.54 1430
Quartz AT-Type Filter Crystals for the Frequency Range 0.7 to 60 Mc/s—R. Bechmann. (Proc. IRE, vol. 49, pp. 523–524; February, 1961.) Design data are given for suppressing the unwanted modes to at least 40 db below the main mode. See also 1899 of 1960.

621.372.413 1431
Design of Cavity Resonators with Maximum *Q*-Factor—W. Otto. (*Nachrtech.*, vol. 10, pp. 205–209, 266–272, and 365–372; May, June, and August, 1960.) A summary of theory and design formulas for cylindrical cavity resonators with details of constructional features and coupling arrangements is given.

621.372.44:621.395.6 1432
Exact Solution of a Time-Varying Capacitance Problem—J. R. Macdonald and D. E. Edmondson. (Proc. IRE, vol. 49, pp. 453–466; February, 1961.) A new method is developed giving a closed-form solution for the harmonics generated by a sinusoidally varying capacitance in series with a fixed resistor and battery. Practical applications are discussed with reference to the condenser microphone, the vibrating-reed electrometer and a special loud-speaker improvement.

621.372.5 1433
The Quadratic Invariances of a Generalized Network—M. C. Pease. (Proc. IRE, vol. 49, pp. 488–497; February, 1961.) In systems where the matrix operator has eigenvalues which are unity or can be taken in reciprocal conjugate pairs, the number of invariant quadratic forms is at least equal to the number of degrees of freedom.

621.372.54 1434
The Imaginary Part of the Characteristic Impedance in the Pass Band of Filter Circuits—W. Herzog. (*Nachrtech. Z.*, vol. 13, pp. 179–182; April, 1960.) The use of filter circuits with complex characteristic impedance in the pass band extends the number of possible bridge-type filters. Filters with variable pass band can be realized.

621.372.54 1435
A Quick Method for the Design of Butterworth Filters—J. S. Bell and P. G. Wright. (*Electronic Engrg.*, vol. 33, pp. 106–108; February, 1961.) A design procedure and worked example are given.

621.372.543.2 1436
A Variable-Bandwidth Band-Pass Filter—J. Holland. (*Electronic Engrg.*, vol. 33, pp. 100–105; February, 1961.) "A method is described whereby several bandwidths may be obtained in an *LC* filter employing one set of inductors only. The design of a typical filter with an actual plot of its attenuation characteristic is included."

621.372.543.2 1437
A Continuously Variable Band-Pass Filter for the Audio-Frequency Range—W. Ohme. (*Frequenz*, vol. 14, pp. 182–186; May, 1960.) The design of a band-pass filter whose bandwidth and lower cutoff frequency are continuously variable is described. The requisite crystal filters are derived from the band-pass elements given by image-parameter theory. The equipment and some of its response curves are illustrated.

621.372.63 1438
Synthesis of *N*-Port Active RC Networks—I. W. Sandberg. (*Bell Sys. Tech. J.*, vol. 40, pp. 329–347; January, 1961.) It is proved that N is the sufficient and, in general, minimum number of controlled sources required to realize an arbitrary $N \times N$ matrix of real rational functions as a transformerless active RC N-port network.

621.373.421.1:539.2:538.569.4 1439
The Hamiltonian Formalism of Damping in a Tuned Circuit—K. W. H. Stevens. [*Proc. Phys. Soc.* (*London*), vol. 77, pp. 515–525; February, 1961.] An LC circuit coupled to a transmission line is studied as an example of introducing damping into the quantum-mechanical treatment of a harmonic oscillator. The problem of coupling a spin system to the current in the inductance is also studied.

621.373.43 1440
The Stability of Relaxation Oscillations Synchronized by Pulses—E. De Castro. (*Alta Frequenza*, vol. 29, pp. 206–231; April, 1960.) A finite-difference equation is derived for the pulse synchronization of relaxation oscillators. Solutions are obtained by a graphical method and possibilities of simplification are considered. The procedure is also applied to the case of frequency division. Calculated parameters are verified by measurements on a multivibrator circuit.

621.373.431.1 1441
Description of the Change-Over Process in Multivibrators—G. Kohn. (*Arch. elekt. Übertragung*, vol. 14, pp. 193–203; May, 1960.) The dynamic processes occurring in multivibrators are described using a graphical method. A bistable circuit is considered as example, and a system of curves representing all possible internal processes is derived on the basis of differential equations for the characteristic time functions of this circuit. The influence of the duration and amplitude of the triggering pulse on transition time is determined.

621.374:681.188 1442
Decoding Circuits with Semiconductor Elements—E. Oehme. (*Hochfrequenz. und Elektroak.*, vol. 69, pp. 52–61; April, 1960.)

Various types of circuits for binary decoding systems are given and requirements regarding the minimum number of semiconductor diodes and transistors to be used are evaluated and compared.

621.374.4:621.396 1443
The Problem of Frequency Synthesis—H. J. Finden. (*J. Brit. IRE*, vol. 21, pp. 95–103; January, 1961.) A review of the basic techniques employed for addition, subtraction and division in frequency synthesizers is given.

621.374.5:534.213–8:535.215 1444
Continuously Variable Glass Delay Line—H. A. Brouneus and W. H. Jenkins. (*Electronics*, vol. 34, pp. 86–87; January 13, 1961.) Stress waves are optically monitored by using the birefringent properties of glass under stress.

621.375.232.4 1445
Designing Grounded-Grid Amplifiers with Controlled Gain—J. W. Rush. (*Electronics*, vol. 33, pp. 50–53; December 23, 1960.) A six-stage low-noise IF amplifier with microminiature ceramic triodes is described.

621.375.4.018.783 1446
Review of Nonlinear Distortion in Transistor Stages, including Cross-Modulation—H. Lotsch. (*Arch. elekt. Übertragung*, vol. 14, pp. 204–216; May, 1960.) The nonlinear transfer function is approximated by a Taylor series expansion, and from this equations are derived for various types of distortion in earthed-emitter transistor circuits. Methods of compensating cross-modulation are considered. The results of cross-modulation measurements are found to be similar to those obtained by Akgün and Strutt (3034 of 1960).

621.375.423 1447
Selective RC Amplifier using Transistors—R. Hutchins. (*Electronic Engrg.*, vol. 33, pp. 84–87; February, 1961.) An investigation of the use of phase-shift networks in transistor circuits is discussed. The dual of the Wien-bridge arm is used to form an oscillator and a selective amplifier, which has constant gain while the selectivity and tuning are continuously variable.

621.375.432 1448
Equation for the Performance of a Nonlinear Transistor Amplifier—I. Gumowski, J. Lagasse, and Y. Sevely. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 1995–1997; March 14, 1960.) A first-order differential equation is derived to describe the performance of a nonlinear amplifier discussed earlier by Gumowski (4150 of 1960).

621.375.9:538.569.4 1449
Investigation of the Shift in Frequency and of the Amplitude of Oscillations of a Maser-Type Self-Oscillator with Liquid Flow—C. Fric. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 2353–2355; March 28, 1960.) An experimental investigation of the effect of detuning a maser oscillator of the type described earlier (798 of 1960), and of the influence of the rate of flow on the resonance line width is given. The range of oscillation extends from 1.7 kc (0.4 G) to 29.6 Mc (7 kG).

621.375.9:538.569.4 1450
Cross-Relaxation Phenomena in Solid-State Masers—S. A. Ahern, P. A. Gould, and J. C. Walling. (*J. Electronics and Control*, vol. 9, pp. 477–480; December, 1960.) With the direction of the applied magnetic field close to the *c*-axis of a ruby maser, amplification for different crystal orientations was confined to two narrow-frequency bands within the 2.8–3.2 Gc range used. Possible mechanisms concerned are spin-spin energy exchanges involving a) four ions, and b) seven ions.

621.375.9:621.372.44:537.311.33 1451
Three-Terminal Variable-Capacitance Semiconductor Device—J. L. Giacoletto. (PROC. IRE, vol. 49, pp. 510–511; February, 1961.) An analysis of the operation of the device which is based on the change in junction transition capacitance, associated with the change in collector voltage gradient, in an avalanche-controlled semiconductor amplifier is given.

621.375.9:621.372.44:621.382.23 1452
A Varactor-Diode Parametric Standing-Wave Amplifier—H. Brett, F. A. Brand, and W. G. Matthei. (PROC. IRE, vol. 49, pp. 509–510; February, 1961.) Details are given of experimental work on the type of amplifier discussed by Landauer (*ibid.*, vol. 48, pp. 1328–1329; July, 1960.).

621.376:621.382.23 1453
Using Voltage-Variable Capacitors in Modulator Design—A. C. Todd, R. P. Schuck, and H. M. Sachs. (*Electronics*, vol. 34, pp. 56–59; January 20, 1961.) Circuits and design equations are given for phase and frequency modulators and for SSB amplitude modulators.

621.376.23:621.316.8 1454
Radio-rectification and Detection by Simple Bilateral Nonlinear Resistors—J. E. Bridges. (PROC. IRE, vol. 49, pp. 469–478; February, 1961.) An analysis of ac/dc converting circuits using symmetrical nonlinear resistors, the operation of which may be largely independent of temperature, is given. Two applications are described: 1) for developing the hv focusing-potential bias for a picture tube, and 2) for improving the detection threshold of a photoelectric sensing device.

GENERAL PHYSICS

537.313:530.17 1455
Gauss's Principle of Least Constraint applied to Electrical Networks—A. von Weiss. (*Arch. elekt. Übertragung*, vol. 14, pp. 235–236; May, 1960.) Gauss's principle of mechanics, when adapted to electrical networks, leads to Kirchhoff's laws.

537.525 1456
Swept Langmuir Probe System for Intense Gas Discharges—H. W. Jones and P. A. H. Saunders. (*J. Sci. Instr.*, vol. 37, pp. 457–459; December, 1960.) Apparatus, including a low-impedance voltage sweep generator and a logarithmic amplifier, is described for measurement of electron temperatures up to $2\times10^{5\circ}$K and ion densities up to 5×10^{14}/cm³.

537.533:537.56 1457
Exploration of Oscillating Beam-Plasma Fields with a Transverse Electron Beam—A. Garscadden and K. G. Emeleus. (*J. Electronics and Control*, vol. 9, pp. 473–476; December, 1960.) The results appear to be compatible with considerable randomization and possible break-up of the main beam by the oscillations.

537.534:537.311.33 1458
Electrodynamics of the Image Force—F. Ollendorff. (*Arch. Elektrotech.*, vol. 45, pp. 169–187; April 11, 1960.) The processes of ion emission and absorption at semiconductor surfaces are analyzed. See 594 of February.

537.56 1459
Electron Density Fluctuations in a Plasma—E. E. Salpeter. (*Phys. Rev.*, vol. 120, pp. 1528–1535; December 1, 1960.) The spatial Fourier transform of the electron distribution in a plasma is obtained, together with its time variation, using assumptions consistent with an ionospheric application.

537.56 1460
Space-Charge Instabilities in Synthesized Plasmas—A. L. Eichenbaum and K. G. Hernqvist. (*J. Appl. Phys.*, vol. 32, pp. 16–21; January, 1961.) Potential transitions, dependent on boundary space-charge conditions, are shown to be possible by a theoretical analysis of an idealized model. Such transitions are observed experimentally. The voltage jumps are in good agreement with the theory.

537.56 1461
Method for the Determination of the Electron Density of a Plasma by Group Velocity—T. Consoli and D. Lepechinski. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 2694–2696; April 11, 1960.)

537.56 1462
Method of Measuring Simultaneously the Confining Magnetic Field and the Electron Density of a Plasma—T. Consoli and D. Lepechinski. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 2813–2815; April 20, 1960.) An extension of earlier work (1461 above) is given.

537.56:538.561.029.64 1463
Microwave Measurements of the Radiation Temperature of Plasmas—G. Bekefi and S. C. Brown. (*J. Appl. Phys.*, vol. 32, pp. 25–30; January, 1961.) Measurements are described of radiation temperature of positive columns of glow discharges in helium, neon and hydrogen; the microwave noise radiation was detected at 3 Gs. The results are compared with calculations and with Langmuir probe measurements of the electron temperature.

537.56:538.566]+621.391.812.63 1464
Nonlinear Phenomena in a Plasma Located in an Alternating Electromagnetic Field—V. L. Ginzburg and A. V. Gurevich. (*Uspekhi Fiz. Nauk*, vol. 70, pp. 201–246 and 393–428; February and March, 1960.) A detailed treatment is given of elementary and kinetic theories of a nonrelativistic nondegenerate plasma in a uniform electric field, with particular reference to the distribution function of electron collisions, the effective electron temperature and total electric current density. Results are applied to determine the nonlinear effects occurring in the propagation of radio waves in the ionosphere, including self-interaction, cross-modulation and the interaction of unmodulated waves. 133 references.

537.56:538.63 1465
Oscillations and Diffusion in Weakly Ionized Plasmas—J. F. Bonnal, G. Briffod, and C. Manus. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 2859–2861; April 25, 1960.) The diffusion of weakly-ionized plasma particles is observed in a direction perpendicular to the magnetic field.

538.11 1466
Magnetic Moment and Magnetic Charge—J. Fischer. (*Arch. Elektrotech.*, vol. 45, pp. 157–161; April 11, 1960.) The definition of magnetic dipole moment and magnetic charge in relation to the theory of permanent magnets is discussed.

538.114/.115 1467
Some Properties of Concentrated and Dilute Heisenberg Magnets with General Spin—R. J. Elliott. (*J. Phys. Chem. Solids*, vol. 16, pp. 165–168; November, 1960.) The constant coupling approximation is used to obtain the thermodynamic properties of a Heisenberg ferromagnet above the transition temperature.

and the dependence of these properties on magnetic concentration in an alloy is considered. Antiferromagnets are also discussed.

538.114/.115 1468

The Behaviour of Magnetic Systems with Dilution—J. B. Smart. (*J. Phys. Chem. Solids*, vol. 16, pp. 169–173; November, 1960.) A generalization of the Bethe-Peirels-Weiss method to the case of alloys, and application to the dilution of classical Heisenberg ferromagnets with nearest-neighbor interaction only is given.

538.222:538.569.4:538.614 1469

Microwave Modulation of Light in Paramagnetic Crystals—N. Bloembergen, P. S. Pershan, and L. R. Wilcox. (*Phys. Rev.*, vol. 120, pp. 2014–2023; December 15, 1960.) The considerations of Dehmelt and others about the modulation of light by RF signals in atomic vapors are extended to paramagnetic solids. Experimental design criteria are discussed and some possible applications reviewed.

538.24 1470

Anisotropic Magnetization—E. R. Callen and H. B. Callen. (*J. Phys. Chem. Solids*, vol. 16, pp. 310–328; November, 1960.) A discussion of the magnitude and temperature dependence of anisotropic magnetization, with application to different classes of materials, is given.

538.566 1471

Electromagnetic Energy Density in Dispersive Media—F. Borgnis. (*Z. Phys.*, vol. 159, pp. 1–6; May 16, 1960.) General expressions for energy densities and losses are derived.

538.566:535.31:534.512.2 1472

Electromagnetic Reflection from Sound Waves—Schmitt and Wu. (See 1387.)

538.566:535.42 1473

Inhomogeneous Waves in the Diffraction Field near Loss-Free Dielectric Circular Cylinders—V. Müller. (*Z. angew. Phys.*, vol. 12, pp. 206–212; May, 1960.) Measurements were made at 3.2 cm λ of the diffraction field near solid and hollow dielectric cylinders of equal external diameter placed in a parallel-plate transmission line. Results are compared with those obtained for a solid metal cylinder of equal diameter and with the field distribution given by a graphical method. An explanation is given of the origin and structure of the inhomogeneous waves which contribute to the measured diffraction field.

538.566:535.42 1474

Diffraction by an Irregular Screen of Limited Extent—B. H. Briggs. [*Proc. Phys. Soc.* (*London*), vol. 77, pp. 305–317; February 1, 1961.] Information that may be derived about the screen by measurements of the diffraction pattern at different distances is considered. Close to the screen the pattern gives information about the small-scale structure; at greater distances, it is related to the over-all extent of the screen. Measurements of the angular diameter of stars, ionospheric irregularities and meteor trails are discussed.

538.566:535.42 1475

Diffraction by Finite Irregular Objects—R. P. Mercier. [*Proc. Phys. Soc.* (*London*), vol. 77, pp. 318–327; February 1, 1961.] The problem considered in 1474 above is treated by a different method. Its application to lunar radio echoes is discussed.

538.566:537.56 1476

Nonresonance Absorption of Electromagnetic Waves in a Magnetoactive Plasma—B. N. Gershman. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 695–704; September, 1959.) An analysis of the absorption of ordinary, extraordinary and plasma waves outside the gyromagnetic resonance regions, taking account of collisions and the specific absorption mechanism, is given.

538.566:537.56 1477

Growth of Electromagnetic Waves in Interpenetrating Infinite Moving Media—G. G. Getmantsev. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 843–846; September, 1959.) A phenomenological treatment of the propagation of plane monochromatic waves in mutually interpenetrating media is given. Equations are derived for the refractive index and are applied to the problem of stability. Build-up and damping factors are determined for the case of a plasma moving through a dispersionless dielectric.

538.566.2:539.23 1478

Influence of Frequency and Structure on the Complex Refractive Index of Metals and Thin Metallic Films—M. Gourceaux. (*Compte. rend. Acad. Sci., Paris*, vol. 250, pp. 2176–2178; March 21, 1960.) The extension of earlier work (3651 of 1959) is discussed.

538.569.2 1479

Quantum Theory of Spatial Dispersion of Electric and Magnetic Susceptibilities—O. V. Konstantinov and V. I. Perel'. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 786–792; September, 1959.) Electromagnetic effects in a homogeneous medium can be described in terms of a conductivity dependent on frequency and wave vector, and a susceptibility dependent only on the wave vector. A universal relation between conductivity and susceptibility is derived.

538.569.4 1480

The Influence of Coherent Magnetic-Dipole Radiation Field on Magnetic Resonance—G. V. Skrotskiĭ and A. A. Kokin. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 802–804; September, 1959.) Corrections to the relaxation time and resonance-frequency shift due to the effect of the radiation field are computed.

538.569.4:535.33:621.375.9 1481

Beam-Type Masers for Radio-Frequency Spectroscopy—K. Shimoda, H. Takuma, and T. Shimizu. (*J. Phys. Soc. Japan*, vol. 15, pp. 2036–2041; November, 1960.) The behavior of the beam-type maser is considered theoretically, with particular regard to sensitivity and line width. Experimental results for a formaldehyde transition are given for comparison.

538.569.4:538.22 1482

General Spin-Wave Dispersion Relations—R. F. Soohoo. (*Phys. Rev.*, vol. 120, pp. 1978–1982; December 15, 1960.) The derivation takes into account the effects of conductivity, exchange and propagation.

538.569.4:538.221:621.318.134 1483

Calculation of the Widths of Ferrimagnetic Absorption Lines and of Relaxation Times in the Case of Spin Waves of Non-negligible Amplitude—P. E. Seiden. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 2530–2532; April 4, 1960.)

538.569.4:538.222 1484

Pulsed-Field Measurements of Large Zero-Field Splittings: V^{3+} in Al_2O_3—S. Foner and W. Low. (*Phys. Rev.*, vol. 120, pp. 1585–1588; December 1, 1960.)

538.569.4:539.2:621.373.421.1 1485

The Hamiltonian Formalism of Damping in a Tuned Circuit—Stevens. (See 1439.)

538.569.4:621.375.9:535.61–1/2 1486

Proposed Fibre Cavities for Optical Masers—E. Snitzer. (*J. Appl. Phys.*, vol. 32, pp. 36–39; January, 1961.) The practical possibilities of employing fibers as dielectric waveguides at optical infrared wavelengths are discussed.

539.2:537.112 1487

Electron Effective Mass in Solids—a Generalization of Bardeen's Formula—M. H. Cohen and F. S. Ham. (*J. Phys. Chem. Solids*, vol. 16, pp. 177–183; November, 1960.) "A formula is derived for the effective mass of an electron in a crystal which replaces the sum over excited states in the usual sum rule by an integral over the surface of the unit cell. The integrand of the surface integral involves the wave function(s) at the symmetry point or band extremum $\boldsymbol{k}_0$ and a second solution of Schroedinger's equation at the same energy but satisfying inhomogeneous boundary conditions on the cell surface."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.12 1488

An All-Electric Universe—C. E. R. Bruce. [*Elec. Rev.* (*London*), vol. 167, pp. 1070–1075; December 23, 1960.] An outline of theory ascribing the evolution of the universe to electrical discharge processes is discussed. An interpretation is given of observed cosmic phenomena. Correction, *ibid.*, vol. 168, p. 20; January 6, 1961.

523.15 1489

The Theory of Force-Free Magnetic Fields—E. Richter. (*Z. Phys.*, vol. 159, pp. 194–211; June 15, 1960.) The general features of force-free magnetic fields are established and several models are discussed in relation to special coordinate systems. See also 784 of 1959 (Woltjer).

523.164.3 1490

A Curious Feature of the Radio Sky—R. H. Brown, R. D. Davies, and C. Hazard. (*Observatory*, vol. 80, pp. 191–198; October, 1960.) An examination is given of existing theories regarding the spur of intense radiation which appears to emerge from the galactic plane at about $l^{II}=30°$ and to run upwards towards the north galactic pole. Radio observations may be interpreted more reasonably in terms of an irregularity in the local spiral structure or a remnant of a supernova.

523.164.3 1491

Observations of Cosmic Radio Noise at 18 Mc/s in Hawaii—W. R. Steiger and J. W. Warwick. (*J. Geophys. Res.*, vol. 66, pp. 57–66; January, 1961.) Cosmic radio noise measurements at Hawaii on a frequency of 18 Mc are described. On numerous occasions considerable flux continued to arrive when the *F*-region critical frequency exceeded 18 Mc. This effect is possibly due to cosmic radio radiation being trapped and propagated under the ionosphere. Attenuation measurements suggest that an ionospheric window effect exists.

523.164.3 1492

Secular Variation of the Flux Density of the Radio Source Cassiopeia A—J. A. Högbom and J. R. Shakeshaft. (*Nature*, vol. 189, pp. 561–562; February 18, 1961.) Comparative measurements at 81.5 Mc on the RF sources in Cassiopeia A and Cygnus A indicate that the flux density of Cassiopeia A is decreasing at a rate of (1.06 ± 0.14) per cent per annum.

523.164.3:551.510.535 1493

Statistical Analysis of Radio Star Scintillation—S. Gruber. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 59–71; February, 1961.) "Radio star scintillation data are analysed by statistical techniques to find the autocorrelation functions and power density spectra of the fluctuations in amplitude and phase. Interpretation is given in terms of a drifting phase-changing screen as postulated by Booker, Ratcliffe & Shinn [428 of 1951]. It is shown how the drift velocity of the irregularities in ionospheric electron density that lead to the scintillations, as well as the mean square amplitude and phase of the fluctuations, may be obtained from these records."

523.164.32 1494
Preliminary Report on Positional Fluctuations of Solar Radio Sources on 200 Mc/s—P. Maltby. (*Astrophys. Norveg.*, vol. 6, pp. 75–84; December, 1958.) An analysis indicates that the magnitude of the positional fluctuations is less when the solar RF source is closely connected with a magnetic field in the active region. No increase in the magnitude of positional fluctuations was found on 200 Mc as the source moved towards the solar limb.

523.164.32 1495
On the Features of Storm Bursts in the 200-Mc/s Range—Ø. Elgarøy and Ø. Hauge. (*Astrophys. Norveg.*, vol. 6, pp. 85–92; December, 1958.) The characteristics of "short pips" which occur during noise storms are discussed. For brief reports of this work, see 1140 and 1141 of 1958 (Elgaroy).

523.164.32:523.75 1496
10.7-cm Solar Noise Burst of November 20, 1960—A. E. Covington and G. A. Harvey. (*Phys. Rev. Lett.*, vol. 6, pp. 51–52; January 15, 1961.) A burst of RF noise of high intensity is associated with a flare from a sunspot area which had just disappeared behind the sun's west limb.

523.165 1497
Mirror and Azimuthal Drift Frequencies for Geomagnetically Trapped Particles—D. A. Hamlin, R. Karplus, R. C. Vik, and K. M. Watson. (*J. Geophys. Res.*, vol. 66, pp. 1–4; January, 1961.) "For charged particles trapped in the geomagnetic field, the frequencies of the mirror oscillations ω_m and the azimuthal drift ω_d are defined as appropriate averages over the helical motion around the field lines and the mirror motion between reflection points in the two magnetic hemispheres. These integrals for ω_m and ω_d are evaluated numerically. Results are tabulated, illustrated, and represented by approximate analytical expressions."

523.165:551.507.362 1498
Investigation of Cosmic Rays and Terrestrial Corpuscular Radiation by means of Rocket and Satellite Flights—S. N. Vernov and A. E. Chudakov. (*Uspekhi Fiz. Nauk*, vol. 70, pp. 585–619; April, 1960.) A detailed report of observations showing that the earth is surrounded by two separate zones of high-intensity radiation is given. The outer zone is composed of electrons and in the equatorial plane extends from 20,000 to 60,000 km from the center of the earth. It comprises two main energy groups of the order of 20 kev and 10^6 ev. The inner zone extends from 600 km to about 6000 km above the earth and is composed of protons with energies of 10^8 ev. See also 1594 and 3086 of 1960 (Vernov, *et al.*).

523.75:523.165 1499
The Cosmic-Ray Flare on November 12, 1960 and Solar Activity during the Period November 10–15, 1960—(*Nature*, vol. 189, pp. 438–440; February 11, 1961.) A short report of optical, radio-astronomical, cosmic-ray and earth-current observations made at Nera (Netherlands), Paramaribo (Surinam), and Hollandia (New Guinea), is given.

550.38:551.594.5 1500
Interplanetary Magnetic Field and the Auroral Zones—J. W. Dungey. (*Phys. Rev. Lett.*, vol. 6, pp. 47–48; January 15, 1961.) The S_D current system is explained qualitatively using Hoyle's suggestion that auroral particles are accelerated at the neutral points between the geomagnetic and interplanetary fields.

550.385 1501
Pulsation of the Earth's Electromagnetic Field with Periods of 1 to 15 Seconds and their Connection with Phenomena in the High Atmosphere—V. A. Troitskaya. (*J. Geophys. Res.*, vol. 66, pp. 5–18; January, 1961.) Pulsations are closely correlated with aurora and the various phases of magnetic storms. The time of onset of a storm's sudden commencement which is determined from pulsation data appears to be the same, within a few seconds, over the whole world. See 133 of 1960 (Troitskaya and Mel'nikova).

550.385.3 1502
World-Wide Characteristics of Geomagnetic Micropulsations—J. A. Jacobs and K. Sinno. (*Geophys. J. R. Astr. Soc.*, vol. 3, pp. 333–353; September, 1960.) The characteristics of *Pt*'s associated with negative or positive bays in the auroral zone are examined and a distinction is drawn between short-period *Pc*'s (15–30 sec) and those of longer periods (30 90 sec). The long-period, continuous pulsations which appear simultaneously with *Pc*'s in polar regions are also discussed. Equivalent overhead current systems are derived for particular cases of each type.

550.385.4 1503
A Theory of Polar Geomagnetic Storms—J. H. Piddington. (*Geophys. J. R. Astr. Soc.*, vol. 3, pp. 314–332; September, 1960.) This is the third of a series of papers which outline a hydromagnetic theory of storms (see 1209 and 1993 of 1960). The three phases of the D_s variations are accounted for by Hall currents set up around space-charge accumulations in the lower ionosphere. These accumulations result from the deformation of lines of force due to interaction with the solar wind near 06 and 08 hours local time.

551.507.362.2:621.391.812.63 1504
The Application of Ray Tracing Methods to Radio Signals from Satellites—Capon. (See 1624.)

551.507.362.2:621.396 1505
Project Oscar—D. L. Stoner. (*QST*, vol. 45, pp. 56 59. 146; February, 1961.) General information is given about a proposed "orbital satellite carrying amateur radio."

551.507.362.2:621.398:621.3.087.4 1506
Telemetry Signals from Sputnik III—R. E. Henderson. (*Electronic Tech.*, vol. 38, pp. 76–79; March, 1961.) The equipment transcribes from magnetic tape to 35-mm film, displaying the coded pulse signals in a raster form which clearly shows pattern changes. The telemetry coding system is given and two transits are analyzed.

551.510.535+550.38 1507
Scale Times and Scale Lengths of Variables: with Geomagnetic and Ionospheric Illustrations—S. Chapman. [*Proc. Phys. Soc.* (*London*), vol. 77, pp. 424–432; February 1, 1961.] The conception of the atmospheric scale height is generalized to apply to any scalar or vector function of time and/or position. Examples are given relating to magnetic fields and the ionosphere.

551.510.535 1508
Maximum of Temperature in the Middle Ionosphere—O. Burkard. (*Nature*, vol. 189, p. 474; February 11, 1961.) A method is outlined for determining variations in scale height *H* and electron density with height from $N(h)$ profiles and from calculations of the electron loss. Results obtained from observations at Puerto Rico show that *H* has a maximum value at a height of about 200 km, with considerable daily and yearly variations. See 3705 of 1959.

551.510.535 1509
A New Method for the Calculation of N(*h*) Profiles from Ionospheric $h'(f)$ Curves—H. Hojo. (*Nature*, vol. 189, pp. 562–563; February 18, 1961.) The basic integra' for virtual height has been rearranged so that the integrand is a function of phase refractive index instead of group refractive index. This technique avoids an infinity in the integrand at the reflection condition.

551.510.535 1510
The Effective Recombination Coefficient of an Ionosphere Containing a Mixture of Ions—S. A. Bowhill. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 19–30; February, 1961.) The effective recombination coefficient in the *E* layer can be estimated from the fall in electron density during the night, from the changes during solar eclipses, and from the time lag in the diurnal peak density. These coefficients can be expressed as functions of the coefficients for each of several different ions present and are shown to be different from each other. It is suggested that O_2^+ and NO^+ are the ions responsible during the day, and that NO^+ ions predominate at night.

551.510.535 1511
The Formation of the Sporadic-*E* Layer in the Temperate Zones—J. D. Whitehead. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 49–58; February, 1961.) A hypothesis is given for E_s formation based on the observation that E_s occurrence is correlated with the horizontal intensity of the earth's field [see 3501 of 1960 (Heisler and Whitehead)]. If the horizontal wind varies vertically, the horizontal field will produce vertical motion of electrons and ions. An accumulation of electrons will occur at the height for which the vertical velocity passes through zero. The theory also accounts for the persistence of E_s.

551.510.535 1512
Seasonal Variations of Mean Electron Density at Heights between 400 and 1200 km—L. Klinker, K. H. Schmelovsky, and R. Knuth. (*Naturwiss.*, vol. 47, pp. 197–198; May, 1960.) Electron-density profiles are given for summer, 1956, and winter, 1958/1959, which are based on observations of Faraday fading of 20-Mc signals from satellite 1958 δ_2.

551.510.535:539.16 1513
A Note on the Cause of Sudden Ionization Anomalies in Regions Remote from High-Altitude Nuclear Bursts—C. M. Crain and P. Tamarkin. (*J. Geophys. Res.*, vol. 66, pp. 35–39; January, 1961.) The sudden changes in VLF propagation characteristics associated with the nuclear explosions over Johnston Island are attributed to bomb-induced ionization in the ionospheric *D* layer. It is suggested that this ionization is caused by β particles resulting from the decay of fission neutrons.

551.510.535:550.385.4 1514
Geomagnetic Disturbance Effects in Equatorial E_s—B. N. Bhargava and R. V. Subrahmanyan. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 81–84; February, 1961.) A correlation between the disappearance of equatorial E_s observed at Kodaikanal, India, and magnetic-storm disturbances is reported. A lowered E_s critical frequency occurs with a decrease in horizontal magnetic force.

551.510.535:550.385.4 1515

The Total Electron Content of the Ionosphere during the Magnetic Disturbance of November 12–13, 1960—G. N. Taylor. (*Nature*, vol. 189, pp. 740–741; March 4, 1961.) Measurements of the total electron content (n_t) of the ionosphere have been made with the radio telescope at Jodrell Bank by observing the differential Faraday fading of lunar radio echoes. Results show that the reduction in maximum electron density (N_m) was accompanied by a proportional reduction in n_t.

551.510.535:[551.594.5+550.385.37 1516

Ionospheric Absorption at Times of Auroral and Magnetic Pulsations—W. H. Campbell and H. Leinbach. (*J. Geophys. Res.*, vol. 66, pp. 25–34; January, 1961.) Variations in auroral absorption during March and April, 1960, were closely related to magnetic-field micropulsations and short-period intensity changes at 3914Å. The amplitudes of the magnetic micropulsations were diminished during periods of polar-cap-type absorption.

551.510.535(98):523.164.3 1517

A Note on Polar Blackouts—T. M. Donahue. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 76–79; February, 1961.) The twilight effects in polar-cap blackouts are explained in terms of the detachment of electrons from negative oxygen ions in the region of 50 km and higher by visible light which has been attenuated in the atmosphere.

551.510.535 "1960" 1518

Ionosphere Review: 1960—T. W. Bennington. (*Wireless World*, vol. 67, pp. 135–136; March, 1961.) Declining sunspot activity forced signal frequencies down.

551.510.536:551.594.6 1519

An Estimate of Electron Densities in the Exosphere by means of Nose Whistlers—J. H. Pope. (*J. Geophys. Res.*, vol. 66, pp. 67–75; January, 1961.) By examining successive members of families of nose whistlers for March 19, 1959, it was possible to choose between three models of electron density distribution for the exosphere. The distribution with the best fit was $N=KR^{-3}\exp(3.03/R)$; where N is electron density, R is the distance from the earth's center and K is a constant. This fact suggests that the density at 5 earth radii is of the order of 10 electrons per cm³.

551.594.5 1520

A Study of Auroral Coruscations—W. H. Campbell and M. H. Rees. (*J. Geophys. Res.*, vol. 66, pp. 41–55; January, 1961.) Coruscations represent 5 per cent of the total light in the 3914-Å region and have a period of 6–10 sec. They are closely related to micropulsations and ionospheric absorption of cosmic radio noise. The electron density profile associated with the aurora was found to have a maximum in the E region of 1.1×10^6 electrons per cm³.

551.594.6 1521

Spaced Observations of the Low-Frequency Radiation from the Earth's Upper Atmosphere—G. R. A. Ellis. (*J. Geophys. Res.*, vol. 66, pp. 19–23; January, 1961.) "Observations of 5 kc/s radio noise with a network of four stations extending across southern Australia are described. It is shown that there is often good correlation of the amplitude variations over a distance of 3000 km and that the observed amplitude differences may be explained by assuming that the radiation propagates in the earth-ionosphere waveguide from geographically large sources. Some discrete sources were observed."

551.594.6 1522

Sferics from Intracloud Lightning Strokes—L. R. Tepley. (*J. Geophys. Res.*, vol. 66, pp. 111–123; January, 1961.) The experimental observation that many ELF atmospherics are of negative polarity indicates that intracloud lightning strokes radiate significant energy at very low and extra low frequencies. See also 1624 of 1960.

551.594.6:539.16 1523

The Magnetic Flash of the Nuclear Test of 13th February 1960 at Reggane—J. Delloue. (*Compt. rend. Acad. Sci., Paris*, vol. 250, pp. 2536–2537; April 4, 1960.) An EM signal originating in the explosion at Reggane has been recorded with a field strength of 0.1 vpm at a distance of 2500 km.

LOCATION AND AIDS TO NAVIGATION

621.396.933:621.391.64 1524

Infrared Scanners for Airborne Reconnaissance—C. M. Cade. (*Brit. Commun. Electronics*, vol. 8, pp. 94–102; February, 1961.) The advantages of infrared scanning systems are compared with those of radar scanners. The practical problems of system design are given detailed consideration.

621.396.933.23 1525

Radio Guidance Elements of the B.L.E.U. Automatic Landing System for Aircraft—J. S. Shayler. (*J. Brit. IRE*, vol. 21, pp. 17–33; January, 1961. Discussion.) A detailed description is given of the magnetic leader cable and FM radio altimeter equipment used during the final stages of an automatic landing.

621.396.96:629.13.052 1526

Electrical Measurement of Altitude: Part 2—Physical Fundamentals—H. J. Zetzmann. (*Arch. tech. Messen*, no. 290, pp. 45–48; March, 1960.) Part 1: 578 of February.

621.396.963:621.3.087.4 1527

Development and Possible Applications of Equipment for Bandwidth Compression of Radar Signals by means of Storage Capacitors—K. Jekelius. (*Nachrtech. Z.*, vol. 13, pp. 225–233; May, 1960.) The problem of storage-system design for radar relay links is considered, and the requirements of such a relay system regarding resolving power and transmission bandwidth are assessed with reference to tabulated data of several radar systems. Equipment for radar-picture transmission over long distances using channels of narrow bandwidth is described.

621.396.963:681.142 1528

Preparation of Radar Data for High-Speed Digital Computers—R. R. Fidler, R. B. Angus, Jr., and P. F. Marino. (*Sylvania Tech.*, vol. 12, pp. 97–105; July, 1959.) Common mathematical signal-in-noise problems are studied and a brief description is given of an implementation of the results for the case of a signal in continuos random noise.

621.396.963.3:621.397.331.24 1529

An Experimental Scan Conversion System for the Production of Bright Radar Displays—D. L. Plaistowe. (*Marconi Rev.*, vol. 23, no. 139, pp. 184–203; 4th Quarter, 1960.) An experimental investigation is given describing the reproduction characteristics of a twin-gun tube in which a radar image is scanned by an electron beam to produce a brightened image on a standard television picture tube.

621.396.967.2 1530

The Elbe-Weser Shore-Based Radar System—C. le Compte, O. Hilke, J. M. G. Seppen, and W. J. Verhoeff. (*Tijdschr. ned. Radiogenoot.*, vol. 25, no. 2, pp. 59–103; 1960. In English.) A description is given of the radar system and associated equipment

621.396.969:551.578.7 1531

Radar Scatter by Large Hail—D. Atlas, W. G. Harper, F. H. Ludlam, and W. C. Macklin. (*Quart. J. R. Met. Soc.*, vol. 86, pp. 468–482; October, 1960.) Radars with wavelengths of 3.3 and 4.67 cm have been used to measure the back-scatter cross sections of individual artificial hailstones and their variations as melting occurs. Results confirm the cross sections computed theoretically.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 1532

Photoelectric Effect in Ag-As-S Alloys—H. Kutlu and M. Öğder. (*Bull. Tech. Univ. Istanbul*, vol. 13, no. 1, pp. 1–31; 1960.) A report of photoelectric measurements on different alloys formed by mixing Ag_2S and As_2S_3 compounds in varying proportions is made. Curves of photocurrent as a function of illumination intensity and time are given and discussed.

535.215 1533

Investigations on Ca-Sb Films—C. Kunze. [*Ann. Phys. (Lpz.)*, vol. 6, pp. 89–106; May 21, 1960.] Optical and electrical measurements were carried out on Cs-Sb films of different composition, and the results relating to absorption, conductivity and spectral distribution of the photo-effect are discussed with reference to the results of other authors. 44 references.

535.215:537.533.8 1534

Secondary Electron Emission of Antimony-Caesium and Bismuth-Caesium Films of Different Composition—G. Appelt and O. Hachenberg. [*Ann. Phys. (Lpz.)*, vol. 6, pp. 67–81; May 21, 1960.]

535.215:537.533.8 1535

Secondary Electron Emission of Antimony-Rubidium Films—W. Kaneff. [*Ann. Phys. (Lpz.)*, vol. 6, pp. 82–88; May 21, 1960.]

535.215:546.48'221'241 1536

Photoconductivity of Cu-Activated Cadmium Sulphide-Selenide—S. Asano. (*J. Phys. Soc. Japan*, vol. 15, p. 2103; November, 1960.)

535.215:546.48'241 1537

On the Electrical and Optical Properties of *p*-Type Cadmium Telluride Crystals—S. Yamada. (*J. Phys. Soc. Japan*, vol. 15, pp. 1940–1944; November, 1960.) From measurements of optical transmission and Hall coefficient of *p*-type CdTe single crystals, the energy gap is found to be about 1.43 ev, the ionization energy of acceptors about 0.2 ev, and the Hall mobility of holes about 80 cm²/v-sec at room temperature.

535.215:546.817'221 1538

Electrical Structure of PbS Films—D. P. Snowden and A. M. Portis. (*Phys. Rev.*, vol. 120, pp. 1983–1995; December 15, 1960.) Electrical properties have been investigated experimentally as a function of frequency from zero to the microwave range. Data are compared with calculations based on a proposed model.

535.37 1539

Tin-Activated Calcium Orthosilicate Phosphors—R. W. Mooney. (*J. Electrochem. Soc.*, vol. 107, pp. 100–104; February, 1960.)

535.376 1540

Electroluminescence of Insulated Particles: Part 2—K. Maeda. (*J. Phys. Soc. Japan*, vol. 15, pp. 2051–2053; November, 1960.) The theory given in Part 1 (1215 of 1959) is shown to be consistent with recent experimental data.

537.226:546.431'824-31 1541
Study of Solid-Solution Single Crystals containing $BaTiO_3$—T. Sakudo. (*J. Phys. Soc. Japan*, vol. 15, pp. 2112–2113; November, 1960.) Dielectric and optical properties of various Ba-Sr and Ba-Pb titanates have been measured.

537.226:621.319.4 1542
Silicon Nitride Thin-Film Dielectric—C. R. Barnes and C. R. Geesner. (*J. Electrochem. Soc.*, vol. 107, pp. 98–100; February, 1960.) Thin films of Si_3N_4 have been deposited pyrolytically on hot Mo substrates to form capacitors capable of operating satisfactorily in the region of 600°C.

537.226.2:546.824.31 1543
Dielectric Constant and Dielectric Loss of TiO_2 (Rutile) at Low Frequencies—R. A. Parker and J. H. Wasilik. (*Phys. Rev.*, vol. 120, pp. 1631–1637; December 1, 1960.) A rise in dielectric constant from 170 at 1 Mc to 30,000 at 10 cps is confirmed. This is explained in terms of an electron-deficient barrier layer of variable thickness.

537.227 1544
Ferrielectricity—C. F. Pulvari. (*Phys. Rev.*, vol. 120, pp. 1670–1673; December 1, 1960.) A new group of materials with ferroelectric properties has been found. They are mixed crystals of antiferroelectric compounds, *e.g.*, $Na(V_xNb_{1-x})O_3$ or $(Na_{1-x}Ag_x)NbO_3$. The onset of the ferroelectric state is a function of the applied field. Switching transients of a form not previously reported are shown.

537.227 1545
Electrical Properties of Lead-Barium Niobates and Associated Materials—P. Baxter and N. J. Hellicar. (*J. Am. Ceram. Soc.*, vol. 43, pp. 578–583; November, 1960.)

537.227 1546
Ferroelectric Properties of $BaLi_{2x}Al_{2-2x}F_{4x}O_{4-4x}$—T. G. Dunne and N. R. Stemple. (*Phys. Rev.*, vol. 120, pp. 1949–1950; December 15, 1960.) Experimental data on a new type of room-temperature ferroelectric is given. A preliminary study of switching properties shows a switching time $<5\mu$sec for 300-v pulses applied to a crystal 0.1 mm thick.

537.227:546.431'824–31:621.317.335.3 1547
Measurement of Microwave Dielectric Constants of Ferroelectrics: Part 1—Dielectric Constants of $BaTiO_3$ Single Crystal at 3.3 kMcs—E. Nakamura and J. Furuichi. (*J. Phys. Soc. Japan*, vol. 15, pp. 1955–1960; November, 1960.) S-band measurements of the dielectric constant (including loss) of $BaTiO_3$ single crystals by a resonant-cavity perturbation method are described [see 966 of 1946 (Horner, *et al.*)]. Results are given for temperatures from 20 to 170°C.

537.311.33 1548
Kinetic Theory of Impact Ionization in Semiconductors—L. V. Keldysh. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 713–727; September, 1959.) The energy dependence of the impact ionization probability near the threshold is significantly different for crystals with low and high dielectric constants. The solution of the kinetic equations is considered in the two cases. Expressions are derived for the equilibrium number of carriers in a strong field, the impact ionization coefficient and the critical field strength. Increasing the electric field leads to a decrease in the recombination velocity and, as a result, the equilibrium number of carriers begins to increase long before the appearance of impact ionization.

537.311.33 1549
Determination of Numbers of Injected Holes and Electrons in Semiconductors—F. van der Maesen. (*Philips Res. Repts.*, vol. 15, pp. 107–119; April, 1960.) Measurements of the photoelectric Hall effect and photoconduction may give information on the deviations of the equilibrium numbers of holes and electrons. The ratio of these deviations is used in computing the diffusion-recombination length. The ratio of Hall mobility to drift mobility for electrons and holes is evaluated for Ge and Si.

537.311.33 1550
The *p-n* Junction in a Temperature Gradient—W. Klose. [*Ann. Phys.* (*Lpz.*), vol. 6, pp. 25–30; May 21, 1960.] The problem of the nonisothermal junction is treated theoretically and the energy transport across the junction is calculated. The possibilities of experimental verification are considered.

537.311.33 1551
A Theory of the Effects of Carrier-Carrier Scattering on Mobility in Semiconductors—T. P. McLean and E. G. S. Paige. (*J. Phys. Chem. Solids*, vol. 16, pp. 220–236; November, 1960.) The theory given is valid at low temperatures where carrier-carrier scattering is most important. Electron-hole scattering can produce the largest effects, altering both the mobility and its temperature dependence. In some cases the opposite drift velocities of electrons and holes can produce a drag effect which is sufficient to give minority carriers a negative mobility.

537.311.33 1552
Atomic Radius, Electronegativities and Activation Energies in Mineral Semiconductor Compounds—J. P. Suchet. (*J. Phys. Chem. Solids*, vol. 16, pp. 265–278; November, 1960. In French.) Empirical relations are deduced between atomic and structural data and the homopolar and heteropolar contributions to the energy gap, and are used to calculate the latter quantities for about a hundred semiconducting binary compounds. See also 2782 of 1960.

537.311.33 1553
Uniformity of Electrical Current Flow in Cylindrical Semiconductor Specimens with Cylindrical Metallic End Caps—R. Simon, J. H. Cahn, and J. C. Bell. (*J. Appl. Phys.*, vol. 32, pp. 46–47; January, 1961.) "The distribution of current is computed in a cylindrical semiconductor specimen provided with cylindrical metallic end caps of the same diameter as that of the specimen and electrical lead wire of much smaller diameter. Nonuniformity of the longitudinal current density of 1% or less can be obtained in specimens with electrical resistivities at least 200 times greater than that of the end caps if the lengths of the specimens are at least equal to their diameters and if the length of each end cap is at least 0.3 diam."

537.311.33 1554
Theory of Tunnelling—E. O. Kane. (*J. Appl. Phys.*, vol. 32, pp. 83–91; January, 1961.) "The theory of 'direct' and 'phonon-assisted' tunnelling is reviewed. Theoretical I-V characteristics are calculated using the constan field model. Generalizations to nonconstant field and more complicated band structure models are discussed briefly."

537.311.33 1555
Tunnelling from a Many-Particle Point of View—J. Bardeen. (*Phys. Rev. Lett.*, vol. 6, pp. 57–59; January 15, 1961.) In the expression for the transition probability $4\Pi^2/\hbar|M|^2\rho_f$, for the case of tunneling between two metals separated by a thin oxide layer, the matrix element M can be treated as a constant; it is independent of small changes in energy and is also unchanged when the metals pass from the normal to the superconducting state.

537.311.33 1556
Therma Expansion and Related Bonding Problems of some III-V Compound Semiconductors—L. Bernstein and R. J. Beals. (*J. Appl. Phys.*, vol. 32, pp. 122–123; January, 1961.)

537.311.33:535.215 1557
Photomagnetoelectric Effect and Photo-Conductivity in Semiconductors—V. Andresciani. (*Alta Frequenza*, vol. 29, pp. 154–205; April, 1960.) A general theoretical treatment of the photomagnetoelectric effect is given; formulas are derived for the photomagnetoelectric and photoconductivity voltages and currents under conditions of practical interest. Measurements of these voltages and currents and of carrier lifetime were made as a function of the magnetic field on specimens of Ge and Si, and the effects of surface recombination and light spectrum and intensity are assessed.

537.311.33:546.28 1558
Electron Mobilities and Tunnelling Currents in Silicon—R. A. Logan, J. F. Gilbert, and F. A. Trumbore. (*J. Appl. Phys.*, vol. 32, pp. 131–132; January, 1961.) Hall effect techniques were used to measure carrier density as a function of resistivity in Si samples doped with As, P and Sb in the concentration range used in Esaki diodes. The Hall mobility decreases in the order $\mu(Sb)>\mu(P)>\mu(As)$. These differences are not predicted by present theories and hence, possible explanations are discussed.

537.311.33:546.28 1559
Hole and Electron Mobilities in Doped Silicon from Radiochemical and Conductivity Measurements—K. B. Wolfstirn. (*J. Phys. Chem. Solids*, vol. 16, pp. 279–284; November, 1960.) Calculations are given of conductivity mobility at 300°K for holes and electrons in Ga-, As- and Sb-doped Si, from electrical and radiochemical measurements, using the effective mass approximation.

537.311.33:546.28 1560
Scattering of Conduction Electrons by Lattice Vibrations in Silicon—D. Long. (*Phys. Rev.*, vol. 120, pp. 2024–2032; December 15, 1960.) A theoretical model is applied to experimental data to determine how accurately the magnitude and temperature dependence of the various electrical transport effects in nearly pure *n*-type Si can be described by present theoretical ideas about lattice scattering in this material.

537.311.33:546.28:537.533 1561
Experimental Evidence for the Existence of Two Distinct Field-Emission Characteristics from Silicon Emitters—R. L. Perry. (*J. Appl. Phys.*, vol. 32, pp. 128–130; January, 1961.) For the Type 1 relation, the log $I/(10^4/V)$ relation shows a point of inflexion for dc and the relation is quite different for pulses. The relations for dc and pulses are almost linear for Type 2 data. The methods of production of the two types are discussed briefly.

537.311.33:[546.28+546.289]:538.614 1562
Microwave Faraday Effect in Silicon and Germanium—J. K. Furdyna and S. Broersma. (*Phys. Rev.*, vol. 120, pp. 1995–2003; December 15, 1960.) The Faraday rotation and ellipticity in a system of quasi-free carriers is discussed and applied to microwave measurements on semiconductors. Theoretical expressions are analyzed for various ranges of parameters and compared with experimental data.

537.311.33:546.281'26 1563
Grown *p-n* Junctions in Silicon Carbide—C. A. A. J. Greebe and W. F. Knippenberg. (*Philips Res. Repts.*, vol. 15, pp. 120–123; April, 1960.) The preparation and properties

are described and the forward characteristics are explained in terms of a p-i-n structure. Recombination radiation was found to contain violet light.

537.311.33:546.289 1564
Noise Measurements on Germanium Single Crystals in the Range of Impact Ionization between 5° and 10°K—G. Lautz and M. Pilkuhn. (*Naturwiss.*, vol. 47, p. 198; May, 1960.) Circuit noise and resistance as a function of field strength were measured on p- and n-type Ge. The characteristics obtained for p-type materials are discussed. See also 2790 of 1958 (Finke and Lautz).

537.311.33:546.289 1565
Theory of Optical Radiation from Breakdown Avalanches in Germanium—P. A. Wolff. (*J. Phys. Chem. Solids*, vol. 16, pp. 184–190; November, 1960.) A theory is given of the optical spectrum from avalanches in terms of known properties of the band structure and breakdown process. Good agreement is obtained with experimental data.

537.311.33:546.289 1566
Photon Emission from Avalanche Breakdown in Germanium p-n Junctions—A. G. Chynoweth and H. K. Gummel. (*J. Phys. Chem. Solids*, vol. 16, pp. 191–197; November, 1960.) Results are given of an experimental study of the visible and near infrared emission spectrum from avalanching junctions. The emission mechanisms in different parts of the spectrum are discussed.

537.311.33:546.289 1567
The Drift Mobility of Electrons and Holes in Germanium at Low Temperatures—E. G. S. Paige. (*J. Phys. Chem. Solids*, vol. 16, pp. 207–219; November, 1960.) Measurements of drift mobility down to 20°K in samples of Ge containing impurity concentrations from 7×10^{12} to 4×10^{15} per cm^3 are reported. Below 100°K, the observed minority carrier mobility is less than that expected from phonon and impurity scattering, probably due to electron-hole scattering.

537.311.33:546.289 1568
Recombination of Electrons and Donors in n-Type Germanium—G. Ascarelli and S. C. Brown. (*Phys. Rev.*, vol. 120, pp. 1615–1626; December 1, 1960.) The rate of recombination of excess electrons, produced by breakdown at liquid He temperatures, is measured in As-doped Ge, as a function of temperature. α is found to vary as T^{-2}. Recombination light is detected. The recombination cross sections found vary from 10^{-12} to 10^{-11} cm.^{2}

537.311.33:546.289 1569
Germanium Saturated with Gallium Antimonide—J. O. McCaldin and D. B. Wittry. (*J. Appl. Phys.*, vol. 32, pp. 65–69; January, 1961.) The solubility of Sb is greatly enhanced by the presence of Ga, but the converse is not true.

537.311.33:546.289 1570
Solubility of Oxygen in Germanium—W. Kaiser and C. D. Thurmond. (*J. Appl. Phys.*, vol. 32, pp. 115–118; January, 1961.) The solubility as a function of temperature was determined and the precipitation of second phase GeO_2 observed. The maximum solubility in 2×10^{18} atoms per cm^3 and crystals containing 10^{17} atoms per cm^6 have been prepared.

537.311.33:546.289:538.632 1571
Anisotropic Hall Coefficients in n-Type Germanium—H. Miyazawa and H. Maeda. (*J. Phys. Soc. Japan*, vol. 15, pp. 1924–1939; November, 1960.) Values of μ_H/μ are deduced from Hall-effect measurements at 77°K, 90°K, 195°K and 298°K on oriented single crystals of n-type Ge. The results are discussed with reference to theoretical scattering formulas.

537.311.33:546.623'18 1572
AlP: Preparation, Electrical and Optical Properties—H. G. Grimmeiss, W. Kischio, and A. Rabenau. (*J. Phys. Chem. Solids*, vol. 16, pp. 302–309; November, 1960. In German.)

537.311.33:546.681'19 1573
Band Structure and Electron Transport of GaAs—H. Ehrenreich. (*Phys. Rev.*, vol. 120, pp. 1951–1963; December 15, 1960.) Existing experimental data are reviewed and analyzed.

537.311.33:546.681'19 1574
Distribution Coefficient of Zinc in Gallium Arsenide—A. Shibata. (*J. Phys. Soc. Japan*, vol. 15, p. 2107; November, 1960.)

537.311.33:546.682'86 1575
Quadratic Deviations from Ohm's Law in n-Type InSb—R. J. Sladek. (*Phys. Rev.*, vol. 120, pp. 1589–1599; December 1, 1960.) Measurements of resistivity of samples with various carrier concentrations are described. Electron mobility $\mu=\mu_0(1+\beta F^2)$, where F=electric field, and values of β give information on electron scattering processes.

537.311.33:546.823–31 1576
Some Properties of Titanium Sesquioxides containing Vanadium Ions—T. Kawakubo, T. Yanagi, and S. Nomura. (*J. Phys. Soc. Japan*, vol. 15, p. 2102; November, 1960.)

537.312.62:538.632 1577
R.F. Hall Effect in a Superconductor—G. Dresselhaus and M. S. Dresselhaus. (*Phys. Rev.*, vol. 120, pp. 1971–1977; December 15, 1960.) The RF Hall field is calculated for the case of an applied dc magnetic field and RF electric field, both parallel to the metal surface.

537.312.62:539.23 1578
Effect of Residual Gases on Superconducting Characteristics of Tin Films—H. L. Caswell. (*J. Appl. Phys.*, vol. 32, pp. 105–114; January, 1961.)

537.323:546.571'241 1579
Thermoelectric Properties of Ag_2Te—P. F. Taylor and C. Wood. (*J. Appl. Phys.*, vol. 32, pp. 1–3; January, 1961.) Measurements of Seebeck coefficient, electrical conductivity and thermal conductivity, made at room temperatures, indicate that Ag_2Te is inferior to Bi_2Te_3 for cooling applications in the temperature range considered.

538.221 1580
Magnetization Processes in Ferromagnetics—L. V. Kirenskiĭ, M. K. Savchenko, and I. F. Degtyarev. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 616–619; September, 1959.) Powder-pattern and Kerr-effect observations on Si-Fe crystals containing 3 per cent Si, indicate that a process of domain structure rearrangement plays an important part in magnetization.

538.221:538.541 1581
Eddy-Current and Spin-Relaxation Losses in Thin Metal Tapes at Frequencies up to about 1 Mc/s—R. Boll. (*Z. angew. Phys.*, vol. 12, pp. 212–223; May, 1960.) Loss measurements were carried out on Ni-Fe and permalloy tapes of thickness 2.3–30μ. Results obtained are discussed with reference to those given by other authors for these materials and for ferrites, and are related to the findings of classical eddy-current theory. Anomalies are partly due to spin relaxation effects. A method is given for separating the eddy-current and spin-relaxation contributions to the total losses. 58 references.

538.221:538.652 1582
Magnetostriction and Crystal Anisotropy of Nickel-Chromium and Nickel-Vanadium Alloys—T. Wakiyama and S. Chikazumi. (*J. Phys. Soc. Japan*, vol. 15, pp. 1975–1981; November, 1960.)

538.221:621.318.134 1583
Nonresonance Absorption of Oscillating Magnetic Field Energy by a Ferromagnetic Dielectric—M. I. Kaganov and V. M. Tsukernik. (*Zh. Eksp. Teor. Fiz.*, vol. 37, pp. 823–832; September, 1959.) The imaginary part of the longitudinal magnetic susceptibility is calculated using spin-wave theory.

538.221:621.318.134 1584
Magnetic After-Effect due to Electron Diffusion in Mg-Mn Ferrites—S. Krupička. (*Naturwiss.*, vol. 47, pp. 153–154; April, 1960.) Results of investigations on ferrites of the series $Mg_xMn_{1.15-x}Fe_{1.85}O_4$ are briefly reported.

538.221:621.318.134 1585
Resistance Changes of Ferrites in a Magnetic Field—H. Schröder. (*Naturwiss.*, vol. 47, pp. 175–176; April, 1960.) The magnetoresistance effect was measured on n-type polycrystalline $NiO \cdot Fe_2O_3$; curves of magnetization and magnetostriction as a funcition of field strength are also given.

538.221:621.318.134 1586
Apparent Permeability and Intrinsic Permeability at Centimetre Wavelengths—R. Vautier and A. J. Berteaud. (*Compte. rend. Acad. Sci., Paris*, vol. 250, pp. 2527–2529; April 4, 1960.) Critical comment is given on the results of Spencer, *et al.* (see 203 of 1957). Measurements made on two polycrystalline yttrium garnet specimens show that the intrinsic permeability, as usually calculated, is not solely dependent on the characteristics of the material.

538.221:621.318.134 1587
On the Ferrimagnetic Resonance of Europium Iron Garnet—W. P. Wolf. (*J. Phys. Soc. Japan*, vol. 15, p. 2104; November, 1960.) A semi-quantitative theory to account for the observed g-value of Eu-Fe garnet and its variation with temperature is given.

538.221:621.318.134 1588
Spin-Wave Spectrum of Yttrium Iron Garnet—R. L. Douglass. (*Phys. Rev.*, vol. 120, pp. 1612–1614; December 1, 1960.)

538.221:621.318.134 1589
A Further Explanation of the Shape of the Hysteresis Loop of "Square Loop" Ferrites—J. E. Knowles. [*Proc. Phys. Soc. (London)*, vol. 77, pp. 225–229; February 1, 1961.) Theoretical hysteresis loops are calculated, given the stress distribution of the materials. See 273 of January.

538.221:621.395.625.3 1590
Particle Interaction in Magnetic Recording Tapes—J. G. Woodward and E. Della Torre. (*J. Appl. Phys.*, vol. 32, pp. 126–127; January, 1961.) Some improvements have been made to the methods described in 1723 of 1960, but the general conclusions are unchanged.

538.222 1591
Polarization Effects in Electrical "Conductivity" of Artificial Sapphire at High Temperatures—J. Cohen. (*J. Phys. Chem. Solids*, vol. 16, pp. 285–290; November, 1960.)

538.222:537.311.31 1592
Theory of Resistance Minimum in Dilute Paramagnetic Alloys—K. Tani. (*J. Phys. Soc. Japan*, vol. 15, pp. 1960–1962; November,

1960.) An investigation is made of the possibility that the resistance minimum arises from the short-range order of the spins of the solute atoms. See also 257 of 1960 (Brailsford and Overhauser).

538.222:538.569.4 1593

Electron Paramagnetic Resonance of Manganese in TiO_2—H. G. Andresen. (*Phys. Rev.*, vol. 120, pp. 1606–1611; December 1, 1960.)

538.222:538.569.4 1594

Spin-Lattice Relaxation Times in Sapphire and Chromium-Doped Rutile at 34.6 Gc/s—J. H. Pace, D. F. Sampson, and J. S. Thorp. [*Proc. Phys. Soc.* (*London*), vol. 77, pp. 257–260; February 1, 1961.) Measurements were made in the range 1.4–56°K, and relaxation times up to 5 msec were found at the lowest temperatures. Both materials are promising for use in masers at mm λ.

539.23:537.533.7 1595

The Energy Distribution of Moderately Fast Electrons during Penetration through Thin Films—G. Syrbe. (*Z. Phys.*, vol. 159, pp. 237–242; June 15, 1960.) A partial differential equation for the energy distribution is derived and the relation between electron energy and penetration range is calculated. Results of measurements on Al and Al_2O_3 films [see *e.g.*, 206 of 1957 (Young)] are compared with those calculated.

MATHEMATICS

517.431:513.813 1596

The Range of Validity of the Equation curl curl v=grad div $v-\Delta v$—F. Müller. (*Hochfrequenz. und Elektroak.*, vol. 69, pp. 62–67; April, 1960.) Proof is given of the general applicability of the equation on the basis of Riemann geometry.

517.433:621.391 1597

Development Trends of Modern Operational Calculus—F. H. Lange. (*Hochfrequenz. und Elektroak.*, vol. 69, pp. 67–75; April, 1960.) The application in information theory of four analytical methods based on integrals of the Laplace, Carson-Laplace, Duhamel and folding type is considered. The methods are compared and their relative value is assessed.

517.751.2 1598

Prolate Spheroidal Wave Functions, Fourier Analysis and Uncertainty: Part 1—D. Slepian and H. O. Pollak. (*Bell Sys. Tech. J.*, vol. 40, pp. 43–63; January, 1961.) "A complete set of bandlimited functions is described which possesses the curious property of being orthogonal over a given finite interval as well as over $(-\infty, \infty)$. Properties of the functions are derived and several applications to the representation of signals are made."

517.751.2 1599

Prolate Spheroidal Wave Functions, Fourier Analysis and Uncertainty: Part 2—H. J. Landau and H. O. Pollak. (*Bell Sys. Tech. J.*, vol. 40, pp. 65–84; January, 1961.) Application of the theory developed is 1598 above to time-limited and band-limited signals is made. In particular, if a finite-energy signal is given, the possible proportions of its energy in a finite time interval and a finite frequency band are found, as well as the signals which do the best job of simultaneous time and frequency concentration.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) 1600

Changes in WWV/WWVH Standard Broadcasts—(*Tech. News Bull. NBS*, vol. 45, pp. 11–13; January, 1961.) Details are given of a 36-bit, 100-pps time code which has been permanently incorporated in the transmissions from stations WWV and WWVH. See 3978 of 1960.

621.3.018.41(083.74):621.396 1601

A Very-Low-Frequency (V.L.F.) Synchronizing System—C. H. Looney, Jr. (PROC. IRE, vol. 49, pp. 448–452; February, 1961.) A receiving system is described by which standard-frequency VLF transmissions are used to control the phase of the local standard. Frequency synchronization to 1 part in 10^{10}, or better, is possible, and when this is achieved a timing accuracy within 10 μsec is obtained.

621.317.2:538.566.08]+534.844.1 1602

A New Reverberation Chamber for Sound and Electromagnetic Waves—Meyer, Kurtze, Kuttruff, and Tamm. (See 1393.)

621.317.2:621.317.321 1603

Microvolt Calibration Console—(*Instrum. Practice*, vol. 15, pp. 197–198; February, 1961.) The console described has been developed by NBS at Boulder, Colorado. It covers the range $1-10^5$ μv at ten frequencies between 30 kc and 400 Mc. Accuracy is within 2–5 per cent according to frequency. See also 1302 of 1959.

621.317.311:621.383 1604

The Measurement of Short-Circuit Currents particularly with Photocells—W. Gründler. (*Arch. tech. Messen*, no. 290, pp. 57–60; March, 1960.) Potentiometer circuits particularly for use in light measurements are described.

621.317.332.1.029.4 1605

Measurement of Impedance at Audio Frequency—N. P. Scholes and J. E. Macfarlane. (*Electronic Tech.*, vol. 38, pp. 106–107; March, 1961.) A simple apparatus is described for measuring modulus and phase angle using standard components only.

621.317.361 1606

Determination of Frequency Fluctuations with an Uncertainty of 10^{-9} to 10^{-13} by the Digital Measurement of the Period of One Beat—R. Mitterer. (*Frequenz*, vol. 14, pp. 157–162; May, 1960.) Methods for measuring short-term frequency fluctuations of crystal oscillators are compared. A method of using an accurate digital timing device for measuring the slow beat between two frequencies is described and an example is given of the comparison of the frequencies of two 1-Mc oscillators, to an accuracy within 1 in 10^{11} for a measurement duration of 1 sec.

621.317.382:538.632 1607

Automatic Electronic Converters for Measurements of R.M.S. Values and Active Power using Hall Generators—G. Rehm. (*Arch. tech. Messen*, no. 290, pp. 61–64; March, 1960.) Examples are given of circuits, for application in telemetry and process control systems, in which Hall generators are used to relate an alternating current to a direct current of equal effective value, or to convert ac active power to obtain proportional direct current.

621.317.44 1608

Field Measurements by the Method of Harmonics: Filter- and Difference-Type Probes—J. Greiner. (*Nachrtech.*, vol. 10, pp. 156–162; April, 1960.) An experimental investigation and comparison of the characteristics of two of the categories of magnetometer tabulated in 1271 of April are given.

621.317.6:621.382.23 1609

Measurement of Varactor Quality—N. Houlding. (*Microwave J.*, vol. 3, pp. 40–45; January, 1960.) A varactor is rated in terms of its reactance per resistance ratio at high microwave frequencies. Methods of measurement and experimental techniques are discussed, in particular slotted-line and reflectometer arrangements for 10 Gc. Suitable forms of the test holder are shown.

621.317.7:621.397.132 1610

Fundamentals of Electronic Measurement Techniques for Colour Television—Neidhardt. (See 1650.)

621.317.74:621.372.852.2 1611

Elimination of Mismatch Error by Means of a Phase-Shifter—M. E. Gertsenshteĭn and L. N. Bryanskiĭ. (*Izmer. Tekh.*, no. 1, pp. 48–51; January, 1960.) A description is given of a method for reducing errors in waveguide measurements of standing-wave ratio without the necessity of matching the generator and the receiving device at each frequency. The method can be applied to systems which cannot be matched. See also *Radiotekh. Elektron.*, vol. 3, pp. 710–721; May, 1958.)

621.317.755.001.4 1612

Measurement of the Nonlinearity of Deflection Factor of Oscillograph Tubes—K. P. Beisse. (*Elektronik*, vol. 9, pp. 129–131; May, 1960.) Photocell equipment for the direct measurement of deviations from linearity of CRT spot deflection is described.

621.317.784:538.639:537.311.33 1613

Multiphase Wattmeters based on the Magnetoresistance Effect of Semiconductors—M. J. O. Strutt and S. F. Sun. (*Bull. schweiz. elektrotech. Ver.*, vol. 50, pp. 452–458; May 9, 1959.) Single-phase and three-phase wattmeter and varmeter circuits using disk-shaped magneto-resistance elements are given. The compensation of temperature effects is investigated for InSb and InAs elements. The Gauss-effect wattmeters are suitable for operation at frequencies up to 10 Mc; it may be possible to raise this limit to about 300 Mc.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.385.833 1614

The Effect of a Periodic Illumination of an Object on the Electron-Optical Image—E. Gütter and H. Mahl. (*Optik, Stuttgart*, vol. 17, pp. 233–243; May, 1960.) The blurring of the image in electron microscopes resulting from ac heating of the cathode was found to be due to thermal fluctuations in the specimen grid. The frequency dependence of this blurring effect was investigated experimentally and a theoretical treatment of the effect is given.

621.398 1615

A New Principle for Contactless Signalling—W. Engel, F. Kuhrt, and H. J. Lippmann. (*Elektrotech. Z., A*, vol. 81, pp. 323–327; April 25, 1960.) Signaling devices incorporating Hall generators and transistor amplifiers are described; they find application in process control and automation system.

621.398 1616

455-Mc/s Telemetry Ground Equipment—F. F. Thomas. (*J. Brit. IRE*, vol. 21, pp. 69–77; January, 1961.) The principles common to current types of receiving and recording equipment are discussed and a description is given of a test set suitable for calibrating telemetry senders.

621.398 1617

Analogue Telemetry Equipment and Systems: Part 2—S. Poole, A. Potton, and C. O. Titley. (*Electronic Engrg.*, vol. 33, pp. 77–83; February, 1961.) Telemetry systems used for research into "flutter" and similar phenomena associated with flight at transonic and supersonic speeds are described. Part 1: 1284 of April (Young).

621.398:621.396.934 1618
Engineering Aspects of Missile Telemetry Equipment—the Airborne Sender for 24-Channel Telemetry—W. M. Rae. (*J. Brit. IRE*, vol. 21, pp. 57–67; January, 1961.) A brief description is given of a 24-channel AM/FM system; design and production problems are discussed in detail.

PROPAGATION OF WAVES

621.391.812.6:537.56 1619
Reflection and Transmission of Electromagnetic Waves at Electron Density Gradients—F. A. Albini and R. G. Jahn. (*J. Appl. Phys.*, vol. 32, pp. 75–82; January, 1961.) Solutions are obtained for the propagation of plane EM waves parallel to a gradient of free electron density. Reflection and transmission coefficients are derived for transition zones between a dielectric and a uniformly ionized gas and they are found to depend strongly on the width of the zone and less strongly on the detailed profile of the transition.

621.391.812.623 1620
A Note on Diffraction Round a Sphere or Cylinder—G. Millington. (*Marconi Rev.*, vol. 23, pp. 170–182; 4th Quarter, 1960.) A simple treatment is given of diffraction of radio waves round a sphere or cylinder, based on the known solution for propagation over a smooth earth.

621.391.812.63+[537.56:538.566 1621
Nonlinear Phenomena in a Plasma Located in an Alternating Electromagnetic Field—Ginzburg and Gurevich. (See 1464.)

621.391.812.63 1622
The Scattering of Radio Waves by an Extended Randomly Refracting Medium—S. A. Bowhill. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 9–18; February, 1961.) The form of the emergent angular power spectrum is derived for the case of an EM wave incident upon a continuous medium containing three-dimensional inhomogeneities of refractive index. The medium cannot be analyzed as a series of thin phase screens with independent phase profiles, owing to diffractive changes in the wave in passing from one inhomogeneity to the next.

621.391.812.63 1623
Analytical Considerations of Ionospheric Windows for Low-Frequency Radio Waves—J. Heading. (*J. Atmos. Terr. Phys.*, vol. 20, pp. 31–39; February, 1961.) Reflection coefficients for ionospheres with free space below and above are examined for zeros. The vertical propagation case is considered for isotropic ionospheres and for anisotropic ionospheres with vertical magnetic field. An Epstein distribution of electrons is shown to give windows for one particular polarized wave. A numerical example is given.

621.391.812.63:551.507.362.2 1624
The Application of Ray Tracing Methods to Radio Signals from Satellites—I. N. Capon. (*Proc. Phys. Soc.* (London), vol. 77, pp. 337–345; February 1, 1961.) The effect which an ionosphere of given form will have on radio waves from an earth satellite is calculated with very high accuracy. Some results are given which apply to the Doppler shift, Faraday effect, and refraction of the received signal.

RECEPTION

621.376.23:621.391.822 1625
The Amplitude Distribution and False Alarm Rate of Noise after Post-Detection Filtering—S. Thaler and S. A. Meltzer. (PROC. IRE, vol. 49, pp. 479–485, February, 1961.) A digital computer is used to simulate the passage of white Gaussian noise through a narrow-band filter, and the amplitude distribution is examined for several different detectors followed by a post detection filter.

621.391.812.63.029.62 1626
Reception of B.B.C. Television Sound Transmissions on 41.5 Mc/s at Halley Bay, Antarctica—L. W. Barclay. (*J. Brit. IRE*, vol. 21, pp. 89–92; January, 1961.) Signals were observed on 130 days between April and October, 1958, a large number of interceptions being due to normal *F*-layer propagation. Disturbance of the layer is correlated with absence of signals.

621.391.823:621.396.669 1627
Radio-Interference Suppression of the Ignition Systems of Motor Vehicles—W. Walter. (*Tech. Mitt. PTT*, vol. 38, pp. 153–164; May 1, 1960. In German and French.) Sources of interference and means of suppression are briefly noted and the methods of interference-suppression adopted in Swiss Post Office vehicles are described, with details of measuring equipment and test results.

621.396.621.53:621.391.822 1628
The Effect of Mixing Two Noisy Signals—N. A. Huttly. (*Marconi Rev.*, vol. 23, pp. 153–169; 4th Quarter, 1960.) An expression for the SNR in the beat note from two noisy signals being mixed is obtained by a general analytical method. This is applied to practical examples with suitable assumptions.

621.396.621.54:621.382.3 1629
Parasitic Oscillations in I.F. Stages and Frequency Changers of A.M. Receivers—H. H. van Abbe, P. Bikker, A. Cense, A. N. van Dijkum and J. Rongen. (*Electronic Applic.*, vol. 20, pp. 41–55; June, 1960.) The cause of oscillations in transistor circuits is briefly discussed and experiments to investigate the influence of circuit parameters on instability are described. Conditions which can render the stages completely stable without appreciable loss of gain are indicated.

621.396.666:621.376.23 1630
Modifications to Directivity due to the Limiting of Signals—B. Picinbono. (*Compte rend. Acad. Sci., Paris*, vol. 250, pp. 2179–2181; March 21, 1960.) Changes in directivity of a receiving system using a correlation technique are studied in relation to limiting of the signal. General formulas are established and applied to a particular example.

STATIONS AND COMMUNICATION SYSTEMS

621.391:621.38 1631
The Role of Electronics in Information Processing—P. Neidhardt. (*Nachrtech.*, vol. 10, pp. 191–195; May, 1960.) The relation of basic logic switching circuits to the concepts of information theory is discussed.

621.396.2:621.396.712.3:621.397 1632
Wireless Studio Control Equipment—A. Rettig, J. Balodis, and H. Vorrath. (*Rundfunktech. Mitt.*, vol. 4, pp. 102–112; June, 1960.) Equipment is described which provides a one-way speech channel between control cubicles and program and technical staff in television studios. Details are given of a) an AM system operating at about 40 Mc and b) an FM system using frequencies in the range 20–70 kc.

621.396.43:551.507.362.2 1633
Active Satellites—L. Pollack. (*Wireless World*, vol. 67, pp. 52–56; February, 1961.) A description is given of a radio communication system using earth satellites as relays with particular reference to the Courier project.

621.396.43:629.19 1634
Orbital Scatter—(*Electronic Ind.*, vol. 19, p. 84; October, 1960.) A brief report of a proposed world-wide communication system using the reflecting properties of a continuous belt of metallic fibers formed at a height of several thousands of miles above the earth's surface is discussed.

621.396.65:621.396.43 1635
Beyond-Horizon Radio Link—K. Hoffmann (*Elektrotech. Z.*, *B*, vol. 12, pp. 320–324; June 27, 1960.) The installation described uses standard radio-link equipment modified for operation over a distance of about 200 km at 2 Gc to link West Berlin to the telephone system of the German Federal Republic. Transmitter power and receiver sensitivity have been raised, the antenna diameters increased from 3 to 10 m and a system of space-diversity reception has been adopted.

621.396.7.004.6:519.24 1636
A Statistical Method for Predicting the Lifetime of Communication Installations—H. Störmer. (*Arch. elekt. Übetragung*, vol. 14, pp. 217–224; May, 1960.) Predictions of the expected useful life of an installation can be made if the average life of its components, though not accurately known, has been determined by sampling.

621.396.712 1637
Planning the New Motala Long-Wave Broadcasting Station—E. Magnusson and F. Strandén. (*E.B.U. Rev.*, no. 61A, pp. 107–113; June, 1960.) The new 2×300-kw transmitter, due to come into operation during 1961, has been designed to overcome the fading and interference problems affecting the existing station. A special circular array of antennas will be used on a new and larger site. For German version, see *Rundfunktech. Mitt.*, vol. 4, pp. 94–101; June, 1960.)

621.396.712.029.62 1638
Selecting Frequencies for V.H.F. Sound Broadcasting Stations—S. Lacharnay. (*E.B.U. Rev.*, no. 61A, pp. 114–116; June, 1960.) The problem of frequency selection is considered for the case of three transmitters radiating different programs and feeding a common antenna through a multiplexer. A diagram is given showing unsuitable frequency spacings for the band 87.5–100 Mc.

SUBSIDIARY APPARATUS

621.3.087.4:621.395.625.3:537.533 1639
Experiments on Magnetic Tape Read-Out with an Electron Beam—M. M. Freundlich, S. J. Begun, D. I. Breitzer, J. B. Gehman, and J. K. Lewis. (PROC. IRE, vol. 49, pp. 498–509; February, 1961.) The feasibility of the method is demonstrated experimentally. Theoretically, video signal with a bandwidth of 3 Mc could be reproduced with a signal/electron-shot-noise ratio of 30 db from a recorded area density of 0.85 cycles per mil².

621.3.087.4:621.398:551.507.362.2 1640
Telemetry Signals from Sputnik III—Henderson. (See 1506.)

621.311.62:621.316.72 1641
Precision Variable-Frequency Power Supply—E. A. Gilbert. (*Electronics*, vol. 34, pp. 99–100; January 13, 1961.) A ballast tube in a thermal regulating bridge controls the amplitude of the output and an adjustable filter controls the frequency from 50 cps to 2 kc.

621.311.69 1642
Unconventional Methods of Generating Electrical Power—B. A. Eagle. (*Trans. S. Afr. Inst. Elec. Engrs.*, vol. 51, pt. 11, pp. 233–253; November, 1960. Discussion, pp. 253–256.) A

review of methods for the direct generation of electricity from other forms of energy, including methods based on electrochemical and photoelectric conversion, solid-state devices, and high-velocity ionized gas is given. A comparative table of the principal devices is included. 32 references.

621.311.69:621.383.5 1643
Considerations on the Solar Cell—D. A. Kleinman. (*Bell Sys. Tech. J.*, vol. 40, pp. 85–115; January, 1961.) The absorption curve of the material and solar spectrum effects are contained in a single function readily obtained by numerical integration.

621.316.721.078.3:538.569.4 1644
Stabilization of a Magnetic Field by a Nuclear-Magnetic-Resonance Maser—H. Hahn. (*Compte rend. Acad. Sci., Paris*, vol. 250, pp. 2335–2337, March 28, 1960.) The system briefly described will give an absolute long-term stability within ± 1 part in 10^6.

621.316.722:621.382.3:621.316.93 1645
Overload Protection for Transistor Voltage Regulators—A. G. Lloyd. (*Electronics*, vol. 33, pp. 56–59; December 23, 1960.) Various circuit modifications are described for protection against overload.

TELEVISION AND PHOTOTELEGRAPHY

621.397:621.391.826.2 1646
Two Methods for Determining the Signal and Noise Field Strengths of Television Reception in Regions with Multipath Reception—H. Bödeker. (*Nachrtech. Z.*, vol. 13, pp. 213–218; May, 1960.) The problem of field plotting in the presence of reflections is considered and a formula is derived for calculating the angle of incidence of the reflected ray and the ratio of its field strength to that of the direct ray. A method is given for the subjective assessment of picture quality in the field to evaluate signal strength and SNR.

621.397.12 1647
S.C.F.M.—an Improved System for Slow-Scan Image Transmission—C. Macdonald. (*QST*, vol. 45, pp. 28–32 and 32–35; January and February, 1961.) Test results show that a system using subcarrier frequency modulation is superior to the original system using amplitude modulation (see 3266 of 1960.) Details of modulator and demodulator circuits for the new system are given.

621.397.132 1648
Signal/Noise Ratios in the N.T.S.C. Colour Television System—N. Mayer. (*Rundfunktech. Mitt.*, vol. 4, pp. 130–139; June, 1960.) The sensitivity of the NTSC system to interference by sinusoidal signals at various frequencies is determined on the basis of subjective tests.

621.397.132 1649
A New Survey of the B.B.C. Experimental Colour Transmissions—I. R. Atkins, A. R. Stanley, and S. N. Watson. (*B.B.C. Engrg. Div. Mono.*, no. 32, 31 pp.; October, 1960.)

621.397.132:621.317.7 1650
Fundamentals of Electronic Measurement Techniques for Colour Television—P. Neidhardt. (*Elektron. Rundschau*, vol. 14, pp. 187–192; May, 1960.) Specialized electronic test equipment is described including color-bar and pattern generators, vector-scopes, and apparatus for phase measurement and for the statistical analysis of picture content.

621.397.132:621.372.55 1651
The Correction of Nonlinear Distortion in Colour Television Transmission Systems—E. Baumann. (*Bull. schweiz. elektrotech. Ver.*, vol. 50, pp. 458–466; May 9, 1959.) A mathematical treatment, on the basis of Volterra's line functions, of distortion in nonlinear transmission networks as a function of input signal is given. The theoretical design of correction circuits is considered including that of an equalizer effective at one given frequency. See also 1652 below.

621.397.132:621.372.55 1652
Equipment for the Correction of Phase Difference—T. Celio. (*Bull. schweiz. elektrotech. Ver.*, vol. 50, pp. 466–468; May 9, 1959.) An equalizer circuit is described for use in NTSC color television systems conforming to CCIR standards. The system is based on a step-function approximation of the signal nonlinearity; phase difference is less than 5° within a 500-kc band around the color subcarrier frequency, with amplitude correction to within 3 per cent.

621.397.2 1653
A High-Resolution Television System—L. L. Pourciau, M. Altman, and C. A. Washburn. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 105–108; February, 1960.) The camera-tube used is a 1-inch vidicon. Its over-all resolution is improved by increasing the focus field from the usual 40 G, and by a redesigned deflection system. A resolution better than 1000 lines horizontal and 700 lines vertical is obtained. Frame rate is 30 cps, field rate 60 cps with a 2:1 interlace. Lines per frame can be varied from 675 to 1035, and the system bandwidth is 20 Mc.

621.397.2:621.391.812.624 1654
Troposcatter Communications for Intercontinental TV Transmission—E. Dyke. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 81–88; February, 1960.) Technical advances in the operation of tropospheric scatter systems for television are reviewed. Performance data relating to different methods of improving SNR and examples of calculations for system design are given. 51 references.

621.397.2:621.395.625.3 1655
The Choice of the Carrier Frequency and Band-width of the F.M. Channel and the Circuit Arrangement of Magnetic Recording Equipment for Television Picture Signals—W. Dillenburger. (*Rundfunktech. Mitt.*, vol. 4, pp. 113–129; June, 1960.) Better exploitation of tape and recording head properties may be achieved by transposing the carrier frequency of the FM band to a higher frequency, *e.g.*, 50 Mc before demodulation. Experimental recording modulator and demodulator circuits are described and the results obtained are discussed.

621.397.232.6 1656
Contribution on the basis of System Theory, to the Problem of the Optimum Form of the Nyquist Slope in Television—K. Bernath. (*Tech. Mitt. PTT*, vol. 38, pp. 113–117; April 1, 1960. In German and French.) Theoretical attenuation and group velocity characteristics are determined by Bode's method for idealized receiver IF filters used on six different television standards including the CCIR standard with three different Nyquist slopes. The distortion effects in vestigial-sideband systems are briefly considered.

621.397.331.22 1657
An Improved Image Orthicon—E. D. Hendry and W. E. Turk. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 88–91; February, 1960.) Limitations of the 3-inch image orthicon are mentioned. $4\frac{1}{2}$-inch tubes have been produced with improvements in SNR, center and corner resolutions, and edge effects.

621.397.331.22:621.397.61 1658
The Design of a $4\frac{1}{2}$-Inch Image Orthicon Camera Channel—G. E. Partington. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 92–98; February, 1960.) The improved characteristics of the 41-inch tube and the higher order of optical, mechanical and circuit engineering required for the camera channel are discussed.

621.397.6.001.4 1659
Present and Future Test-Line Signals of the French Television Service—A. Pouyferrié and G. Frachet. (*E.B.U. Rev.*, no. 61A, pp. 102–106; June, 1960.) The characteristic features of the vertical-interval test signals used on the French television network are described. For German version, see *Rundfunktech. Mitt.*, vol. 4, pp. 153–157; August, 1960.) See also 1411 of 1960 (Fröling).

621.397.61 1660
Tetrode Television Transmitters for Bank IV/V—U. Finkbein, J. Holle, and S. Tobies. (*Elektrotech. Z., A*, vol. 81, pp. 332–338; April 25, 1960.) The requirements for transmitters operating in the frequency range 470–790 Mc are summarized. The construction and operation of a television transmitter which incorporates new types of UHF triode are described. The rating of the basic transmitter unit is 2 kw for vision and 0.4 kw for sound; higher powers are obtained by combination with appropriate amplifier units.

621.397.61:621.397.132 1661
Brief Presentation of the Additional Requirements for Monochrome Television Transmitters used for the Transmission of Colour Images—G. Coldewey. (*Nachrtech. Z.*, vol. 13, pp. 175–178; April, 1960.) Details are given of proposed signal waveform specifications for the transmission, by monochrome transmitters, of NTSC-system color television modified to CCIR standards [see 4212 of 1959 (Davidse)].

621.397.621:621.385.832 1662
Review of Modern Settling Techniques for Television Screens with One or More Components—U. Fischer. (*Nachrtech.*, vol. 10, pp. 166–171; April, 1960.) Several sedimentation methods used in the production of television picture tubes are described and the problem of adhesion of luminescent materials is discussed.

TRANSMISSION

621.396.61(083.7) 1663
I.R.E. Standards on Radio Transmitters: Definitions of Terms, 1961—(PROC. IRE, vol. 49, pp. 486–487; February, 1961.) Standard, 61 IRE 15. S1.

621.396.712 1664
Bisamberg High-Power Transmitting Installation—(*Electrotech. u. Maschinenbau.*, vol. 77, pp. 185–248; May 1, 1960.) A number of papers are given dealing with various aspects of planning, construction and operation of the high-power broadcast transmitters near Vienna, and include the following titles:

a) **The Expansion of the Austrian Transmitter Network**—W. Füchal, pp. 187–190.

b) **The Tasks of a Medium-Wave Broadcast Transmitting Installation in the Vicinity of the Austrian Federal Capital**—G. Caspar, pp. 190–195.

c) **The Medium-Wave Transmitting Installation Bisamberg in relation to the Planning of Transmitter Networks**—J. Burgstaller, pp. 195–200.

d) **The Transmitters of the Medium-Wave High-Power Broadcast Installation Bisamberg and their Power Supply**—H. Kikinger, pp. 200–211.

e) **The Electrical Equipment of the Bisamberg Aerial Installation**—R. Kayser, pp. 211–219.

f) **The Power Balance of the Bisamberg**

High-Power Transmitting Installation—H. Kikinger, pp. 219–221.

g) **The Common-Frequency Control System of the Bisamberg High-Power Transmitting Installation**—G. Klement, pp. 221–224.

h) **The Transmitter Test and Monitoring Rack of the Bisamberg High-Power Transmitting Installation**—L. Jelinek, pp. 225–230.

i) **The Equipment for Automatic Transmitter Operation in the Bisamberg Installation**—B. Eiermann, pp. 230–233.

j) **The Building of the Bisamberg High-Power Transmitting Installation**—E. Mühlberg, pp. 234–236.

k) **The Aerial Masts of the New Bisamberg Medium-Wave Transmitting Installation**—E. Melan, pp. 236–244.

l) **The Air-Heating, Ventilation and Air-Conditioning Plant of the Bisamberg High-Power Transmitter**—E. Heyna, pp. 245–248.

TUBES AND THERMIONICS

621.382.23 **1665**

Germanium Tunnel Diodes for the High-Frequency Region—G. Kesel, A. Ottmann, and H. N. Toussaint. (*Nachrtech. Z.*, vol. 13, pp. 191–195; April, 1960.) The operating principle and construction are described; performance characteristics of a prototype are given.

621.382.23 **1666**

Evidence for States (Bands) in the Forbidden Gap of Degenerate GaAs and In P: Secondary Tunnel Currents and Negative Resistances—H. Holonyak, Jr. (*J. Appl. Phys.*, vol. 32, pp. 130–131; January, 1961.) The *I/V* characteristics are shown for two types of tunnel diode. The peculiarities of the characteristics are discussed and some tentative conclusions given.

621.382.23:621.317.6 **1667**

Measurement of Varactor Quality—Houlding. (See 1609.)

621.382.3.012 **1668**

Transistor Frequency Response—J. R. James and D. J. Bradley. (*Electronic Tech.*, vol. 38, pp. 80–82; March, 1961.) A graphical presentation of f_T as a function of the dc working point is described and the importance of f_T as a figure of merit is discussed.

621.382.3.012.8 **1669**

An Application of Matrix Methods to the Determination of the Equivalent Circuits of Earthed-Base and Earthed-Emitter Transistors at Low and High Frequencies—R. Mezencev. (*Compte. rend. Acad. Sci., Paris*, vol. 250, pp. 2338–2340; March 28, 1960.) Impedance matrices are developed and applied to T-type equivalent circuits, taking into account capacitance effects at high frequencies.

621.382.323 **1670**

The Alcatron—A. V. J. Martin. (*Electronic Ind.*, vol. 19, pp. 98–100; October, 1960.) A description of an experimental four-terminal field-effect device having a transconductance of 6 ma/v and output power of 6 w is given. Improved performance is considered possible if a semiconductor material such as GaAs is used.

621.382.333 **1671**

Transient Behaviour and Fundamental Transistor Parameters—C. le Can (*Electronic Applic.*, vol. 20, pp. 56–83; June, 1960.) A detailed analysis of the transient behavior of alloy junction transistors, both when switching on and switching off, is given.

621.382.333 **1672**

Symmetrical Transistors—G. H. Parks. (*J. Brit. IRE*, vol. 21, pp. 79–88; January, 1961.) Significant differences in performance compared with unsymmetrical alloy junction transistors are revealed by theoretical examination. Brief descriptions are given of circuits in which the symmetrical type offers improvements.

621.382.333 **1673**

Alloy-Diffused Transistors—E. Wolfendale. (*Electronic Engrg.*, vol. 33, pp. 88–93; February, 1961.) It is shown why conventional alloy junction transistors cannot be further exploited at VHF, and alloy-diffused transistors and their advantages are described. The performances of the two types are compared and the future of the new type is assessed.

621.382.333:621.317.3 **1674**

Measuring Power Transistor Parameters by Pulse Techniques—D. H. Breslow. (*Electronics*, vol. 34, pp. 120–122; January 6, 1961.) The advantages of pulse methods of measurement are discussed and circuit details of test equipment are given.

621.382.333:621.375.4.018.783 **1675**

Investigations of Distortion in Low-Frequency Junction Transistors—H. Hönicke. (*Nachrtech.*, vol. 10, pp. 163–166, 209–216, and 273–277; April–June, 1960.) Second- and third-order distortion coefficients are calculated for small-signal conditions at low frequencies in the three basic circuit configurations. The dependence of distortion on operating conditions is considered in detail.

621.382.333:621.396.822 **1676**

Variation of L.F. Noise Figure of a Junction Transistor—S. Deb and A. N. Daw. (*J. Brit. IRE*, vol. 21, pp. 49–56; January, 1961.) Noise measurements have been made over a 300-cps bandwidth centered on 1 kc at temperatures in the range -20 to $+45°C$ and collector currents between 0.3 and 1.6 ma. The major part of the noise can be attributed to surface recombination.

621.383.5:538.63 **1677**

Variation of Photovoltaic Response with Magnetic Field for a Germanium *p-n* Junction—W. Dunstan. [*Proc. Phys. Soc.* (*London*), vol. 77, pp. 459–466; February 1, 1961.] Measurements are made of the variation with magnetic field of the open-circuit voltage from a Ge *p-n* junction photocell. The effect depends on field orientation and is proportional to illumination. The results are partially explained.

621.385.032.212.3 **1678**

Rise-Time Measurements in MgO Cold-Cathode Diodes—A. Sussman. (PROC. IRE, vol. 49, pp. 517–518; February, 1961.) Current rise times on application of a step voltage have been measured as a function of the initial current and magnitude of the step itself

621.385.032.266 **1679**

A Hollow-Beam Focusing System—P.M. Lally. (PROC. IRE, vol. 49, pp. 514–515; February, 1961.) The system is based on solid-beam Brillouin focusing, the central core of a solid beam of electrons being replaced by a central cylindrical conductor.

621.385.1.029.6:003.62 **1680**

Graphical Symbols for Microwave Valves—G. W. Epprecht. (*Bull. schweiz. elektrotech. Ver.*, vol. 51, pp. 457–461; May 7, 1960.) An attempt is made to rationalize the system of microwave-tube symbols by creating a number of elementary symbols which are then combined to represent any particular electron-beam device. 69 symbols and symbol combinations are given.

621.385.2 **1681**

Theory of the Space-Charge-Limited Diode—H. Pötzl and K. Richter. (*Arch. elekt. Übertragung*, vol. 14, pp. 225–234; May, 1960.) A simplified theory of one-dimensional electron flow is derived, and, assuming the kinetic temperature of the electrons to be constant, the equations of the space-charge-limited diode are solved for the stationary and nonstationary case. The boundary conditions are determined, and the diode admittance is calculated and compared with experimental data.

621.385.3.029.64 **1682**

Small-Signal Performance and Noise Properties of Microwave Triodes—M. T. Vlaardingerbroek. (*Philips Res. Repts.*, vol. 15, pp. 124–221; April, 1960.) A report is given of theoretical and experimental studies of a triode at 4 Gc. An equivalent circuit is deduced to account for internal feedback, noise behavior and electrons in the active space. Methods of measuring the four-pole coefficients and characteristic noise quantities are described. The results of measurements using these methods are compared with theory. 63 references.

621.385.6:537.533 **1683**

An Experiment of Ion Relaxation Oscillation in Electron Beams—Y. Koike and Y. Kumagai. (PROC. IRE, vol. 49, pp. 525–526; February, 1961.) Results obtained using a Hernquvist-type ion-trapping electron gun producing a strip beam are described.

621.385.63 **1684**

A Unified Theory of Electron-Beam Interaction with Slow-Wave Structures, with Application to Cut-Off Conditions—R. M. Bevensee. (*J. Electronics and Control*, vol. 9, pp. 401–437; December, 1960.) Small-signal transmission-line equations are obtained for a longitudinally-confined beam, including relativistic effects, under the action of cavity solenoidal electric field and space-charge irrotational field.

621.385.63 **1685**

Interaction of Premodulated Electron Streams with Propagating Circuits—J. E. Rowe and J. G. Meeker. (*J. Electronics and Control*, vol. 9, pp. 439–466; December, 1960.) Linear and nonlinear analyses are presented for prebunched beams interacting with a travelling RF wave in both growing-wave and beating-wave devices.

621.385.63.032.269.1 **1686**

Space-Charge Effects in Ultra-Low Noise Electron Guns—J. Berghammer. (*RCA Rev.*, vol. 21, pp. 369–376; September, 1960.) The dc characteristics of a typical electron gun used in low-noise traveling-wave tubes are investigated. Variations in positive beam-forming potentials cause sudden jumps in the cathode current. This space-charge effect, rather than temperature limitation, is the main cause of current saturation.

621.385.63.032.269.1 **1687**

Design Considerations for Grid-Controlled Electron Guns for Pulsed Travelling-Wave Tubes—H. J. Wolkstein. (*RCA Rev.* vol. 21, pp. 389–413; September, 1960.) Methods are suggested for determining the parameters for an additional grid-type aperture, current transmission, and cutoff amplification factor. A discussion of the methods of grid positioning and associated design is given for the convergent-flow gun.

621.385.632 **1688**

Correlation between the Theory and Performance of a *C*-Band Travelling-Wave Tube—T. S. Chen. (*J. Electronics and Control*, vol. 9, pp. 321–348; November, 1960.) Performance was predicted on the basis of small-signal normal-mode theory and was found to be in good agreement with measured characteristics

Curves are given of the CW output power over the range 3800–6400 Mc and for 0.007, 0.07 and 0.1 duty cycles. Details are given of the gun and helix design and the measured voltage SWR over the range.

621.385.632 **1689**

Slow-Wave Structures for Electrostatically Focused High-Power Traveling-Wave Tubes—E. F. Belohoubek. (*RCA Rev.*, vol. 21, pp. 377–388; September, 1960.) Two structures are described; one is of the crossed-bar type in which the focusing electrodes are individually supported by chokes, and the other is a structure of the folded-line type with inserted focusing electrodes.

621.385.632 **1690**

Multiple-Transit Amplification by Travelling-Wave Tube—D. A. James. (*Electronic Tech.*, vol. 38, pp. 108–109; March, 1961.) A signal with 1-Mc bandwidth was passed twice through a traveling-wave tube, Type CV 2188, firstly at 4024 Mc, secondly at 4096 Mc. The over-all gain achieved was 10 db, but 23 db should be possible.

621.385.632:621.372.2 **1691**

An Interaction Circuit for Travelling-Wave Tubes—P. J. Crepeau and I. Itzkan. (PROC. IRE, vol. 49, p. 525; February, 1961.) Characteristics of a pair of coplanar meander lines are given.

621.385.64 **1692**

Energy Build-up in Magnetrons—D. Kobayashi. (*Rev. Elect. Commun. Lab., Japan*, vol. 8, pp. 266–269; May/June 1960.) A previous theoretical analysis (719 of 1960) has been extended to energy build-up and is used to explain some characteristics of ppm noise.

621.385.832.032.269.1 **1693**

Emission Decay below Narrow Grid Apertures—A. Sander. (*Le Vide*, vol. 15, pp. 373–380; September/October, 1960. In French and English.) A sharp increase in current-density occurs at the cathode center as the grid aperture is reduced in size. This increase causes failure of the protective space charge and subsequent chemical poisoning of the cathode surface.

621.387.132.223:621.396.96 **1694**

Gas Clipper Tubes for Radar Service—W. W. Watrous and J. McArtney. (*Electronics*, vol. 33, pp. 80–83; December 16, 1960.) Hydrogen-filled thyratrons prevent the building up of dangerous voltages in the pulse-forming circuits.

Translations of Russian Technical Literature

Listed below is information on Russian technical literature in electronics and allied fields which is available in the U. S. in the English language. Further inquiries should be directed to the sources listed. In addition, general information on translation programs in the U. S. may be obtained from the Office of Science Information Service, National Science Foundation, Washington 25, D. C., and from the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C.

PUBLICATION	FREQUENCY	DESCRIPTION	SPONSOR	ORDER FROM:
Acoustics Journal (Akusticheskii Zhurnal)	Quarterly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Automation and Remote Control (Avtomatika i Telemekhanika)	Monthly	Complete journal	National Science Foundation—MIT	Instrument Society of America 313 Sixth Ave., Pittsburgh 22, Pa.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept of Commerce Washington 25, D. C.
Journal of Abstracts, Electrical Engineering (Reseratívnyy Zhurnal: Electronika)	Monthly	Abstracts of Russian and non-Russian literature		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Journal of Experimental and Theoretical Physics (Zhurnal Eksperimentalnoi i Teoreticheskoi Fiziki)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Journal of Technical Physics (Zhurnal Tekhnicheskoi Fiziki)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Proceedings of the USSR Academy of Sciences: Applied Physics Section (Doklady Akademii Nauk SSSR: Otdel Prikladnoi Fiziki)	Bimonthly	Complete journal		Consultants Bureau, Inc. 227 W. 17 St., New York 22, N. Y.
Radio Engineering (Radiotekhnika)	Monthly	Complete journal	National Science Foundation—AIEE	Royer & Roger, Inc. 41 E. 28 St., New York 16, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Radio Engineering and Electronics (Radiotekhnika i Elektronika)	Monthly	Complete journal	National Science Foundation—AIEE	Royer & Roger, Inc. 41 E. 28 St., New York 16, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Solid-State Physics (Fizika Tverdogo Tela)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Telecommunications (Elektrosviaz')	Monthly	Complete journal	National Science Foundation—AIEE	Royer & Roger, Inc. 41 E. 28 St., New York 16, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Automation Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York 35, N. Y.
Electronics Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York, 35, N. Y.
Physics Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York 35, N. Y.
Express Contents of Soviet Journals Currently being Translated into English	Monthly	Advance tables of contents of translated journals		Consultants Bureau, Inc. 227 W. 17 St., New York 22, N. Y.
Technical Translations	Twice a month	Central directory in the U. S. of translations available from all major sources in the U. S.	OTS and Special Libraries Assoc.	Superintendent of Documents U. S. Gov't Printing Office Washington 25, D. C.

Professional Group Meetings

Antennas and Propagation

Los Angeles—March 9

"Study and Design of Unfurlable Antennas," P. D. Kennedy, Lockheed, Sunnyvale, Calif.

San Francisco—March 8

"Collision Cross Sections and Collision Frequency," Dr. C. Cook, Stanford Research Institute, Menlo Park, Calif.

San Francisco—February 15

"Communication Potentialities of Exospheric Scatter," Dr. V. R. Eshleman, Stanford University, Palo Alto, Calif.

San Francisco—February 8

"Plasma, A Propagating Medium and Source of Radiation," Dr. O. Buneman, Stanford University, Palo Alto, Calif.

Washington, D. C.—March 23

"Television Transmitting Antennas with Special Emphasis on a Novel Traveling Wave Structure," Dr. R. W. Masters, Melpar, Inc., Falls Church, Va.

Antennas and Propagation Microwave Theory and Techniques

Boston—March 9

"Dept. of Defense Ionospheric Research Facilities, ARCEBO, Puerto Rico," P. Blacksmith, Jr., and R. S. Allen, ARCRC, Lincoln, Mass.

"A Line Source Feed for a 1000-foot Spherical Reflector," Dr. A. F. Kay, TRG, Somerville, Mass.

Columbus—March 22

"Anisotropic Properties of Artificial Dielectric Media," Dr. R. E. Collin, Case Institute of Technology, Cleveland, Ohio.

Columbus—January 10

"Plasma Dynamics," G. I. Cohn, Ill. Inst. Tech., Chicago.

Orange Belt—March 22

"Applications of the Solid State to Microwaves," Dr. L. Hogan, Motorola, Inc., Phoenix.

"Applications of the Solid State to Microwaves," J. Cacheris, Motorola, Inc., Phoenix.

Orange Belt—February 28

"Recent Advances in Antenna Arrays," Dr. N. Yaru, Hughes Aircraft, Fullerton.

"Computing Mutual Impedance of Elements in an Array," L. Kurtz, Rantec Corp.

"Symposium with Dr. N. Yaru, Dr. R. Elliott, Dr. Bickmore and L. Kurtz as panel members.

Philadelphia—March 15

"Design of Multi-Channel Rotary Joints," D. Bowman, ITE Circuit Breaker, Philadelphia.

"High Power Testing of Rotary Joints," C. Mann, ITE Circuit Breaker, Philadelphia.

San Diego—March 14

Tour of Navy Electronics Lab., Antenna Measurement Facilities, under the guidance of B. Small.

Syracuse—December 15

"New Developments in Microwave Field Measuring Techniques," E. Damon, Ohio State Antenna Lab., Ohio State University.

Audio

Milwaukee—April 11

"The WHAT, WHY and HOW of Stereo Sound," G. Carrington, Allied Radio Corp., Chicago, Ill.

Milwaukee—March 14

"An Improvement in Simulated 3-Channel Stereo," P. W. Tappan, Warwick Mfg. Corp., Chicago.

Philadelphia—March 22

"FM Receivers—a High Fidelity Component," G. Meyer, Fisher Radio Corp., Long Island City.

San Francisco—March 15

"Two-Way Professional Quality Speaker System Using 6–8 Cubic Foot Enclosures" J. B. Craig and G. L. Augspurger, Jim B. Lansing Corp., Los Angeles.

Automatic Control

Baltimore—February 8

"Advancements in Space Vehicle Guidance and Control System Mechanization Techniques," A. G. Buckingham, Westinghouse Electric Corp., Baltimore.

Baltimore—January 12

"An Engineering Presentation of the Second Method of Lyapunov, with Applications," E. J. Lefferts, The Martin Co., Baltimore.

(Continued on page 86A)

Professional Group Meetings

(*Continued from page 84A*)

Dallas-Fort Worth—March 14

"Ferrites As Circuit Elements for Electrically Controlled Microwave Devices," Dr. Von Aulock, Bell Telephone Labs., Whippany, N. J.

Los Angeles—March 14

"X-15 Guidance and Control," N. Cooper, North American Aviation, El Segundo, Calif.

Bio-Medical Electronics

Cleveland—March 1

"Electrostatic Effects in Protein Interactions," H. Bensusan, Ph.D., Benjamin Rose Hospital, Western Reserve University, Cleveland.

"Electrical Manifestations of Muscular Activity," Dr. N. Speralakis, Western Reserve Medical School, Cleveland.

Houston—March 23

"Project Mercury and Telemetry of Physiological Data," Dr. T. F. McGuire and L. A. Geddes, Baylor Medical College, Houston, Tex.

(*Continued on page 88A*)

Professional Group Meetings

(Continued from page 86A)

Houston—February 14

"Problems of Manned Space Flight," Col. J. P. Stapp, USAF, Aerospace Medical Center, Brooks AFB, Tex.

Los Angeles—February 16

"Solid State Devices and Microelectronic Circuits Applied to Biological Instruments," I. Weiman and H. L. Richter, Jr., Electro-Optical Systems, Pasadena, Calif.

Portland—March 23

"The Artificial Kidney," Dr. P. Selling

Portland—February 23

"Experimental Biological Results Using High Energy X-Rays," Dr. J. Bowling, Linfield Research Institute, McMinnville, Ore.

"Generation of High Energy X-Rays with the Linfield Field-Emission Electron Gun," J. Griffith, Electro-Glass Labs., Beaverton, Ore.

Portland—January 26

"Applications of Gamma Ray Analysis and Radiation Instrumentation," V. R. Roberts, Tracerlab, Inc., and J. Rogers, Tektronix, Inc.

(Continued on page 90A)

"Frankly, we've been first in High Power Signal Generators but last in advertising"

Professional Group Meetings

(Continued from page 88A)

Bio-Medical Electronics Nuclear Science

Los Angeles—March 16

"Isotope Dating Techniques," G. J. Fergusson, UCLA.

Broadcasting

Omaha-Lincoln—March 30–1

"Solid State Short Course for Radio and Television Engineers," Guest Speakers: J. Gardner, Tektronix, Mission, Kans.; T. Pierce, Crossley, St. Paul, Minn.; D. K. Haahr, Collins Radio Co., Cedar Rapids, Iowa.

University of Nebraska staff speakers: N. M. Bashara, R. G. Combs, C. M. Hyde, R. M. Ibata, J. Kohl, D. J. Nelson, E. A. Pearlstein, and W. C. Robison.

Philadelphia—March 9

"New Developments in Video Recording," R. Sirinsky, Ampex Corp.

Circuit Theory

Los Angeles—February 21

"A General Method for Determining the Stability of Non-Linear Networks—

(Continued on page 92A)

TRANSISTOR/DIODE MULTIPLES FROM FAIRCHILD

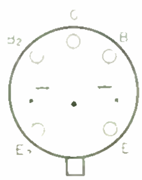
Multiple PLANAR transistor and/or diode dice are packaged in a single TO-5-or TO-18-sized header with common or isolated electrical connections for a variety of circuit functions. PLANAR Multiples offer these advantages: ideal thermal matching, significant space reductions (up to 80%), fewer external soldered connections, reduced component- and circuit-assembly costs, superior device reliability.

PLANAR makes it possible

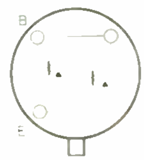
Fairchild's PLANAR process provides the uniformity, parameter stability and high yield to make "multiples" economically and operationally practical.

The Products

PLANAR Multiples can be produced in many device configurations. Some of the production units now available include:



CHOPPER (FSP-1)
h_{FE} 35 at 100 microamps



DARLINGTON AMPLIFIER
(FSP-4)
h_{FE} 2000 minimum @ 150 mA

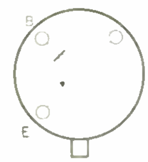
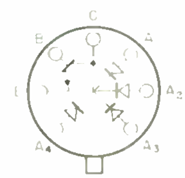
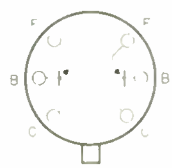


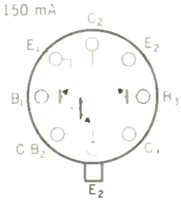
PHOTO-TRANSISTOR (FSP-5)
Extremely low drift



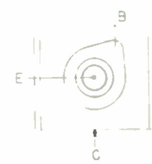
LOW-LEVEL LOGIC (FSP-14)
High speed logic at
low power levels



COMPLEMENTARY AMPLIFIER
(FSP-15
2N1613 and 2N1132



AMPLIFIER MODULE (FSP-17)
High voltage low drift
amplifier



TAB MOUNT TYPE II (FSP-42)
Gold clad moly tab
varnish coated

Special Products Handbook available on request.

MATCHED PERFORMANCE/REDUCED SPACE/IMPROVED RELIABILITY

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A wholly owned subsidiary of Fairchild Camera and Instrument Corporation

FSP-60 Triple Darlington
High Gain Low Noise Amplifier
TO-5 size header
3 - 2N1613 Planar Transistors
h_{FE} typical = 100,000

Professional Group Meetings

(*Continued from page 90A*)

The Second Method of Lyapunov," Dr. L. K. Timothy, Autonetics.

"Analysis of Varactor Diode Circuits," B. J. Leon, Hughes Res. Labs., Malibu, Calif.

San Francisco—March 1

"The Gain Bandwidth Concept in Circuits and in Human Endeavors," K. R. Spangenberg, Consultant, Palo Alto, Calif.

Communications Systems

Los Angeles—February 28

"Using Computers to Solve Problems on System Utilization and System Design," Dr. R. Kalaba, Rand Corp., Santa Monica.

Oklahoma City—March 28

Tour of KOCO television facilities. Various engineers acted as guides and lecturers.

Oklahoma City—January 10

"Weather Facilities," Capt. R. McKissack, USAF, Tinker AFB, Oklahoma City.

Rome-Utica—February 21

"Engineers," J. Bridges, U. S. Dept. of Defense, Washington, D. C.

San Francisco—February 23

"A Synchronous Satellite Relay for Communications," D. Williams, Hughes Aircraft Corp., Los Angeles.

Toronto—April 3

"Command and Dispatch Consoles for Mobile and Point to Point Systems," G. W. Steck, Westrex Corp., New York, N. Y.

Communications Systems Vehicular Communications

Omaha-Lincoln—April 6

"'Bell Boy' Personal Signaling," A. A. Little, Northwest Bell Telephone Co., Omaha.

Component Parts

Los Angeles—March 13

"Development and Applications of Solid State Transducers," H. K. Manning, Fairchild Controls Corp., Los Angeles.

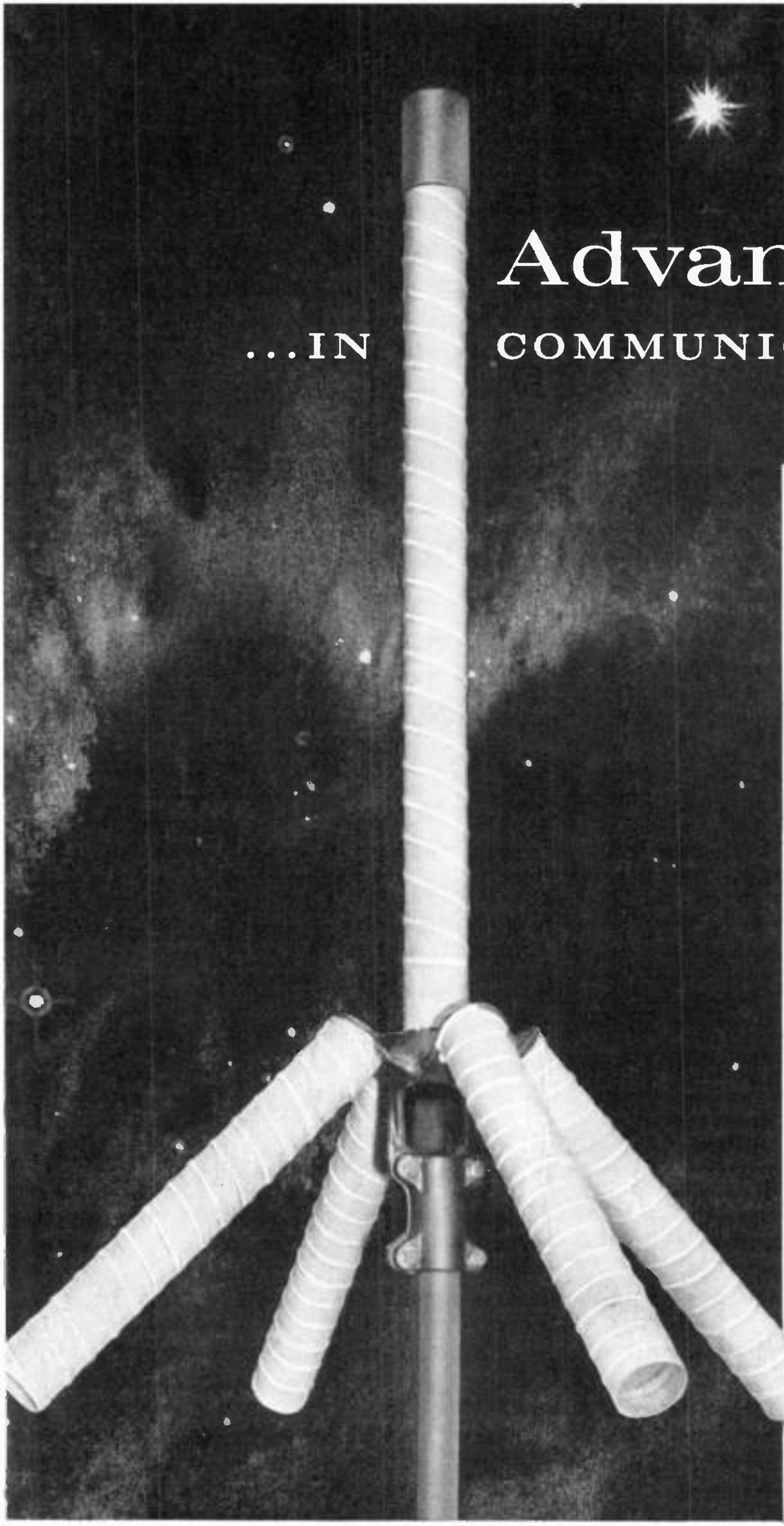
"State of the Art Summary of Solar Cells as Power Convertors for Use in Space," K. A. Ray, Hoffman Electronics Corp., El Monte, Calif.

Electron Devices

Los Angeles—March 14

"Recent Advances and Applications of

(*Continued on page 94A*)



Advanced design

...IN COMMUNICATION ANTENNAS

HELIPOLE

The ANDREW Type 902 HELIPOLE* is the first basically new 30-50 mc fixed station antenna to appear on the 2-way radio scene in the past 12 years. Type 902 employs a new design concept that combines improved performance with mechanical convenience. It is the result of an extensive ANDREW development program.

HELIPOLE CONSTRUCTION

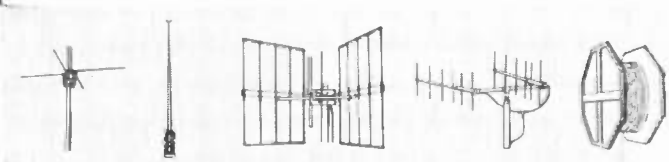
The foreshortened radiator employs a bifilar helical element which is encased in Fiberglass for strength, durability and corrosion resistance. One helix is grounded, providing a static drain path. The other is fed. Ground rods employ single helix conductors which also are embedded in Fiberglass. Size reduction is shown by comparing the 57 inch radiator of Type 902 with 101 inches of a conventional antenna at 30 mc. Ground rods are also shortened by a proportionate amount.

Lightweight and strong – with a maximum total weight of 13 pounds, Type 902 is designed to withstand 30 psf load with ½ inch of radial ice. The focal point of this mechanical strength is found in an aluminum casting to which ground rods and radiator are bolted with stainless steel hardware. Direct mounting is provided for members from 1¾ to 2½ inches in diameter. VSWR of this unity gain antenna is less than 1.5.

Economically priced ANDREW Type 902 is the best performing, corrosion resistant high wind load antenna on the market.

Write or call your ANDREW sales engineer for complete information or request Bulletin 8467.

*TRADEMARK



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Professional Group Meetings

(Continued from page 92A)

High Power TWT," J. T. Mendel, Hughes Aircraft Co., Culver City, Calif.

San Francisco—March 1

"Ion Propulsion," Dr. G. Brewer, Hughes Aircraft Co., Malibu, Calif.

Syracuse—March 7

"Trends in Semiconductor Devices," R. N. Hall, General Electric Co., Schenectady.

Electronic Computers

San Francisco—March 28

"Teaching Machines," Dr. R. S. Hirsch, IBM, San Jose, Calif.

San Francisco—February 28

"Recent Trends in Reliability Theory," Dr. L. C. Hunter, General Telephone and Electronics Lab., Inc.

Engineering Management

Los Angeles—March 29

"Current and Potential Extensions of PERT/PEP Techniques," Dr. D. Green, Operations Research, Inc., Santa Monica.

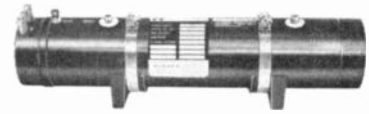
(Continued on page 96A)

1 kw Hughes traveling wave tubes in S-Band

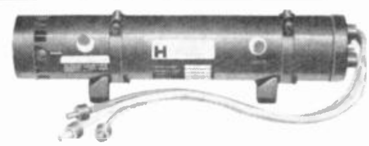
Now available in production quantities, these new and improved tubes offer you 1 kw of pulsed output power, with low power input, minimum heat generation and high reliability.

All these Hughes S-band tubes are lightweight, compact and ruggedly built to withstand the most severe environmental conditions—and provide long life. Each has been fully tested in the field.

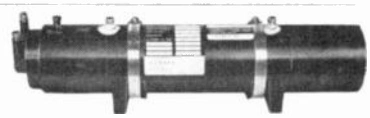
Three of these tubes provide full octave frequency ranges of 2.0 to 4.0 kmc and you have a choice of either ½ or 1% duty, in either ungridded or gridded versions, and with gains up to 37 db. All are permanent magnet periodically focused.



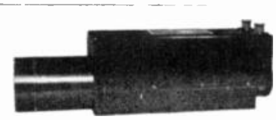
311H Gridded, 1 kw minimum peak power output, 1% duty, 36 db small signal gain @ 50 mw input. Weight: 13 lbs. Length: 17-7/16". Meets usual customer requirements of MIL-E-5400, Class I environmental tests.



312H Gridded 1 kw minimum peak power output, ½% duty, 36 db small signal gain @ 50 mw input. Weight: 11 lbs. Length: 15-3/8". Meets usual customer requirements of MIL-E-5400, Class I environmental tests.



304H Ungridded, 1 kw minimum peak power output, 1% duty, 37 db small signal gain @ 1 mw input. Weight: 12¼ lbs. Length: 17-31/32". Meets usual customer requirements of MIL-E-5400, Class I environmental tests.



313H Ungridded, 1 kw minimum peak power output over the center portion of the band, ½% duty, 36 db small signal gain @ 1 mw input. Weight: 17½ lbs. Length: 16-5/8". Meets usual customer requirements of MIL-E-1 environmental tests.

CREATING A NEW WORLD WITH ELECTRONICS

HUGHES

HUGHES AIRCRAFT COMPANY
MICROWAVE TUBE DIVISION

For information wire or write:
11105 Anza Ave.
Los Angeles 45, Calif.

Professional Group Meetings

(Continued from page 94A)

Engineering Writing and Speech
Military Electronics

Boston—March 14

"The Writing and Evaluation of Technical Proposals," A. P. Hill, Mitre; R. Kendall, Itek, *et al.*

Information Theory

Boston—March 7

"Coding in Practical Communications Systems," Prof. P. Elias, MIT, Cambridge.

Instrumentation

Los Angeles—March 8

"Survey of Storage and Display Equipment," H. Grief, STL.

"The Inconorama Data Display System" Mr. Miller, Fenske, Frederick and Miller.

Trip through Telemetry Data Reduction and Ground Station Facility of STL.

San Francisco—February 28

"Large Scale Data Handling Concepts," R. L. Sink, Consolidated Electrodynamics Corp., Pasadena.

Instrumentation
Space Electronics and Telemetry

Washington, D. C.—March 21

"TIROS II System and Performance," Dr. R. A. Stampfl, Goddard Space Flight Center (NASA), Greenbelt, Md.

Microwave Theory and Techniques

Boston—February 28

"The Limitations of Microwave Duplexers," Dr. L. Gould, Microwave Associates, Burlington, Mass.

Los Angeles—February 9

"Parametric Amplifiers," W. H. Louisell, Bell Telephone Labs., Murray Hill, N. J.

Orlando—March 1

"Some Aspects of Microwave Components Design to Achieve Economical Systems Application," L. H. Fisher, PRD Electronics Co.

Orlando—January 9

Business meeting—Election of Officers.

Washington, D. C.—March 14

"Microwave Components for the 1.5-3 Millimeter Wave Region," L. L. Bertan, FXR Inc., Woodside, N. Y.

(Continued on page 98A)

NOW TIN OXIDE RESISTOR RELIABILITY FOR JUST 6¢

Now you have a happy combination you can play two ways. Use our C resistors in place of composition types to boost product performance at virtually the same cost *or* to maintain the high performance of precision type resistors while cutting costs markedly, saving space.

These C resistors are available in ½ and 1 watt sizes. Both are available in ±5% tolerance. They have the inherent stability of a tin oxide conductor fired onto a glass substrate. We add a special solventproof insulation. Current noise level is less than 0.1 microvolt per volt of applied signal.

The C is the ideal resistor for any of your applications which involve radio or television components, instruments, computers, or other communications equipment.

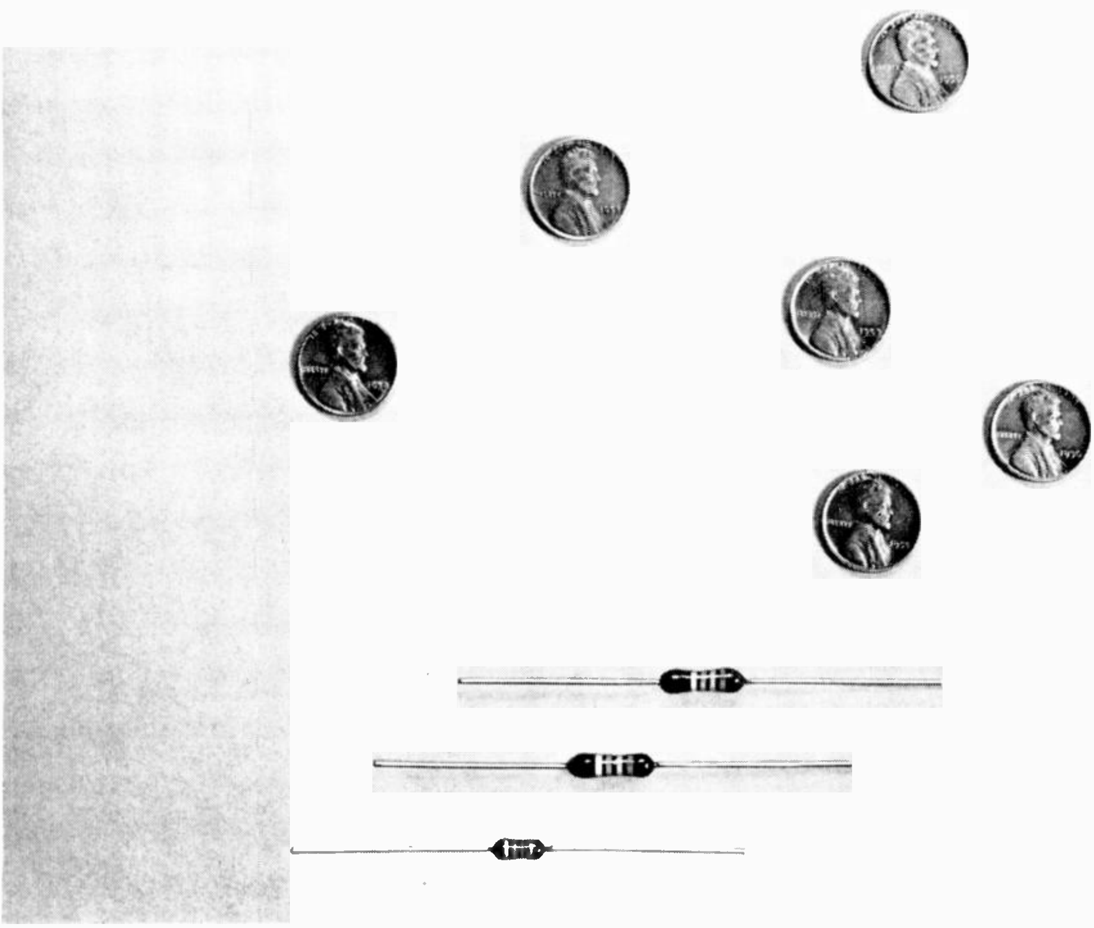
If you're interested in higher wattages, you can get the same basic construction in our low-cost LPI series, which ranges from 3 to 10 watts.

Typical values of Corning C resistors:

Type	Resistance	Wattage	Load Life	Moisture Resistance	Temperature Coefficient	Nominal Dimensions
C-20	51 to 150K	½	.5%	0.3%	150 ppm/°C. (55°C. to +150°C.)	.375"x.138"
C-32	51 to 470K	1	.5%	0.3%	150 ppm/°C. (−55°C. to +150°C.)	.562"x.200"

You can get off-the-shelf delivery from your local Corning distributor.

For complete specs on both C and LPI types, write to Corning Glass Works, 542 High St., Bradford, Pa.



CORNING ELECTRONIC COMPONENTS

CORNING GLASS WORKS, BRADFORD, PA.

Professional Group Meetings

(Continued from page 96A)

MILITARY ELECTRONICS

Boston—March 14

"The Writing and Evaluation of Technical Proposals," A. P. Hill and Panel, Mitre Corp., Bedford, Mass.

Boston—February 23

"The Nature of Command and Control," Maj. Gen. K. P. Berquist, USAF, Bedford, Mass.

Long Island—March 14

"Advanced Radar Concepts," K. E. Forsberg, Sperry Gyroscope Co., Great Neck, N. Y.

Northwest Florida—January 10

"Infrared Theory and Techniques in the Space Age," C. Phillippi, Infrared Section, APGC.

Omaha-Lincoln—March 17

Tour of Atlas Missile Site: Launch Control, GE Radio Guidance, Launcher Missile.

Rochester—February 28

"Satellite Communication Systems," S. Benson, General Dynamics/Electronics.

San Francisco—February 23

"A Synchronous Satellite Relay for Communication," D. Williams, Hughes Aircraft Co., Culver City.

Syracuse—February 23

"High Power Microwave Research," R. Beitz, Cornell Aeronautical Lab., Buffalo, N. Y.

"Activities of National PGMIL," Dr. E. G. Witting, Dept. of Army, Washington, D. C.

NUCLEAR SCIENCE

Los Angeles—March 18

"Isotopes Dating Techniques," G. J. Fergusson, UCLA.

PRODUCT ENGINEERING AND PRODUCTION

Boston—January 10

"Sonic Energy Engineering," T. J. Bulat and C. Schultz, Bendix Corp.

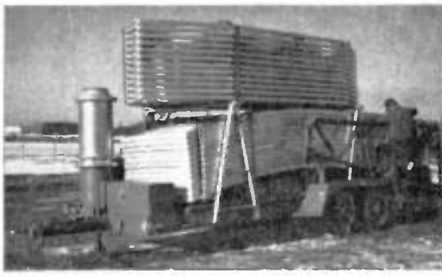
San Francisco—February 28

"Hidden Gold through Work Simplification," T. E. Scatchard, Beckman/Berkeley Div., Richmond, Calif.

San Francisco—January 24

"Product Engineering and Fabrication of Masers," R. Roberts, Micro Engineering Lab., Palo Alto.

(Continued on page 112A)



The antenna completely packed and ready for transit.

The antenna assembled but not erected.

The antenna partially erected.

ASK ASI!

30-foot MOBILE ANTENNA offers the solution to tactical communication problems . . . This is the new Model 111 — a 30′ mobile scatter antenna by Antenna Systems, Inc. designed for tactical applications in scatter links.

In transit, this lightweight, self-contained unit, complete with 30′ mesh reflector, torque tube support, and all associated hardware nests snugly and securely on a steel flat bed trailer. On location, the antenna can be rapidly assembled. A winch or a hydraulic lifting device, mounted on the trailer, provides for quick erection into operating position.

ASI is fully capable of modifying this mobile and versatile package to meet individual requirements . . . to support smaller or larger antennas or pedestals for el-az tracking. For full information, write:

ANTENNA SYSTEMS INC.
HINGHAM, MASSACHUSETTS

FOR ANSWERS TO PROBLEMS IN THE DESIGN, FABRICATION AND INSTALLATION OF ANTENNA SYSTEMS, ASK ASI!

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

BIOPHYSICIST—ENGINEER

Ph.D., MSEE., wishes faculty appointment teaching and research. Publications and author of two books. Can develop biomedical instrumentation program. Equivalent industrial positions considered. Box 3020 W.

TEACHING

Naval officer, aged 38, B.S. in Engineering Electronics plus 35 graduate hours. Retiring in July 1961. 5 years teaching experience, both graduate and undergraduate. Desires teaching position at university in West or Southwest. Textbook author. Resume upon request. Box 3021 W.

R & D MANAGER

Desires assignment in industry or university. 20 years experience in industry, government and universities in R & D teaching, and management encompassing broad fields of physics, electronics, earth sciences, education and administration. Box 3022 W.

PUBLIC RELATIONS EXECUTIVE

Desires opportunity to help put medium or large company and its management on map with small to medium budget; strong with financial community, press, educators; electronics, wire service background. Now with a "top ten" company. Box 3025 W.

TECHNICAL WRITER

Electronics Technician Chief would like to ghost or write under a dual byline with electronics engineers. If you have ideas which you feel fit the popular market but do not have the time to develop and write a practical do-it-yourself article we could probably work to our mutual advantage. Would also like technical writing assignments. Box 3026 W.

INTERNATIONAL OPERATIONS

Former Lieutenant-Colonel Marine Corps with strong civilian background for executive or liaison responsibilities heavy experience South America and Europe. Technically trained to assist manufacturing, sales or field engineering management overseas. Box 3027 W.

REPRESENTATIVE / ENGINEERING

Engineer with electronic and executive experience; age 32; married with no family desires association with firm requiring representation western Canada or anywhere abroad. Position in engineering offering a challenge. Experience in sys-

(Continued on page 102A)

THERE IS SOMETHING NEW UNDER THE SUN

Yes, there *is* something new under the sun. Science is proving this every day. With new discoveries. New explorations. New concepts.

Nowhere is this more evident than in the field of technology. For example: On the drawing boards of Lockheed Scientists and Engineers, new designs are constantly being born—designs in Spacecraft and Aircraft that will reinforce and enlarge our growing knowledge of Outer Space.

These new designs are rapidly developing. And their number is rapidly increasing. The pace is fast. Yet it needs to become faster. To keep pace, Lockheed needs more Scientists, more Engineers. Result? The future for Lockheed was never more promising—the opportunities never greater.

Lockheed feels that trained men will do well to examine thoughtfully the Company's current openings. Notable among these are: Aerodynamics engineers; thermodynamics engineers; dynamics engineers; electronic research engineers; servosystem engineers; electronic systems engineers; physicists (theoretical, infrared, plasma, high energy, solid state, optics); hydrodynamicists; ocean systems scientists; physio-psychological research specialists; electrical-electronic design engineers; stress engineers; and instrumentation engineers.

Scientists and Engineers: To learn more about the opportunities at Lockheed, write Mr. E. W. Des Lauriers, Manager Professional Placement Staff, Dept. 1806, 2402 No. Hollywood Way, Burbank, California. All qualified applicants will receive consideration for employment without regard to race, creed, color, or national origin. U.S. citizenship or existing Department of Defense industrial security clearance required.

LOCKHEED CALIFORNIA DIVISION

Positions Wanted

By Armed Forces Veterans

(Continued from page 102A)

erable management and supervisory experience. Interested in a challenging position which utilizes both engineering and pilot experience. Box 3947 W.

SUPERVISORY OR MANAGERIAL

Supervisory or managerial type position is desired; BS. in mathematics and physics MA. in administration and mathematics; 3 years industrial experience as supervisor of data reduction section; 6 years teaching experience, 4 at the college level and 2 in high school; 4 years experience as Executive Officer USNR; age 34; married with a family. Box 3948 W.

SERVO ENGINEER

BSEE. 1959, E.I.T. 1959; 7 months missile experience; 1½ years R&D design and checkout of fire control equipment. Additional graduate work in engineering and business. 7 months Naval Electronic School; 2 years as Naval Electronics Technician. Desires position in design or sales engineering with opportunity for advancement and further study. Age 27; married, 1 child. Willing to relocate. Box 3949 W.

COMPUTER ENGINEER

BEE., R.P.I. 1956, MSEE., M.I.T. 1958, MS. Physics. N.U. 1961. 5 years experience in computer systems and component development. Looking for challenging position in computer development, applications and/or sales. Desires New York City area. Willing to travel. Box 3951 W.

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
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COMMUNICATIONS ENGINEER

College graduate with several years' experience and good technical background, including radio system application or installation. Consulting engineering firm in New York City. Box 2044.

ASSOCIATE PROFESSOR OF E. E.

Ph.D. required and special competence in electromagnetic theory desirable. Should have some teaching and research experience. Position available Sept., 1961. Write Head of E.E. Dept., State University of Iowa, Iowa City, Iowa.

ELECTRONICS ENGINEER

Inductive Devices. Well established and growing company located in Culver City, Calif., has an excellent opportunity for a capable engineer to organize and manage the design and manufacture of low pass, high pass and band pass filters and electro magnetic delay lines. Salary will be commensurate with training, experience and ability to manage. Liberal company benefits. Box 2047.

(Continued on page 108A)

America's FIRST man into space will rely on a Honeywell designed and developed Attitude Stabilization and Control System for controlling his space capsule. This system automatically damps out initial launch rates, orients and maintains the capsule in proper orbital plane, and provides for the correct descent trajectory and re-entry angle. This device is just one of the many contributions being made by Honeywell scientists and engineers to our nation's space programs.

Honeywell

Military Products Group

WEAPONS SYSTEMS ENGINEERS

Increasing needs for new capabilities in total weapons systems research has created these high-level professional openings in the Aero Division of the Honeywell Military Products Group.

Project Engineer—Minimum 12 years' experience in all aspects of small missile and space weapon systems. Requires thorough understanding of all analytical disciplines, including propulsion, structure, aerodynamics, dynamics, thermodynamics and electronics as well as project management experience including major subcontract management. Missile and space proposal experience required.

Project Development Engineer—Capable of defining total test program for a missile system. Including component testing, sub-system, system, and remote site testing. Know facilities and tests required, monitor all in-house and subcontractor testing. Know military qualification testing. Solid engineering experience necessary. 10-13 years' design and analysis experience.

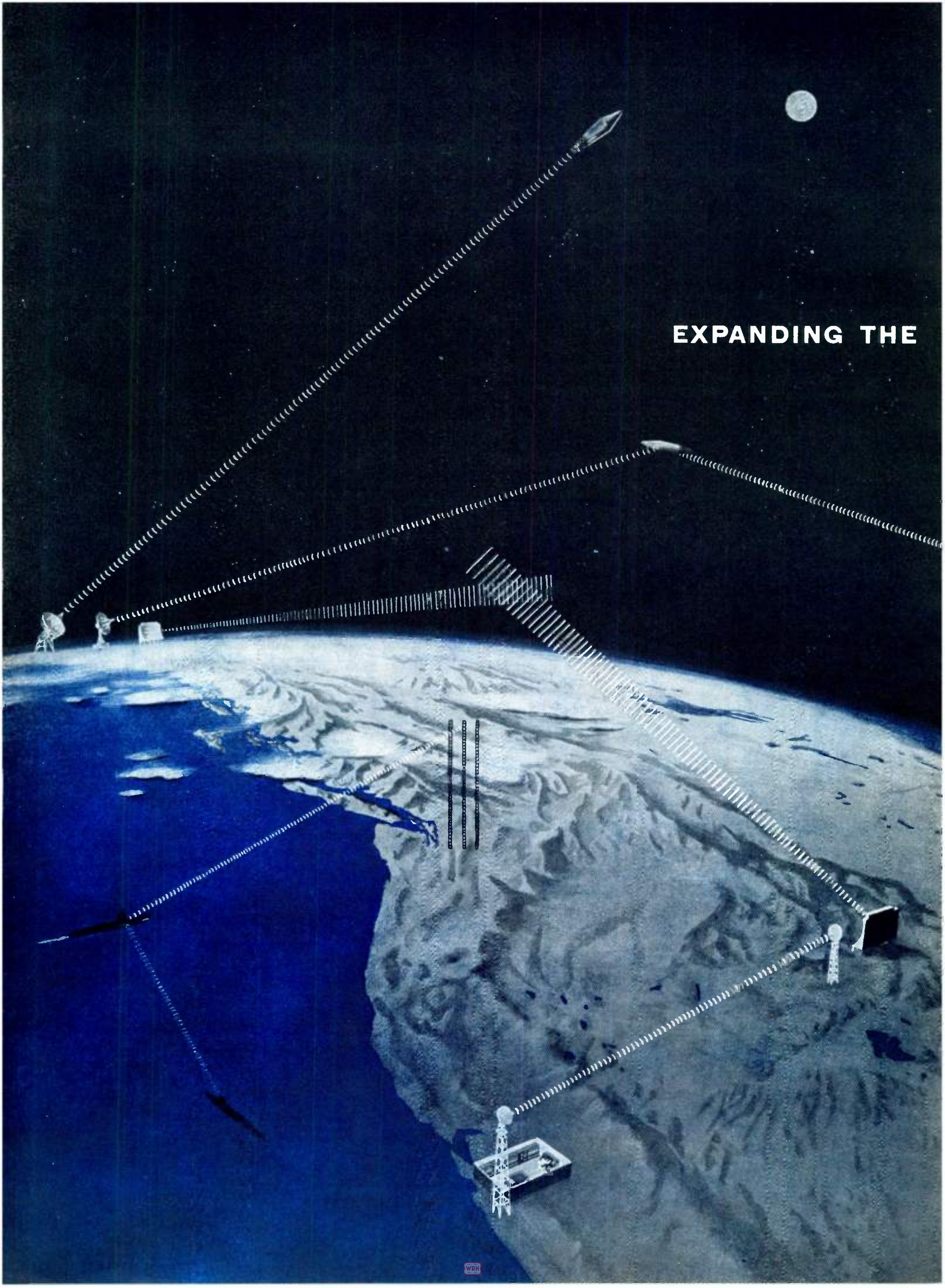
Structures Analysis Specialist and Designer—Minimum 10 years in missile and space-craft structures analysis work. Self-starting and capable of analyzing load paths and stresses in all materials used in space-craft and missiles, defining a structural test program and guiding structural designers.

Aerodynamicist Specialist—Minimum 10 years in missile and space-craft aerodynamics. Analytical and design experience sub-sonic and hyper-sonic aerodynamics; trajectory formulation; computer applications and analysis; aerodynamic configuration design; magnetic fluid dynamics; flight performance analysis; stability and control; wind tunnel design and testing.

Thermodynamics Specialists—Staff Engineer and Engineer—Minimum 10 years in thermodynamics field (missile and space-craft). Analytical and design knowledge of thermal control techniques; emissivity measurements; heat balance; conduction; radiation, convection; heat balance in electronic packages; heat transfer through sub-sonic and hyper-sonic structure; sputtering and meteorite effects; aero-thermo chemistry.

Send your résumé stating your areas of interest, or request for further information to: Mr. Clyde W. Hansen, Technical Director, Aeronautical Division, 2644 Ridgway Road, Minneapolis 40, Minnesota. All qualified applicants considered regardless of race, creed, color, or national origin.

To explore professional opportunities in other Honeywell operations, coast to coast, send your application in confidence to: Mr. H. B. Eckstrom, Honeywell, Minneapolis 8, Minnesota.


EXPANDING THE

Herodotus, the historian, records (490 B.C.) the use of burnished shields for military signaling. This was the forerunner of the heliograph, invented by Sir Henry C. Mance, which came into wide use centuries later.

FRONTIERS OF SPACE TECHNOLOGY IN

COMMUNICATIONS

Lockheed's interest in developing the science of communications extends from the depths of the oceans to deep space. Its Missiles and Space Division research programs deal with the development and application of statistical communication and decision theory in such areas as countermeasures; telemetry multiplexing and modulation; scatter communications; multiple vehicle tracking; millimeter wave generation and utilization; sonic signal detection and processing; avoidance of multipath degradation; and interference avoidance.

Associated research and development efforts are directed toward propagation studies and advanced antenna design; low noise amplifiers; vehicle borne signal transmission and reception, data storage and processing; solid state materials and devices.

The scope of such activities extends from advanced studies of naval communication problems on and under the oceans; the many applications to satellite vehicles; on to the specialized communication problems of deep space explorations. Latter needs are exemplified by high frequencies, low weight and power, high stability, low effective bandwidth, extreme reliability and basic simplicity requirements.

Engineers and Scientists: Investigating the entire spectrum of communications is typical of Lockheed Missiles and Space Division's broad diversification. The Division possesses complete capability in more than 40 areas of science and technology – from concept to operation. Its programs provide a fascinating challenge to creative engineers and scientists. They include: celestial mechanics; communications; computer research and development; electromagnetic wave propagation and radiation; electronics; the flight sciences; human engineering; magnetohydrodynamics; man in space; materials and processes; applied mathematics; oceanography; operations research and analysis; ionic, nuclear and plasma propulsion and exotic fuels; sonics; space medicine; space navigation; and space physics.

If you are experienced in work related to any of the above areas, you are invited to inquire into the interesting programs being conducted and planned at Lockheed. Write: Research and Development Staff, Dept. M-18B, 962 W. El Camino Real, Sunnyvale, California. U. S. citizenship or existing Department of Defense industrial security clearance required. All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

Lockheed / MISSILES AND SPACE DIVISION

Systems Manager for the Navy POLARIS FBM and the Air Force AGENA Satellite in the DISCOVERER and MIDAS Programs

SUNNYVALE, PALO ALTO, VAN NUYS, SANTA CRUZ, SANTA MARIA, CALIFORNIA
CAPE CANAVERAL, FLORIDA • HAWAII

WRH

Positions Open

(Continued from page 104A)

ELECTRICAL ENGINEERING TEACHING POSITIONS

Ph.D. degree required. Teaching experience desirable but not necessary. Excellent opportunity for young man interested in teaching electronics, network theory, control systems and computers at undergraduate and graduate level. Appointment effective Sept. 1961. Write, Chairman, E.E. Dept., University of Houston, Houston 4, Texas.

ASSISTANT & ASSOCIATE PROFESSOR

Applications are invited for Assistant and Associate Professor of E.E. Candidates should be well qualified academically, preferably to the doctorate level, and should have some research, design or teaching experience in control systems. Duties include teaching at undergraduate and graduate levels, organization and direction of laboratory classes, conducting research and supervising research students. Salary scales are open and competitive with those of industrial and research establishments. Additional stipends are offered to professors who remain on the campus for 11 months of the year and carry out research during this period. Write to Chairman, Dept. of E.E., McMaster University, Hamilton, Ontario.

DESIGN ENGINEER

Design Engineer for low noise and wide band UHF and VHF amplifiers. Outstanding opportunity. Chief Engineer potential. Small, vigorous firm, located in beautiful central Pennsylvania. Educational opportunities across the street at Penn State. Stock option. Send resume to Community Engineering Corp., P.O. Box 824, State College, Pa.

TEACHING POSITIONS

The E.E. Dept. of the City College of New York has several positions available on the teaching staff beginning Sept. 1961. Rank and salary commensurate with qualifications and experience. Opportunity for graduate study. Applicants must be present residents of the U.S. Address inquiry to Prof. H. Taub, Dept. of E.E., The City College, Convent Ave. at 139th St., New York 31, N.Y.

SENIOR ELECTRONIC ENGINEERS

Electronic Engineers are needed for development of new types of power supplies and other electronic instruments. Experience is desired in the fields of power supplies, AC line regulators, electronic instruments and magnetic amplifier and transistorized circuits. Salary is open and is commensurate with applicant's background and ability. Company benefits. Apply Perkin Electronics Corp., 345 Kansas St., El Segundo, Calif.

ENGINEERS

Openings for Electrical Engineering Dept. Chairman and for Assistant or Associate Professor. Must have Doctorate. Opportunity for research in Bio-medical Electronics. Excellent salary and environment. Send resume to Chairman, Dept. of E.E., University of Vermont, Burlington, Vermont.

SCIENTIST—TRANSLATORS RUSSIAN TO ENGLISH

Supplement your income while you keep abreast of latest Soviet research in your field. Guaranteed steady flow of translation in your specialty. Must have graduate study and experience in Physics (atomic, plasma, solid state) Crystallography, Instrumentation, Automation. Native command of English necessary. Chinese-English scientist-translators also wanted. Apply Consultants Bureau, 227 West 17th St., New York 11, N.Y.

(Continued on page 110A)

Positions Open

(Continued from page 110A)

This is a staff and faculty appointment with full participation in our graduate school program. Write for information, academic catalog, and application form to: The Director of Personnel, The Institute of Paper Chemistry, Appleton, Wisconsin.

ELECTRONICS INSTRUCTOR

Junior College: to teach general electronics theory and math., pulse circuits, measurements, transistors. State experience in the fields of TV, radar, computers, magnetic amplifiers, microwaves. Send resume and references to College Supervisor, Certified Selection, Personnel Div., Los Angeles City Board of Education, 450 North Grand Ave., Los Angeles 12, Calif.

PART TIME EMPLOYMENT

Part time job for semi-retired salesman experienced in electrical-electronic-mechanical field. Newly established company. Commission basis. Write Box 2051.

ENGINEER—MATHEMATICIAN

We should like to add a Statistical Mathematician to our staff, who would be concerned with problems in the information processing area—more specifically, projects utilizing statistical decision theory, perception recognition and factor analysis. An example of some current research in this area is Pattern Recognition Studies: decision criteria, tests, statistical analysis, and rating factors applicable to pattern recognition problems are being investigated. Mathematical models including the use of the analysis of stochastic processes will be used to develop criteria and system parameters within the various areas of recognition research, information retrieval and artificial intelligence. Send inquiries to Dr. L. S. Sheingold, Director, Applied Research Lab., Sylvania Electronic Systems, 100 First Ave., Waltham 54, Mass.

Professional Group Meetings

(Continued from page 98A)

RADIO FREQUENCY INTERFERENCE

Philadelphia—April 11

"RCI and Radiation Hazard Reduction on BMEWS," E. M. Brown, RCA Victor Div., Moorestown, N. J.

San Francisco—March 14

"Interference in Power Systems," F. Rowe, Northern California Electrical Bureau, San Francisco; R. Lake, Pacific Gas and Electric, San Francisco.

RELIABILITY AND QUALITY CONTROL

Columbus—February 1

"Circuit Synthesis Techniques," L. Stember, Jr., Battelle Memorial Institute.

(Continued on page 114A)

Professional Group Meetings

(Continued from page 112A)

Columbus—January 18

"Field Failure Reporting Systems," R. A. Yereance, Battelle Memorial Institute.

Columbus—December 15

"Acoustical Testing and Measurement," Mr. Moskal.

Columbus—December 9

"Radiation Environment," A. Plumer, Battelle Memorial Institute.

Columbus—November 16

"Theory of Mechanical Testing," R. J. McCrory, Battelle Memorial Institute.

Los Angeles—March 20

"The Role of the Reliability Engineering Department," P. A. Adamson, Hughes Aircraft, Fullerton, Calif.; W. K. Warner, North American Aviation, Missile Div.; and R. A. Orr, Aerojet General Corp., Azusa, Calif.

Los Angeles—December 5

"Reliability Aspects of Secondary Power Supplies for Space Vehicles Using Photovoltaic Energy Converters," Dr. M. Wolf, Hoffman Electronics.

"Microminiature Semiconductor Devices for Space Applications," E. E. Maiden, Pacific Semiconductors, Inc.

"Reliability in Solid State Circuits" J. R. Nall, Fairchild Semiconductors.

"Sources of Unreliability in Space Vehicles," C. King, Space Technology Labs.

"A Reliable Design for an Orbiting Space Craft," I. Doshay, Aerojet Gen. Corp.

"The Space Environment and its Effect on Materials and Components" S. N. Lehr and V. J. Tronolone, Space Technology Labs.

"Design by Worst Case Analysis," W. D. Ashcraft and W. Hochwald, Autonetics Inertial Navigation Division of North American Aviation.

"Production Reliability Programs for Space Vehicles," L. R. Brown, Hughes Aircraft Co.

Philadelphia—March 28

"ARINC Monograph #9, Concepts Associated with System Effectiveness," Dr. E. L. Welker, ARINC, Washington, D. C.

"Prediction of the Distribution of Down Time for the Tandem Connection of Multichannel Units," W. B. Rohn, Bell Telephone Labs., Whippany, N. J.

Space Electronics and Telemetry

San Francisco—February 23

"A Synchronous Satellite Relay for

(Continued on page 116A)

Professional Group Meetings

(Continued from page 114A)

Communications," D. Williams, Hughes Aircraft Co., Culver City.

San Francisco—January 17

"Radar Techniques for Satellite Tracking," E. K. Stodola, Reeves Instrument Corp., Garden City, L. I.

Vehicular Communications

Los Angeles—March 16

"Mechanical Filters for FM Mobile Applications," R. A. Johnson, Collins Radio Co., Western Div., Burbank, Calif.

Membership

The following transfers and admissions were approved and are now effective:

Transfer to Senior Member

Breese, M. E., Huntington, L. I., N. Y.
Brown, E. M., Collingswood, N. J.
Brown, J. L., Jr., University Park, Pa.
Coe, G. J., St. Louis, Mo.
DeClaris, N., Ithaca, N. Y.
Dryden, V. W., Palo Alto, Calif.
Dungan, M. R., Whippany, N. J.
Eidson, H. G., Jr., Winston-Salem, N. C.
Forsman, M. E., Gainesville, Fla.
Hager, C. K., Garland, Tex.
Iwanovsky, A., Arlington, Va.
Laeser, P. B., Milwaukee, Wis.
Lang, W. W., Poughkeepsie, N. Y.
McCollom, K. A., Idaho Falls, Idaho
McCotter, J. D., Jr., Lansdale, Pa.
Melos, C. M., Saratoga, Calif.
Richmond, G. E., Snyder, N. Y.
Robinson, R. B., Bellevue, Wash.
Rudin, M. B., Tustin, Calif.
Shevel, W. L., Jr., Yorktown Heights, N. Y.
Stone, L. M., Torrance, Calif.
Talley, T. J., New York, N. Y.
Van Den Meersche, A. J., Ghent, Belgium
Voorhoeve, E. W., Ambler, Pa.
Walls, A. B., Baltimore, Md.
Wilk, J., Hamilton, Ont., Canada
Willenbrock, F. K., Cambridge, Mass.
Woestman, J. W., Havertown, Pa.
Wolfe, R. E., State College, Pa.

Admission to Senior Member

A'cs, E., Budapest, Hungary
Allen, G. Y. R., Ishington, Ont., Canada
Anderson, P. M., Ames, Iowa
Balber, D., Little Falls, N. J.
Bogle, R. E., Sierra Vista, Ariz.
Chemerys, M., Roslyn, Pa.
Chipman, L. D., Winston-Salem, N. C.
Cohen, A. E., Wanamassa, N. J.
Commons, H. E., Lancaster, Calif.
Cornog, R., Santa Monica, Calif.
Craig, J. H., Cleveland, Ohio
Cuccia, C. L., Los Angeles, Calif.
Dallos, A., Budapest, Hungary
Frederick, G. E., Chevy Chase, Md.
Gallo, M. A., Zurich, Switzerland
Gergely, G. J., Budapest, Hungary
Groshans, H. H., Godfrey, Ill.
Hall, T. C., Culver City, Calif.

(Continued on page 118A)

Membership

(Continued from page 116A)

Halligan C. W., Bedford, Mass.
Hegedus, G. J., San Pedro, Calif.
Holloran, T. P., Gardena, Calif.
Johnson, C. E., Jr., Minneapolis, Minn.
Jolly, S. A., Owensboro, Ky.
Jones, A. H., Anaheim, Calif.
Katz, L., Baltimore, Md.
Klopfenstein, A., La Habra, Calif.
Miyagi, M., Yokohama, Japan
Oatley, C. W., Cambridge, England
O'Neill, R., Berkeley, Mo.
Peairs, M. H., Pacific Palisades, Calif.
Pitrone, J. A., Mayfield Heights, Ohio
Reeves, R. J. D., Bracknell, Berkshire, England
Sarwate, M. B., Geneva, Switzerland
Selby, P., Dallas, Tex.
Shores, M. W., Pomona, Calif.
Sietz, A. J., Pennsauken, N. J.
Spalinger, E. L., Calabasas, Calif.
Stark, L., Cambridge, Mass.
Sulla, V. F., Battle Creek, Mich.
Uhlig, E. R., Becket, Mass.
Weeks, N. E., Winchester, Mass.
Whiteman, R. A., Chicago, Ill.
Wieland, J. A., Amityville, L. I., N. Y.
Yoh, J. W., New York, N. Y.
Young, C. E., Buffalo, N. Y.

Transfer to Member

Baird, J. W., Elizabethtown, N. C.
Bennett, R. A., Victoria, B. C., Canada
Churchill, D. W., Springfield, Vt.
Clark, G. J., III, Massapequa, L. I., N. Y.
Cook, G. V., Redwood City, Calif.
DeVinney, J. S., Wantagh, L. I., N. Y.
Dolbar, J. L., Northlake City, Ill.
Doll, R. E., Papillion, Neb.
Fain, D. L., Boulder, Colo.
Keating, L. M., Houston, Tex.
Kiel, J. H., Huntington Station, L. I., N. Y.
Kozicki, L. J., Omaha, Neb.
Lum, G. C., Sunnyvale, Calif.
McKee, R. M., Scottsdale, Ariz.
Oster, E. W., Jr., Lutherville, Md.
Palmer, W. L., Scottsdale, Ariz.
Perini, J., Syracuse, N. Y.
Rubin, S., Dedham, Mass.
Scharg, G. A., Westwood, N. J.
Sterner, W. J., Kansas City, Mo.
Strack, H., Rochester, N. Y.
Thomae, M. A., Buenos Aires, Argentina
Vandoch, S. C., Cincinnati, Ohio
Welling, R. L., Pomona, Calif.
Wheeler, R. D., Omaha, Neb.
Wood, E. L., Marinwood, San Rafael, Calif.
Zadro, A. P., Elmhurst, L. I., N. Y.
Zutkoff, A. L., Timonium, Md.

Admission to Member

Acker, W. M., Dallas, Tex.
Ackerley, L. D., Floral Park, L. I., N. Y.
Akamatsu, K., Los Angeles, Calif.
Allen, R. R., Waltham, Mass.
Allison, D. K., North Hollywood, Calif.
Anderl, J. H., Eatontown, N. J.
Anderman, A., Canoga Park, Calif.
Anderson, E. R., Hayward, Calif.
Andrus, P. G., Columbus, Ohio
Arden, H. G., Gillespie, Ill.
Arnold, B. J., Winston-Salem, N. C.
Ashley, F. W., Oklahoma City, Okla.
Ball, J. L., Teaneck, N. J.
Barrett, J. R., Utica, N. Y.
Barton, D. K., Seattle, Wash.
Bates, D., Montreal, P. Q., Canada
Baynham, G. E., Ottawa, Ont., Canada
Beckwith, J. F., Melbourne, Fla.
Beers, R. H., Holmes, Pa.
Benedon, A., Queens Village, L. I., N. Y.

Berman, A. L., New Brunswick, N. J.
Bernstein, R., Washington, D. C.
Berthold, J. R., Clear, Alaska
Bird, W. P., New York, N. Y.
Bischoff, R. E., Utica, N. Y.
Bjork, L. W., Tewsbury, Mass.
Blodget, D. M., Jr., Monroe, Conn.
Bondar, M. M., Los Angeles, Calif.
Bosselaers, R. J., Marlboro, Mass.
Boxenhorn, B., Long Island City, N. Y.
Brambrut, S. A., Great Neck, L. I., N. Y.
Brar, C. S., Pullman, Wash.
Bratt, P. P., Cincinnati, Ohio
Braun, P. F., Southampton, Pa.
Breeden, T. D., Moorestown, N. J.
Breisch, W. M., Jr., Philadelphia, Pa.
Brejcha, E. H., Philadelphia, Pa.
Brennan, J. P., Jr., Haddon Heights, N. J.
Bress, E., Pittsfield, Mass.
Breyer, J. P., South Norwalk, Conn.
Briggs, A. S., Rome, N. Y.
Brill, G. D., Honolulu, Hawaii
Brock, W. J., Winston Salem, N. C.
Brode, A. C., Arlington, Va.
Brown, B. C., Madison, Ala.
Brown, H. J., Havertown, Pa.
Brown, R. G., Ames, Iowa
Brun, J. H., Brooklyn, N. Y.
Bruzel, F., Philadelphia, Pa.
Buddenhagen, C. G., Falls Church, Va.
Burienne, J. L., Chester, N. J.
Cacciamani, E. R., Jr., Washington, D. C.
Callahan, J. F., New Hyde Park, L. I., N. Y.
Cameron, E. A., Prince Albert, Saskatchewan, Canada
Carmeci, P., Red Bank, N. J.
Carr, K. J., Elkhart, Ind.
Carroll, F. F., Lynnfield, Mass.
Carson, J. A., Santa Clara, Calif.
Carter, J. M., Seattle, Wash.
Case, A. P., Palos Verdes Estates, Calif.
Chalon R., Copiague, L. I., N. Y.
Chan, K., New York, N. Y.
Chapman, W. P., Jr., Morristown, N. J.
Chaudhry, K. G., Aylmer East, Que., Canada
Chen, P C., New York, N. Y.
Chesley, G. D., Teaneck, N. J.
Cinosso, J. J., Paterson, N. J.
Clark, J. O., Baltimore, Md.
Cohen, M. S., Brooklyn, N. Y
Cole, J. A., Hicksville, L. I., N. Y.
Collett, A. J., Mattapan, Mass.
Collins, L. E., Burtonsville, Md.
Condon, S. F., Haddonfield, N. J.
Cooper, F. M., San Angelo, Tex.
Costigan, W. J., Burlington, Vt.
Conrad, W. A., Plainfield, Ind.
Courtright, E. J., Buckner, Mo.
Cragon, H. G., Dallas, Tex.
Crane, L. D., Hamden, Conn.
Crews, C. W., Silver Spring, Md.
Crowder, R. S., Jr., Newark, Del
Dachan, L. S., North Wales, Pa
Dale, A. P., Salisbury, Southern Rhodesia, Africa
Darmofal, W. A., Somerdale, N. J
da Silva, J. F., West New York, N. J.
Davidson, S., Woodland Hills, Calif.
Day, C. F., Jr., Falls Church, Va
Day, J. C., Lexington, Ky.
Deas, W., Belleville, Ont., Canada
Deese, C. K., Holloman AFB, N. M.
Defonzo, G. J., Brooklyn, N. Y.
Del Guercio, V. G., Caldwell, N. J.
Deo, N., Pasadena, Calif.
Detwiler, W. S., Jr., Leesburg, Va.
Deutsch, W. E., Los Angeles, Calif.
Devoe, D. B., Stoneham, Mass.
Dierman, J. P., Waltham, Mass.
Ditum, V. F., Hackensack, N. J.
Diodato, G., Rome, Italy
Dolinko, M., New York, N. Y.
Dombrosky, E. P., Dayton, Ohio
Donato, G. R., Cambridge, Mass.
Doornebos, H. A., Watertown, Mass.
Doshay, I., Seattle, Wash.
Douglas, F. G., Lauzon Levis, Que., Canada

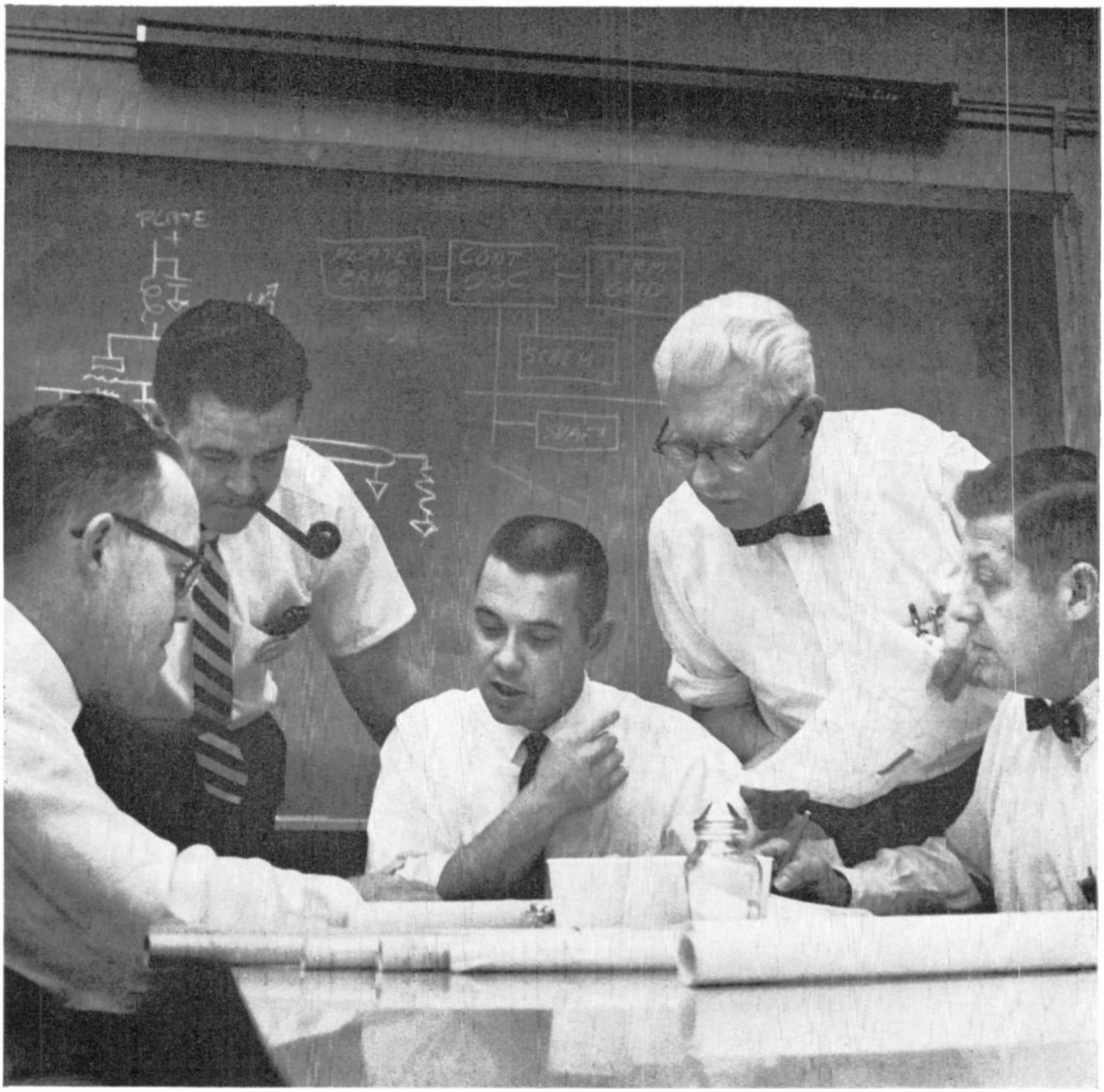
(Continued on page 120A)

Membership

(Continued from page 119A)

Doyle, L. B., East Orange, N. J.
Dubrowsky, L., Baltimore, Md.
Eckard, R. D., Livermore, Calif.
Eckberg, L. D., St. Paul, Minn.
Edwards, W. J., Chevy Chase, Md.
Ellington, R. E., Concord, N. H.
Ellis, M. C., Birmingham, Ala.
Ellis, R. R., Kansas City, Mo.
Else, J., Ottawa, Ont., Canada
Emert, G. G., Washington, D. C.
Enterline, J. R., New York, N. Y.
Epstein, H. S., Great Neck, L. I., N. Y.
Evangelista, J. J., Garden Grove, Calif.
Faller, D. M., Wayland, Mass.
Farthing, W. H., Riverdale, Md.
Fekete, R., Watertown, Mass.
Feller, R. G., Passaic, N. J.
Fields, J. R., Greensboro, N. C.
Fischbein, E., New York, N. Y.
Fleischer, R. J., Los Angeles, Calif.
Folse, R. C., Marlton, N. J.
Ford, W. W., Jr., Dalton, Mass.
Fox, E. D., Haddonfield, N. J.
Frank, K., Aberdeen, Md.
Franklin, S. G., Milford, N. J.
Frazier, W. R., Jr., Summit, N. J.
Gabusi, J. J., Brentwood, L. I., N. Y.
Gagnon, G. A., Concord, N. H.
Gaiser, W. L., Holloman AFB, N. M.
Gamzon, R. R., Rehovoth, Israel
Garrity, J. F., Jr., Baltimore, Md.
Gary, W. L., Jackson Heights, L. I., N. Y.
Gates, H. K., Oak Harbor, Wash.
Gersten, E. J., North Syracuse, N. Y.
Giacoponello, J. A., Springfield, Pa.
Gianelle, W. H., Camden, N. J.
Glasgal, R., Long Island City, N. Y.
Glaz, G., Levittown, Pa.
Glunt, J. D., Los Angeles, Calif.
Godfrey, W. R., Beverly, Mass.
Gold, J. L., Riverton, N. J.
Goldberg, A. A., Massapequa, L. I., N. Y.
Goldsmith, A. L., Haddonfield, N. J.
Goodlet, J. L., Jr., Jackson Heights, L. I., N. Y.
Goodman, P. C., Royal Oak, Mich.
Goodnight, F. E., Kannapolis, N. C.
Gordon, D., Moorestown, N. J.
Graham, K. J., Alexandria, Va.
Granborg, B. S. M., Schenectady, N. Y.
Grant, R. J., Redwood City, Calif.
Green, L. W., Eau Gallie, Fla.
Green, P. A., South Ozone Park, L. I., N. Y.
Green, W. I. L., Montclair, Calif.
Griffin, A. G., Jr., Eau Gallie, Fla.
Grigg, R. D., Haddonfield, N. J.
Guilbaud, G. L., Orsay, S. & O., France
Guillette, R. A., Methuen, Mass.
Guldstrand, R. H., Pasadena, Calif.
Gunthard, H. H., Zurich, Switzerland
Guy, E. H., Philadelphia, Pa.
Halim, M. A., Sheffield, England
Hall, A. H., Boston, Mass.
Handy, B. F., Jr., Los Angeles, Calif.
Harris, G. A., Malvern, Pa.
Harrold, J. S., Seattle, Wash.
Haselwood, J. L., Columbus, Ohio
Haspel, M. S., New York, N. Y.
Hayes, J. F., Brighton, Mass.
Haylock, G. V., Sidcup, Kent, England
Hays, C. S., Philadelphia, Pa.
Haywood, A. L., Dayton, Ohio
Hedlund, C. U., Brooklyn, N. Y.
Heitmann, R. F., Teaneck, N. J.
Henry, W. E., Jr., Charlotte, N. C.
Henschen, L. G., Indianapolis, Ind.
Hess, H. M., Holmdel, N. J.

(Continued on page 122A)



there's a young man on nearly every Collins team

At Collins, young men do not come up with *all* the new ideas, but they do create their fair share. A look at the technical progress being created at Collins is proof enough that Collins senior engineers encourage new men with new ideas.

At Collins, many great ideas come out of creative engineering teams like the one pictured above. The young man in the center, a relative newcomer at Collins, contributes materially to several new projects.

Other team members take pride in his accomplishments. They enjoy "showing him the ropes", but they consider *his* opinions, too. He belongs to their team.

A new Collins man is *new* for a very short time indeed. Collins steady growth creates a continuous need for EE's, ME's, mathematicians, technical writers and men with business training. Recent grads and experienced alumni may find Collins the ideal opportunity to build a satisfying and profitable career.

Learn more about a Collins career by writing for an illustrated brochure. To arrange a personal interview, send your qualifications to L. R. Nuss, Manager Professional Employment, Collins Radio Company, Cedar Rapids, Iowa, or E. H. Watkins, Manager, Employment, Collins Radio Company, 1220 North Alma Road, Richardson, Texas.

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COLLINS RADIO COMPANY • CEDAR RAPIDS, IOWA • DALLAS, TEXAS • NEWPORT BEACH, CALIF.

Membership

(Continued from page 120A)

Hiester, H. B., Point Pleasant, N. J.
Hite, H. F., Jr., Los Angeles, Calif.
Hof, A. R., Little Rock, Ark.
Hoffman, J. E., Hurley, N. Y.
Holmes, W. H., Northbridge, N.S.W., Australia
Holseberg, H. R., North Charleston, S.C.
Hrbek, C. J., Jackson Heights, L. I., N. Y.
Hudak, N. J., Liverpool, N. Y.
Huggard, R. F., Jericho, L. I., N. Y.
Hughes, F. W., Roseville, Calif.
Huttar, D. E., Mount Holly, N. J.
Iannotti, J. D., Omaha, Neb.
James, B. D., Palo Alto, Calif.
Janssen, A. H., Dayton, Ohio
Joglekar, P. J., Bangalore, India
Johnson, R. A., Sunland, Calif.
Jones, E. H., Jr., St. Louis, Mo.
Jones, G. S., II, New York, N. Y.
Jones, R. L., Lawnside, N. J.
Jung, W. K., New York, N. Y.
Kalinchuk, W., Indianapolis, Ind.
Kampe, T. W., Glendale, Calif.
Kaplan, B., Hollis, L. I., N. Y.
Karpeles, M. A., West Orange, N. J.
Kass, S., Framingham, Mass.
Kelliher, F. J., Concord, N. H.
Kemper, D. A., Burbank, Calif.
Kendall, D. F., Dallas, Tex.
Kessler, E. P., Playa Del Rey, Calif.
Khvostchinsky, V., San Mateo, Calif.
Kim, D., Chicago, Ill.
Kinman, D. M., Ridgecrest, Calif.
Klein, R. S., Malver, Pa.
Knochel, J. J., Baltimore, Md.
Knoll, A., Seattle, Wash.
Kogelnik, H., Murray Hill, N. J.
Korff, M., Haddonfield, N. J.
Kornfuehrer, G. R., Dallas, Tex.
Kostiw, N., Plainview, L. I., N.Y.
Kressel, H., Highland Park, N. J.
Kumar, A., Bangalore, India
Kunz, M., Fairfield, Conn.
Kuypers, J. A., San Carlos, Calif.
Laird, R. J., Plainview, L. I., N. Y.
Lanes, A., Flushing, L. I., N. Y.
Lapsins, U., Xenia, Ohio
Laubacher, J. E., LaCrescenta, Calif.
Laurens, A., Saint-Mande, (Seine), France
Lauver, H. O., Lompoc, Calif.
Lawlor, E. D., Levittown, Pa.
Leben, M., Brooklyn, N. Y.
Leichliter, C. V., LaGrange, Ill.
Lentini, J. J., Brooklyn, N. Y.
Letow, A. M., East Meadow, L. I., N. Y.
Lim, Y. S., Whippany, N. J.
Limberis, W. R., Greenbett, Md.
Linge, J. M., Long Island City, N. Y.
Liskow, C. L., Ypsilanti, Mich.
Liubinskas, J., Jr., Chicago, Ill.
Lovell, J. E., Yorktown Heights, N. Y.
Lovejoy, C. A., Eau Gallie, Fla.
Lowe, G. C., Bergenfield, N. J.
Lowrey, G. R., Jr., Fairfax, Va.
Lutton, B. W., State College, Pa.
Lynch, F. E., San Antonio, Tex.
Magleby, K. B., San Jose, Calif.
Magnuson, C. R., Chicago, Ill.
Marshall, H. B., Jr., San Diego, Calif.
Marshall, T., Jr., Rutledge, Pa.
Martin, R. J., Washington, D. C.
Mason, N. E., Ottawa, Ont., Canada
Matthews, H., Whippany, N. J.
Mayne, T. F., Pacifica, Calif.
McClain, D. B., Baltimore, Md.
McClelland, R. L., Van Nuys, Calif.
McCurnin, T. W., Tucson, Ariz.
McFadden, M. H., Belfast, North Ireland
McGrath, E., Richmond Hill, L. I., N. Y.
McKaba, E., Sunnyvale, Calif.
McKern, B. L., Haddon Heights, N. J.
McQueen, V., Seattle, Wash.
Melfi, V. A., New Hartford, N. Y.
Melroy, D. O., Springfield, N. J.
Meyer, G., San Francisco, Calif.
Miller, E. T., Haddonfield, N. J.
Miller, G. F., Quincy, Mass.
Mitchell, H. T., Nashville, Tenn.
Mitchell, W. F., Syracuse, N. Y.
Moll, K. W., Berne, Switzerland
Moore, F. E., Newark, Del.
Moreno, J. A., Claremont, Calif.
Morgan, R. D., Florissant, Mo.
Mortenson, A. R., San Francisco, Calif.
Moulden, W. W., Palo Alto, Calif.
Mowatt, P. A., Rockville, Md.
Moy, H. L., Canoga Park, Calif.
Murphy, R. H., Akron, Ohio
Murray, R. J., Long Beach, Calif.
Newton, C. L., St. Louis, Mo.
Nielsen, R. E., Minneapolis, Minn.
Norris, G. A., Columbus, Ohio
O'Gorman, A. L., Montreal, Que., Canada
Olsta, E., Bloomfield, N. J.
Orr, R. D., Corona, Calif.
Ort, H., Rolling Hills Estates, Calif.
Osadchy, V. S., Hazleton, Pa.
Owen, H. W., Silver Spring, Md.
Page, J., Clay, N. Y.
Parent, M. L., Jr., Bridgeport, Conn.
Pautler, R. A., St. Louis, Mo.
Perotto, P. G., Torino, Italy
Perrotti, E. J., Baltimore, Md.
Petty, H. C., Haddonfield, N. J.
Phillips, L. G., Mobile, Ala.
Plimpton, R. J., Oxon Hill, Md.
Poe, W. R., Adelphi, Md.
Polinsky, M. A., Newark, N. J.
Poole, L. B., Barner, N. C.
Possamai, L., Fairview, N. J.
Preston, W. L., McLean, Va.
Purdy, R. E., Indianapolis, Ind.

(Continued on page 124A)

Membership

(Continued from page 122A)

Querin, A. G., Yonkers, N. Y.
Randolph, J., Asheville, N. C.
Rath, S., Dt. Sambalpur, Orissa, India
Rekasius, Z. V., Lafayette, Ind.
Ritchie, M. L., Dayton, Ohio
Rizzi, W. J., Bedford, Mass.
Roark, M. A., Forest Hills, L. I., N. Y.
Roberts, B. B., Milwaukee, Wis.
Robinson, C. W., Jr., Uwchland, Pa.
Romer, T., Plainview, L. I., N. Y.
Rosasco, R. H., Brooklyn, N. Y.
Rosen, B. P., Williamstown, Mass.
Roush, W. B., Poughkeepsie, N. Y.
Rowell, J. H., Joliet, Ill.
Rozelle, H. G., New Orleans, La.
Rubin, A., Philadelphia, Pa.
Russell, R. K., Orlando, Fla.
Safar, J. G., Milwaukee, Wis.
Sall, I., Norwood, Mass.
Santora, C. G., Pleasantville, N. J.
Sargeant, F. W., Beaconsfield, P. Q., Canada
Schaeper, H. R. A., Powell, Ohio
Schellenberg, A. J., Winston-Salem, N. C.
Schwartz, M., Little Neck, L. I., N. Y.
Scott, D. C., Bryn Mawr, Pa.
Scott, J., Silver Spring, Md.
Scott, W. L., Whitier, Calif.
Selover, F. F., River Edge, N. J.
Shelkey, D. L., Jr., Dahlgren, Va.
Sherman, H. R., Stamford, Conn.
Shuman, B. H., Philadelphia, Pa.
Silverman, G., East Orange, N. J.
Silverstein, S., East Brunswick, N. J.
Skinner, W. L., Pacoima, Calif.
Smith, G. R., Philadelphia, Pa.
Smith, H. J., Hartford, Conn.
Smith, H. R., Sierra Vista, Ariz.
Smith, N. D., Torrance, Calif.
Smith, R. R., Syracuse, N. Y.
Sonde, B. S., Bangalore, India
Spaid, W. L., Yellow Springs, Ohio
Speltz, S. E., Los Angeles, Calif.
Stalcup, R. E., Champaign, Ill.
Stanland, J. E., Pleasantville, N. Y.
Steel, R. S., Jr., Arnold, Md.
Stephens, R. W., Dallas, Tex.
Sterman, B., Saddle Brook, N. J.
Stern, A., Katamon, Jerusalem, Israel
Stewart, J. H., South Orange, N. J.
Stewart, M. E., Mercury, Nev.
Stine, E. V., Lanham, Md.
Stumman, H. E., Los Angeles, Calif.
Sullivan, D. J., Poughkeepsie, N. Y.
Sussman, D. B., Chicopee Falls, Mass.
Svihula, G. F., Haddonfield, N. J.
Swartz, J. A., Jr., Pennsylvania Furnace, Pa.
Sweet, R. M., South Huntington, L. I., N. Y.
Swisshelm, J. T., Kokomo, Ind.
Tai, J., Dayton, Ohio
Taylor, J. L., Richardson, Tex.
Taylor, W. C., Menlo Park, Calif.
Tencza, J. W., Wallingofrd, Conn.
Tepper, T., Baltimore, Md.
Testo, C. J., Winston-Salem, N. C.
Thomas, L. A., Salisbury, Md.
Tiso, J. M., Northridge, Calif.
Tjon Pian Gi, W. E., Poughkeepsie, N. Y.
Trojanowski, M. S., Ottawa, Ont., Canada
Trowill, J. L., Jr., Pittsfield, Mass.
Turpin, J. A., Dallas, Tex.
Ujimoto, K. V., Ottawa, Ont., Canada
van Benthem, B., Chelmsford, Mass.
Van Reymersdal, J. G. A., Fairless Hills, Pa.
Vinson, H. J., North Little Rock, Ark.
Vitka, J. A., Jr., Waltham, Mass.
Waxman, M. M., Owego, N. Y.
Way, T. L., Cambria Heights, L. I., N. Y.
Weaver, J. P., Lancaster, Pa.
Webster, D. R., Baltimore, Md.

Weeks, H. L., New York, N. Y.
Westermann, K. H., Seaford, L. I., N. Y.
Whalen, A. A., Baltimore, Md.
Whalen, J. J., Baltimore, Md.
White, D. F., Dorking, Surrey, England
Whitmore, P. G., St. Paul, Minn.
Wickett, B. H., Riverside, Calif.
Williams, D. R., Collingswood, N. J.
Williams, J. R., Nashua, N. H.
Wilson, J. P. A., Hyattsville, Md.
Wilson, R. A., Southfield, Detroit, Mich.
Wilson, W. A., Orlando, Fla.
Wittig, K-H., Los Angeles, Calif.
Wittry, D. B., Los Angeles, Calif.
Wood, H. G., Asbury Park, N. J.
Wurr, J. C., Sunnyvale, Calif.
Yazombek, F. W., Chicago, Ill.
Young, W. J., North Andover, Mass.
Zachos, T. H., Astoria, L. I., N. Y.
Zavistovich, R. R., Washington, D. C.
Zimmermann, K. A., Trenton, N. J.
Zuckswert, Philadelphia, Pa.

Admisson to Associate

Acosta, J. J., Tewksbury, Mass.
Andersen, E., Copenhagen N., Denmark
Andrew, E. J., Wooster, Ohio
Apitz, J. M., Exeter, N. H.
Apostolakis, A., Montreal, Que., Canada
Armstrong, J. N., Los Angeles, Calif.
Aube, G. R., New Haven, Conn.
Avila, J. R., San Antonio, Tex.
Bandler, M. L., Rockville, Md.
Bantum, W. F., Los Angeles, Calif.
Barker, R. R., King of Prussia, Pa.
Barry, H. M. S., Agra, W. P., India
Bauler, E. K., Shreveport, La.
Baumgarth, J. G., II, Mountain View, Calif.
Beckett, F. J., Brisbane, Australia
Bennett, R. A., Victoria, B. C., Canada
Birenbach, A., Brooklyn, N. Y.
Blanden, H. E., Oaklawn, Ill.
Blau, A. B., Los Angeles, Calif.
Bourgeois, J. M. F., Bruxelles, Belgium
Bowler, R. W., Washington, D. C.
Breslau, S .W., Valley Stream, L. I., N. Y.
Brivio, D. P., Milan, Italy
Buboltz, C. P., Fullerton, Calif.
Burger, H. D., Los Angeles, Calif.
Cady, L. D., New York, N. Y.
Candiani, G., Genova, Italy
Carpenter, D. L., Pico-Rivera, Calif.
Chasson, A. T., Port Arthur, Tex.
Chuartzman, P., Buenos Aires, Argentina
Cline, W. M., Miami, Fla.
Corliss, F. B., Natick, Mass.
Cotter, P., Norwood, Mass.
Cross, A. W., North Largo, Fla.
Cummings, D. J., Chicago, Ill.
Curran, T. V., Granada Hills, Calif.
Dalgleish, D. A., Philadelphia, Pa.
Danzo, P. A., Jr., Union City, N. J.
Daponte, K. J., Boston, Mass.
Davidson, L. M., Forest Hills, L. I., N. Y.
Davis, C. G., Washington, D. C.
Davis, J. J., Cleveland, Ohio
DeClereque, K. L., St. Petersburg, Fla.
Devlin, J. N., Lakeview, N. Y.
Divieti, L. D., Milan, Italy
Dodge, J. L., New York, N. Y .
Driscoll, J. J., Burlington, Mass.
Dughi, L. J., Westfield, N. J.
Dunlap, L. A., Ontario, Calif.
Earle, J. L., San Francisco, Calif.
Ensley, S. B., Groton, Conn.
Esposito, R. V., East Orange, N. J.
Ettelson, B. L., Van Nuys, Calif.
Falcini, S., Pisa, Italy
Farmar, P. D., Escondido, Calif.
Fifer, T. J., Cleveland, Ohio
Fillingham, R. A., Denver, Colo.
Fitzpatrick, J. G., Dayton, Ohio
Foreman, J. D., Las Vegas, Nev.
Fowler, F. B., White Plains, N. Y.
Gallagher, J. L., Old Bridge, N. J.

(Continued on page 126A)

Membership

(Continued from page 125A)

Geduld, G. E., Toronto, Ont., Canada
Geisel, T., Chicago, Ill.
Gerontzos, J. P., Chicago, Ill.
Gianguzzi, L. C., New York, N. Y.
Glasgo, M. L., Omaha, Neb.
Goldammer, F. R., Ossining, N. Y.
Gombos, J. E., Upper, Montclair, N. J.
Gossett, S. D., New York, N. Y.
Gottesfeld, M. L., Laurelton, L. I., N. Y.
Greco, S. V., Lexington Park, Md.
Green, L. R., Glendale, Calif.
Greenwood, D. B., Vancouver, B. C., Canada
Hale, J. N., Montebello, Calif.
Hall, J. M., Washington, D. C.
Halsted, J. E., Baraboo, Wis.
Hancock, D. W., Rome, N. Y.
Hansen, R. P., Barksdale AFB, La.
Harker, D. R., Arlington, Va.
Harrison, W., Winston-Salem, N. C.
Hart, F. D., Newark, N. J.
Hayter, W. D., San Jose, Calif.
Heistermann, F., Ridgefield, Conn.
Hemmie, D. L., Burlington, Iowa
Hendrickson, H. L., Brooklyn, N. Y.
Hermanos, R. L., New York, N. Y.
Herron, E. H., Jr., New York, N. Y.
Hess, R. L., Norfolk, Va.
Hockersmith, W. E., Jackson Heights, L. I., N. Y.
Hoffman, A. G., Fremont, Neb
Hoffman, K. L., Fremont, Neb.
Hollis, E. L., Van Nuys, Calif.
Horan, L. W., Minneapolis, Minn.
Hunt, W. C., Anchorage, Alaska
Hunter, G. B., Vancouver, B. C., Canada
Hurtubise, W. E., Fargo, N. D.
Hurley, D. F., Lynnfield Centre, Mass.
Hich, D. F., Flushing, L. I., N. Y.
Itchkow, K. M., Levittown, L. I., N. Y.
Jako, G. J., Boston, Mass.
Jalaluddin, Troy, N. Y.
Jenkins, P. R., Goleta, Calif.
Johnson, J. A., Baltimore, Md.
Johnson, W. H., Washington, D. C.
Johnston, R. C., Pomona, Calif.
Jones, T. T., Poughkeepsie, N. Y.
Katz, N. I., Jackson, Heights, L. I., N. Y.
Kaytar, J. J., Edison, N. J.
Key, B. G., Greencastle, Pa.
Khoury, G. J., Jr., Westford, Mass.
Kienzle, G. C., Lindenhurst, L. I., N. Y.
King, F. W., Nanaimo, B. C., Canada
Korabell, I., Buffalo, N. Y.
Kosty, W. G., Winnipeg, Man., Canada
Krieger, A. E., Wright-Patterson AFB, Ohio
Krishman, V., Egmore, Madras, India
Krishnaswamy, V., Pilani, Rajastan, India
Kroupa, R. G., Middle Village, L. I., N. Y.
Krug, W. H., Janesville, Wis.
Laederich, G., Paris, France
Laut, S., Baltimore, Md.
Laychak, S. F., Norwalk, Conn.
Lazarus, L. S., Rosedale, L. I., N. Y.
Lemon, E. A., Los Angeles, Calif.
Lenkey, O. A., Hillside, N. J.
Li, H-Y., Boston, Mass.
Lindgren, E. A., Culver City, Calif.
Luce, J. E., Tampa, Fla.
Lykins, W. S., Haddonfield, N. J.
Macdonald, K. G., Downsview, Ont., Canada
Mace, A. E., Cheyney, Pa.
Magnatta, V. B., Palisade, N. J.
Mann, J., New York, N. Y.
Marraccini, A. J., Staten Island, N. Y.
Martin, A., Philadelphia, Pa.
McAvoy, W. R., Studio City, Calif.
McBurney, F. J., Brooklyn, N. Y.
McGuire, J. C., Waldwick, N. J.
McRitchie, K., Dayton, Ohio
Mendez, M. R., Las Salinas, Chile
Mielen, R. V., Victoria, B. C., Canada
Miesch, R. A., Grabill, Ind.
Miller, C. E., Culver City, Calif.
Mittnacht, A. J., San Jose, Calif.
Mock, G. W., Inglewood, Calif.
Moher, T. F., Poughkeepsie, N. Y.
Moore, R., Tekamah, Neb.
Morgan, J. E., Arlington Heights, Ill.
Morris, J. E., Whippany, N. J.
Moyer, L. N., Los Angeles, Calif.
Natale, S. V., New York, N. Y.
Nielson, R. W., New York, N. Y.
Nordbo, O. J., Seattle, Wash.
Norman, P., Jr., Trotwood, Ohio
Nys, A. D., Concord, N. H.
Oberste, J. A., Little Rock, Ark.
Pendergraph, G. S., Arlington, Va.
Penny, J. R., Jr., Miami, Fla.
Philbrick, E. L., Jr., Wellfleet, Mass.
Piatoff, E. B., Brooklyn, N. Y.
Pradervand, M., Princeton, N. J.
Prodan, R., Buenos Aires, Argentina
Pyenson, J. I., Toms River, N. J.
Remeikis, A. A., Toronto, Ont., Canada
Riddle, R. L., Kannapolis, N. C.
Ringelheim, I. B., New York, N. Y.
Roberts, W. S., Philadelphia, Pa.
Rodgers, J. C., Emporium, Pa.
Roll, J. L., McLean, Va.
Ross, R. R., Albuquerque, N. M.
Rubin, M. S., Framingham Center, Mass.
Sadler, J. A., Rexdale, Ont., Canada
Salsberg, A. P., Jackson Heights, L. I., N. Y.
Saltzman, H. S., Bayside, L. I., N. Y.
Seifert, N. A., Brookline, Mass.
Serchuk, A., Jackson Heights, L. I., N. Y.
Shively, G. C., Alamogordo, N. M.
Simonsen, P. W., Valley Stream, L. I., N. Y.
Skala, D. J., Washington, D. C.
Skinkle, G. W., Lanham Park, Md.
Slye, R. E., Jr., Washington, D. C.
Smiley, B. E., Burlington, Mass.

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Section Meetings

ALAMOGORDO-HOLLOMAN

"Antennas in Conducting Media" or "Communication Between Submerged Submarines," R. K. Moore, Univ. of N. Mex.; Election of Officers. 3/21/61.

ATLANTA

Tour of F.A.A. Airway Control Center. 3/31/61.

BAY OF QUINTE

"Origin and Structure of the Universe," G. A. Harrower, Queen's Univ. 3/15/61.

BEAUMONT-PORT ARTHUR

"Recent Developments in Electronic Computers," J. G. Kamena, Remington Rand Univac; "Communication Satellite," Phillip Yearly, Lamar College. 4/4/61.

BENELUX

"Human Reaction Times & Their Electronic Measurement," J. F. Schouten, Inst. for Perception Res., Netherlands. 4/6/61.

CENTRAL FLORIDA

"Systems Applications of Electron Beam Parametric Amplifiers," Robert Adler, Zenith Corp. 3/16/'61.

CENTRAL PENNSYLVANIA

"Infrared Physics & Modern Technology," William Birtley, HRB-Singer, Inc. 3/21/61.

CHICAGO

"How to Get Along with Your Boss or Be One!", G. S. Speer, Illinois Inst. of Tech. 11/'11/60.

"Properties & Uses of Infrared Radiation," J. A. Sanderson, U. S. Naval Res. Lab. 12/9/60.

"World-Wide Communications Using Satellite Repeaters," R. D. Campbell, AT&T. 1/13/61.

CINCINNATI

"Radio Communication Via Satellites," J. L. Glaser, Bell Tele. Lab.; "Community Renewal Program for Cincinnati," C. H. Stamm, City of Cincinnati. 3/23/'61.

CLEVELAND

"Guided Plasma Waves," O. K. Mawardi, Case Inst. of Tech. 3/9/'61.

COLUMBUS

"Common Stocks & Uncommon Profits," Roger White, Paine, Webber, Jackson & Curtis. 4/12/61.

DALLAS

"Relationship Between Devices, Models and Networks," J. G. Linvill, Stanford Univ. 3/28/'61.

DETROIT

Panel Discussion on Engineering Writing and Speech: T. M. Farrell, Michigan State Univ., S. S. Attwood, Univ. of Mich., C. J. Freund, Univ. of Detroit, J. S. Johnson, Wayne State Univ., W. P. Smith, Mich. State Univ. 1/20/'61.

"Infrared & Optical Masers," D. F. Nelson, Bell Tele. Labs. 3/17/61.

EL PASO

"Communication Peculiarities of White Sands Missile Range," Pat Walton, General Electric Co. 3/23/61.

EMPORIUM

"Communications Engineering for Ballistic Missile Early Warning System," Western Elec. 3/21/61.

Fort Huachuca

Film: "Gateways to The Mind" Bell Tele. Labs., Jim Virden, U. S. Army Electronics Proving Ground, 3/27/61.

Gainesville

"Recent Concepts in Microwave Crossed Field Tubes," J. A. Saloom, S.F.D. Labs. of Varian Assocs. 4/12/61.

Hamilton (Ontario)

"Applications of Ultrasonic Energy," C. J. McReynolds, Westinghouse Co., Ltd. 2/6/61.

"Communications in Antarctica," M. Cosgrove, CHCH TV, 3/20/61.

"Japan," R. M. Robinson, Canadian GE Co., Ltd.; Election of Officers, 4/10/61.

Houston

"Why Transistors in VHF Radio?", by Mike Bennett, for GE Co., J. H. Wofford Radio Communications Services; "Fundamentals of Correlation Functions," W. P. Schneider, Schlumberger Well Surveying Corp.; "Iterative Differential Analyzer Function & Control," M. C. Gilliland, Beckman Instruments; "Transistor Specifications & Parameters," Don Able, Texas Instruments, 3/14/61.

"A Process Dynamics Recording Trailer As Used by Monsanto Chemical Co.," W. R. Isaacs, Southwestern Industrial Electronics Co. 3/21/61.

Indianapolis

"Scan Conversion for Bright Tube Display," W. E. Miller & Lawrence Ii, Hazeltine Tech. Devel. Ctr. 3/16/61.

Ithaca

"Superconductivity" E. C. Satterthwaite, Westinghouse Elec. Corp., Joint meeting with AIEE, 3/3/61.

Kansas City

Student Award Presentation; Bell Tele. movie "Sound Reproduction." 4/11/61.

Kitchener-Waterloo

"Digital Computer," Douglas Lawson, Univ. of Waterloo; "Analogue Computer," George Fleming, Univ. of Waterloo. 3/20/61.

Las Vegas

"Instrumentation Systems of the Tonopah Missile Range," J. C. Eckhart, Sandia Corp.; Tour of Tonopah Missile Range. 3/10/61.

Little Rock

"Stereo Geometry," P. W. Klipsch, Klipsch Assocs. 3/17/61.

London (Ontario)

Student Awards. 2/27/61.

"DRITE Topside Sounder Satellite," R. K. Brown, Defence Res. Telecommunications Establishment. 4/10/61.

Long Island

Six Lectures on Microwave Measurements:

"Impedance," Tore Anderson, FXR, Inc. 2/2/61.

"Frequency & Q," Raymond Svenson, Bomac Labs. 2/9/61.

"Power," L. H. Fisher, Poly. Res. & Devel. 2/16/61.

"Loss, Gain & Phase," Patricia Loth Wheeler Labs., Inc. 2/23/61.

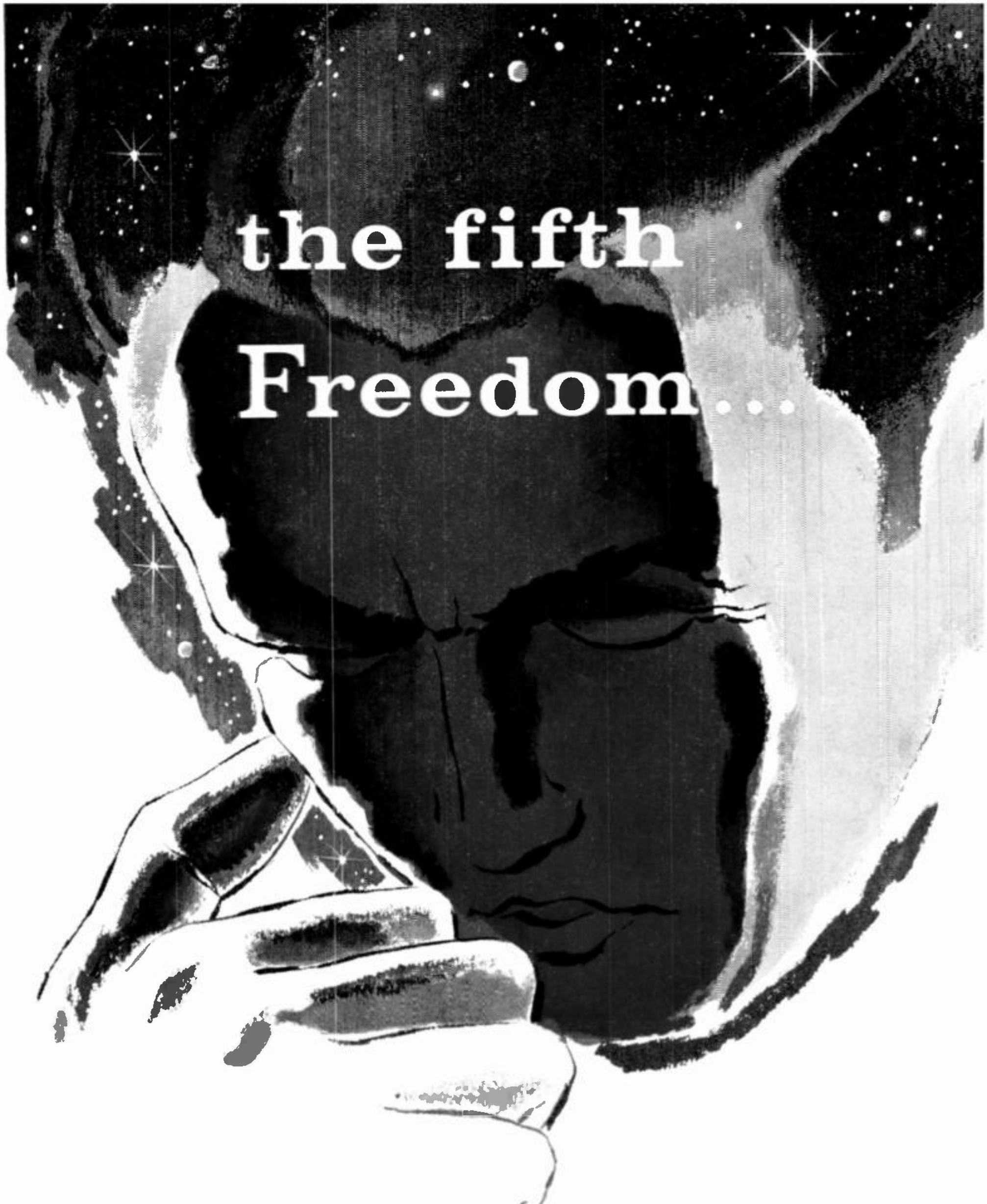
"Noise Figure," Matthew Lebenbaum, Airborne Instruments Lab. 3/2/61.

"Radiation & Propagation," Christian Berger, CDB Enterprises. 3/9/61.

Fellow Award Presentation. 3/19/61.

"Discoverer Satellite Program," R. Flothow Lockheed Aircraft Co. 4/11/61.

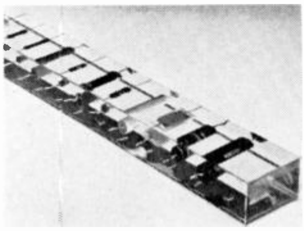
(Continued on page 135A)



NEWS New Products

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

Transparent Encapsulant



Cast in a flat block for processing, cured encapsulant is flexible.

When cured, encapsulant can be twisted or rolled to conform with unique circuit placement requirements.

A new silicone encapsulating material that permits visual inspection of circuits and components within potted, embedded or encapsulated assemblies was introduced at the IRE Convention in New York in March. Developed by **Dow Corning Corp.,** Midland, Mich., the new encapsulant will be designated as Sylgard 182 Resin.

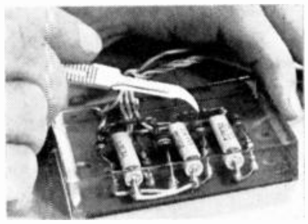
Applied as an almost colorless liquid—after blending with its curing agent—Sylgard 182 cures in place even in totally confined enclosures, to form a transparent mass having good dielectric properties, good moisture resistance and the physical attributes of flexibility and toughness.

Curing time for Sylgard 182 Resin can be varied by changing the curing temperature. At 150°C, it cures in 15 minutes; at 65°C, four hours; at 25°C, three days. Neither the resin nor its curing agent is toxic to the skin. No toxic fumes are given off during mixing or curing.

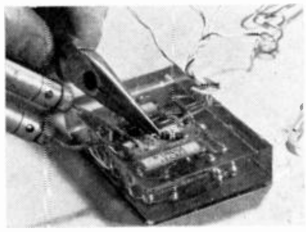
Components and circuits encapsulated in Sylgard 182 Resin are clearly visible, simplifying replacement or repair procedures. The cured resin can be cut away with a sharp knife so that defective components can be repaired or replaced. New resin poured over the repaired area will adhere to the original material, restoring the encapsulant to the original condition.

While this new material exhibits some rubber-like properties, it is not a silicone rubber, but a flexible silicone resin. This difference in basic chemical structure results in a toughness not found in present silicone rubbers.

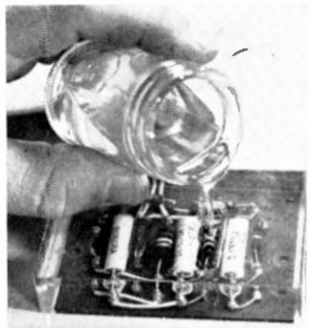
Additional information on this new transparent silicone encapsulant is available from the firm.



This resin can be cut away from around defective part with a sharp knife.



Defective part is removed and its replacement soldered into the circuit without damaging encapsulant.



After circuit repair, additional resin is poured over replacement part. Replaced material bonds to original encapsulant.

(Continued on page 132A)

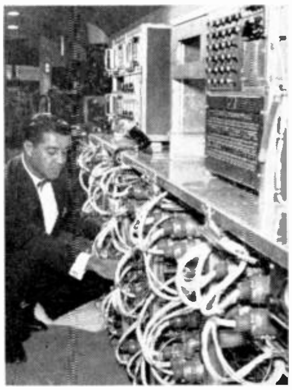
NEWS New Products

(Continued from page 130A)

Airborne Missile Monitoring Stations

Two airborne missile monitoring stations—airplanes that "hear" and "talk" like a missile in flight—are operating over the 400-mile U. S. Air Force Eglin Gulf Missile Test Range off the Gulf of Mexico coast of eastern Florida.

They are T-29s, the Air Force navigator-trainer version of the Convair 240 twin-engine transport, which have undergone complete modification at **Convair (San Diego) Div., General Dynamics Corp.**, San Diego, Calif. The modification converted the planes into airborne stations for frequency monitoring and interference control (FMIC).



These are the first complete airborne monitoring stations ever devised. The stations enable the missile ground control stations to evaluate the conditions under which a test missile is receiving command signals. Control of a missile's flight is difficult from ground stations situated below the missile's horizon and interference from other ground stations while the missile is flying cannot be detected at the launch site.

Three missions for the FMIC airplanes are outlined here:

1. They check out—by flying the missile range—ground electronic systems used to support missile tests. This includes radar, telemetering instrumentation and radio communication.
2. They monitor during countdown the frequencies being used for a missile test. If the channels are not free of interference, a countdown may be stopped, thus avoiding a launching under unreliable conditions which might mean loss of the missile.
3. During the missile firing the airplane loiters at a specific altitude and to one side of the course. In this position the FMIC aircraft is able to hear and record ground station commands given the missile; hear and record signals from the missile including telemetered information; and hear any interfering signals received by the missile which originated from other stations and were not intended for the missile.

The FMIC airplane can talk like a missile by transmitting directly to ground stations. If the missile hears something not intended for it and malfunctions as a result, this fact can be discovered and reported by the airplane.

It was noted that an FMIC airplane would be useful not only at the 400-mile Eglin range used primarily for medium and small missiles, but it is proposed that the aircraft be used for down range checking at all missile test and launching ranges. It may also be used for checking the functional and electronic environment of DEW Line installations and other defense system ground installations equipped with such components as SAGE and Texas Towers.

At Eglin the two FMIC airplanes are flown by the Air Proving Ground Center, Air Research and Development Command.

Push-Button Communications Device

Development of a push-button communications device that enables a jet pilot to "talk" to his base with greater speed, more accuracy and assurance that his message will cut through heavier atmospheric disturbance was disclosed today by **Hughes Aircraft Company.**

Called "Digikey" by its inventors, the device permits a pilot or crewman to "talk" through heavy electrical interference at a rate of 150 words per minute (twice normal speed) by punching keys on a small semi-automatic keyboard. The device is said to be so simple that anyone can be trained to operate it in a matter of minutes.

Digikey was recently installed in a Strategic Air Command aircraft and subjected to a week-long airborne evaluation by SAC officers and technicians.

Digikey converts a message into digital form and transmits through a communications system to a ground station where the digits are reconverted into words and displayed in printed form on a receiver unit. Large volumes of information can be sent in a short time with reduction in chances of human error or ambiguity. The system operates as follows:

A belt, or roll chart, on which message words are printed covers the keyboard with words di-

(Continued on page 135A)

Professional Group on Broadcast and Television Receivers

The field of broadcast receivers is one which is closely associated with the general public, perhaps more so than any other branch of the radio engineering field. In fact, to the layman the term "radio" is synonymous with "broadcast receiver."

As a result, the receiver engineer has been concerned with an additional factor not generally common to other fields, namely, that of responding to—or endeavoring to create—public demand for a product. This factor has played a prominent role in such developments as FM, car radios, portable receivers, and black-and-white television sets. It is now conspicuously evident in connection with current efforts to produce and market color television receivers and stereo equipment.

The IRE Professional Group on Broadcast and Television Receivers is playing a major role in making available vitally needed technical information, not only on color television and stereo, but on all aspects of the receiver field. Through this exchange of information, the radio and television industry is gaining important data which will be helpful in solving the engineering problems it faces and in successfully meeting the "public demand" factor mentioned above.

The Group has been particularly active in sponsoring technical sessions at most of the national meetings held throughout the country during the year: the Radio Fall Meeting, the Spring Television Conference in Chicago, the IRE International Convention, and Wescon to mention but a few.

The Group also publishes its own technical publication, called Transactions, which is distributed to some 1900 members as a part of their $4 assessment fee. The Transactions has become a chief source of information on the latest technical developments in the field of broadcast and television receivers.

Ernst Weber
Chairman, Professional Groups Committee

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 132A)

rectly over the keys. The operator is, in effect, given a "multiple choice," and selects a word from the first group displayed and punches that key, registering the word in the system's memory. The belt advances to a new set of words and another message is selected. At the same time other information from the aircraft's instruments is being recorded in the memory. On completing the message the operator presses a "transmit" button and the message is automatically broadcast as many times as desired.

A weather report used as an example:

The operator would punch a weather report key. Digikey would automatically transmit the aircraft's call letters and identification from the system's memory, the plane's position, heading and altitude as furnished by the ship's instruments.

The keyboard would advance to display a selection of words to describe the nature of the weather such as moderate winds, poor visibility, electrical storm and the like. The next display would describe precipitation, then cloud formation, the latter illustrated by pictograms, symbols illustrating the formations.

One of the applications of Digikey is its use with the Hughes-built long-range communication system, "Hacon," now operational with the U. S. Air Force's B-58 Hustler supersonic bomber.

Digikey and a digital converter allow Hacon to talk digits as well as voice. The difficulty of communicating through heavy electrical interference is overcome by breaking up a message into a series of coded electrical pulses which are transmitted over several frequencies. Only a few pulses need to get through for the receiver to recognize the character transmitted.

Information from a large number of aircraft can be fed directly into automatic command and control systems for data processing because there is no need to translate the digits into computer language.

Section Meetings

(Continued from page 129A)

LOS ANGELES

"The PERT System-Management by Computer," J. G. Sliney, Hughes Aircraft Co.; Presentation of Fellow Awards; Presentation of Student Awards. 3/7/61.

LUBBOCK

Joint Meeting with Student Branch & AIEE. 3/13/61.

MIAMI

"Man in Space," G. M. Knauf, USAF, MC. 2/27/61.

"The Space Program at Motorola," J. F. Byrne, Motorola. 3/28/61.

MOBILE

"World Wide Activities for Defense," J. R. Booth, Philco Corp.

NEW ORLEANS

"Latest Developments in Microwave Transmission, Transistors & Solar Batteries," Tom Mardis, Western Elec. Co. 3/16/61.

NORTH CAROLINA

"Magnetic Properties as Determined by Neutron Diffraction," T. J. Turner, Wake Forest College. 3/17/61.

"Transistor Physics, Manufacturing & Circuitry," W. F. Dimick, Kellogg Div. of IT&T. 4/14/61.

NORTHWEST FLORIDA

"Communication System for Eglin Gulf Test Range," J. J. Schauble, Range Devel. Div., APGC. 1/31/61.

"Standardization of Semiconductor Devices," Harold Champion, Martin Co. 2/28/61.

"Orbital Behavior of Artificial Satellites," P. M. Fitzpatrick, Ballistic Branch, Eglin AFB. 3/30/61.

OKLAHOMA CITY

"Electronic Weapons Systems—Key to Defense," Harold Schutz, Westinghouse Elec. Co., Joint meeting with AIEE. 1/16/61.

"Wave Motion," George Gibson, American Tele. Co., 2/13/61.

"Bioelectronic Telemetry System," M. C. Oviatt, FAA Civil Aeromedical Inst. 3/13/61.

OMAHA-LINCOLN

"Contributions of Military Professional Societies to Our National Security," J. F. Byrne, Motorola, Inc. 3/17/61.

ORLANDO

"Satellite Communication System," G. S. Shaw, Radiation, Inc. 2/22/61.

"Electron Beam Parametric Amplifier," Robert Adler, Zenith Radio Corp. 3/15/61.

OTTAWA

Student's Night—Presentation of papers. 2/9/61.

"Electron Beam Parametric Amplifiers," F. Osborne & A. B. Cutting, Canadian Marconi Co. 3/2/61.

"Frequency Control Systems," M. B. Bloch, Bulova Watch Co. 4/6/61.

PHILADELPHIA

"Project OZMA, A Search for Intelligent Life in Outer Space," F. D. Drake, National Radio Astronomy Observatory. 3/8/61.

PORTLAND

"Silicon Photovoltaic Power Sources in Space Vehicles," Gerald L. Pearson, Stanford Univ. 1/19/61.

Report to Local High School Students on Engineering as a Profession, Glen M. DeKraker, Sangamo Elec. Co. 2/22/61.

"A Career of Opportunity," Panel Discussion: D. C. Strain, Cullen Macpherson, Francis McCann, Tektronix, Inc. 2/25/61.

"Ferromagnetism & Ferromagnetic Domains," J. K. Galt, Bell Tele. Labs. 3/16/61.

PRINCETON

"An Implantable Cardiac Pacemaker," Wilson Greatbach, Wilson Greatbach, Inc. 2/9/61.

QUEBEC

"Hydro-Quebec Communication Systems," J. P. Geuvremont, Quebec Power Co. 3/13/61.

REGINA (Canada)

"The Communication Aspects of the Power Sub-Station Control," Harold Kaldor, Saskatchewan Power Corp. 11/22/60.

(Continued on page 136A)

Section Meetings

(Continued from page 135A)

ROCHESTER

"Generation of Electric Power in Outer Space," H. W. Paige, GE Co. 3/9/61.

SALT LAKE CITY

"Instrument System at Marquardt," Don Ord, Marquardt Jet Lab.; Film: "Ram Jet Engines" & "Capabilities of Marquardt Test Facilities." 2/9/61.

"Synchronized Sampling of Repetitive Wave Forms," R. C. Woodbury, Brigham Young Univ. 3/9/61.

"Microwave Power Tubes" (Illustrated Lecture), C. W. Carnahan, Varian Assocs. 3/30/61.

"130th Aircraft Control & Warning Squadron Group Activities," William Evans, Utah Air National Guard; Tour of Utah Air National Guard Base. 4/13/61.

SAN DIEGO

"Doppler Navigation," C. N. Bates, Ryan Electronics. 4/5/61.

SCHENECTADY

"Chemical Plants & Processes; a Challenge for Control Engineering," Walerian Kipinak, MIT. 2/14/61.

"Space Telemetry," Conrad Hoeppner, Radiation, Inc. 3/14/61.

SHREVEPORT

Student Award Presentation; Joint meeting with Louisiana Tech AIEE, Shreveport AIEE and Monroe AIEE. 3/14/61.

"Instrumentation in Medicine," L. A. Geddes, Baylor School of Medicine. 4/4/61.

SYRACUSE

"Early Days of Radio and the Events Leading up to the Formation of the Syracuse Section," C. A. Priest, Onondaga Pottery Co.; Election of Officers. 3/16/61.

"Ultrasonic Navigation & Detection Systems of Bats & Moths," Kenneth Roeder, Tufts Univ. 3/20/61.

TUCSON

"Performance of Tiros I," A. Schnapf, RCA. 3/17/61.

TWIN CITIES

"Multi-Computer Systems," George Chapin, Remington Rand Univac. 2/22/61.

VANCOUVER

"Surveillance Radar," W. K. Newton, Vancouver Airport. 3/20/61.

VIRGINIA

"Satellites as Navigational Aids," J. L. Loeb, U. S. Navy. 2/24/61.

"Digital Communications, Future Techniques for Strategic Air Command," G. A. Kious, E. J. Braunner, H. C. Drehr, GE Co. 3/10/61.

"Digital Tropospheric Scatter," J. R. Poppe, GE Co. 3/17/61.

WASHINGTON

"Peaceful Co-existence of Engineers & Writers," A. G. Norris, Vitro Engrg. Co. 3/17/61.

WESTERN MASSACHUSETTS

"Use of Electronics in Crime Detection," R. Millen, FBI. 4/5/61.

SUBSECTIONS

BUENAVENTURA

"The Molecular Electronics Approach to Micro-miniaturization," W. B. Hugle, Westinghouse Elec. Corp. 3/8/61.

EASTERN NORTH CAROLINA

"The Nike Zeus Story," F. E. Nimmcke, Bell Tele. Labs. 3/10/61.

FAIRFIELD COUNTY

"Micro Module Construction and Use," V. J. De Fillipo, RCA; Election of Officers. 10/20/60.

"Microwave Modulation of Light," Nicolaas Bloembergen, Harvard Univ. 11/17/60.

Field trip to Indian Point Atomic Reactor. 12/10/60.

"Dystac Computer by Fast Repetitive Operation of Analog Computers," Irwin West, Computer Systems, Inc. 2/28/61.

LANCASTER

"The Challenge of The Deep," B. F. McMahon, RCA. 2/28/61.

"Scientific Aids to Crime Detection," R. L. Millen, FBI. 3/28/61.

LEHIGH VALLEY

"Recent Advances in the Development of Miniature Electron Tubes," J. E. Beggs, GE Res. Labs. 2/15/61.

"Electronics in the Steel Mill," A. C. Chamberlin, Bethlehem Steel Co. 3/22/61.

MEMPHIS

Tour of AT&T Plant. 3/27/61.

MID-HUDSON

"Electronic Switching Systems" A Discussion of Installation at Morris, Ill., B. J. Yokelson, Bell Tele. Lab. 3/16/61.

"Scientific Techniques of Criminal Investigation," I. W. Conrad, FBI; Election of Officers. 4/18/61.

ORANGE BELT

"Around the World in 90 Minutes—Project Mercury," E. R. Hinz, Convair Astronautic Div., General Dynamics Corp. 3/21/61.

PALM BEACH

"Air Traffic Control," General Queseda, USAF, (ret.) 2/23/61.

Plant Tour of Franklin Systems, M. Cohen Franklin Systems. 3/21/61.

PANAMA CITY

"The Technical Education Program at Gulf Coast Jr. College," Richard Morley, Gulf Coast Jr. College. 3/30/61.

"A Miniaturized Radio Transmitter for Location Purposes," C. D. Canfield, Seattle Devel. Lab. 4/13/61.

SAN FERNANDO VALLEY

"Exploration of the Moon," Richard Davies Jet Propulsion Lab. 3/15/61.

SANTA ANA

"Oceanography—Hole in the Noosphere," David Heebner, Hughes Aircraft Co. 2/21/61.

"Bionics—Promise or Menace?", Frank Rosenblatt, Cornell Aeronautical Lab. 3/7/61.

SANTA BARBARA

"Tunnel Diode Applications," Victor Van-Duzer, Hewlett-Packard. 2/24/61.

"National Security & Arms Control," Richard Raymond, GE Tempo. 3/8/61.

WESTCHESTER

Field Trip—Indian Point. 3/18/61.

WESTERN NORTH CAROLINA

Color Films- "Seconds for Survival" and "Project Echo"; Election of Officers. 12/9/60.

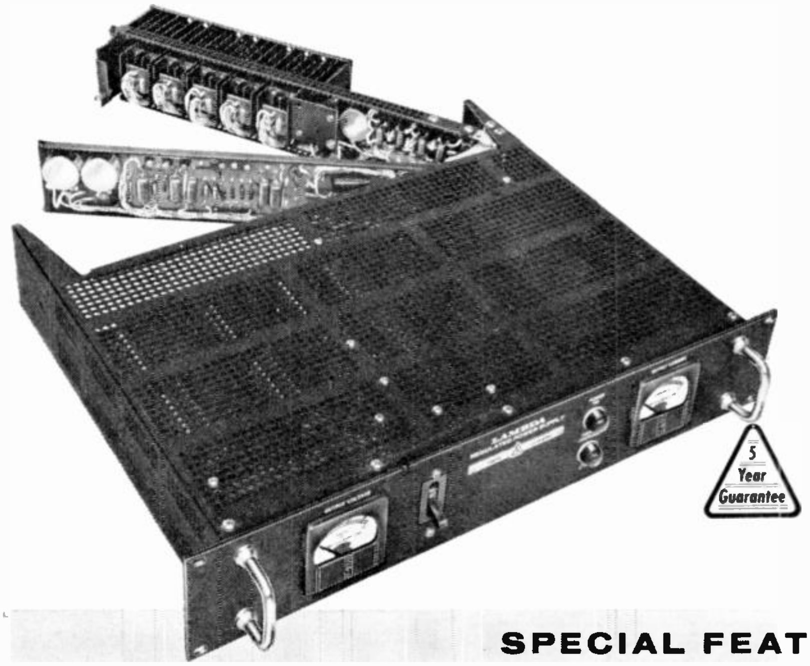
"Radio Aids to Navigation," W. B. Daniels, Federal Aviation Authority, Douglas Airport. 3/24/61.

LAMBDA

Convection Cooled Transistorized Regulated Power Supplies

LA SERIES

0- 34 VDC	5 AMP
0- 34 VDC	10 AMP
0- 34 VDC	20 AMP
20-105 VDC	2 AMP
20-105 VDC	4 AMP
20-105 VDC	8 AMP
75-330 VDC	0.8 AMP
75-330 VDC	1.5 AMP
75-330 VDC	3 AMP



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SPECIAL FEATURES

- Convection Cooled—No internal blowers
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- No Voltage Spikes or overshoot on "turn on, turn off," or power failure
- Short Circuit Proof
- Guaranteed 5 years
- Remote programming over Vernier band
- Hermetically-sealed transformer designed to MIL-T-27A
- Easy Service Access
- Constant Current Operation—Consult Factory

CONDENSED DATA

DC OUTPUT (Regulated for line and load)

Model	Voltage Range (1)	Vernier Band (2)	Current Range (3)	Price (4)
LA 50-03A	0- 34 VDC	4 V	0- 5 AMP	$ 395
LA100-03A	0- 34 VDC	4 V	0-10 AMP	510
LA200-03A	0- 34 VDC	4 V	0-20 AMP	795
LA 20-05A	20-105 VDC	10 V	0- 2 AMP	350
LA 40-05A	20-105 VDC	10 V	0- 4 AMP	495
LA 80-05A	20-105 VDC	10 V	0- 8 AMP	780
LA 8-08A	75-330 VDC	30 V	0- 0.8 AMP	395
LA 15-08A	75-330 VDC	30 V	0- 1.5 AMP	560
LA 30-08A	75-330 VDC	30 V	0- 3 AMP	860

(1) The DC output voltage for each model is completely covered by four selector switches plus vernier range.
(2) Center of vernier band may be set at any of 16 points throughout voltage range.
(3) Current rating applies over entire voltage range.
(4) Prices are for unmetered models. For metered models add the suffix "M" and add $30.00 to the price.

Regulation (line)....... Less than 0.05 per cent or 8 millivolts (whichever is greater). For input variations from 100-130 VAC.

Regulation (load)...... Less than 0.10 per cent or 15 millivolts (whichever is greater). For load variations from 0 to full load.

Ripple and Noise....... Less than 1 millivolt rms with either terminal grounded.

Temperature Coefficient........... Less than 0.025%/°C.

AC INPUT 100-130 VAC, 60 ± 0.3 cycle(5)

(5) This frequency band amply covers standard commercial power line tolerances in the United States and Canada. For operation over wider frequency band, consult factory.

Size

LA 50-03A, LA20-05A, LA 8-08A	3½" H x 19" W x 14⅜" D
LA100-03A, LA40-05A, LA15-08A	7" H x 19" W x 14⅜" D
LA200-03A, LA80-05A, LA30-08A	10½" H x 19" W x 16½" D

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LA118

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(Continued from page 79A)

George F. Houlroyd (A'49–M'55), formerly plant manager, has been elected Vice President—Manufacturing for Foto-Video Electronics, Inc., Cedar Grove, N. J., according to a recent announcement.

G. F. HOULROYD

From 1949, until he joined the staff of Foto-Video in 1959, he served as plant manager for Boonton Radio Corporation, where he advanced techniques for the manufacture of electronic instruments. During this period he also instituted safety procedures that won recognition for the company through awards by the Liberty Mutual Insurance Company and the New Jersey Department of Labor and Industry.

Since joining Foto-Video Electronics, he has instituted and directed many production improvements. Notable among these have been the application of production equipment and techniques that have greatly increased production capacity and worker efficiency.

He began his career in the production of electronic equipment when he joined the staff of Hardwick-Hindle, Inc., as a foreman of resistor manufacturing in 1927. Eight years later he became plant superintendent and foreman for Foote-Pierson and Company, Inc., where he specialized in the production of precision gaging for electronic products and instruments.

Mr. Houlroyd is a director of the Boonton, N. J., Safety Council and Boonton Evening Adult Education.

❖

Stephen J. Jatras (M'52–SM'60), who had been guiding the expansion of the Avionics and Industrial Products Division of Lockheed Electronics Company for nearly a year as Acting General Manager, has been named General Manager of that Division and Vice President of Lockheed Electronics Company.

S. J. JATRAS

The Avionics and Industrial Products Division is responsible for the development and production of quality electronic components for industry and the military.

Mr. Jatras, who joined the Lockheed organization in 1956, has held several engineering and administrative positions with Lockheed's Missiles and Space Division and the former Lockheed Electronics and Avionics Division, from which LEC's Avionics and Industrial Products Division was organized.

Before coming to Lockheed, he was associated with Midwestern Instruments in Tulsa, Okla., where he served as Vice President and Chief Engineer. He was graduated from the Carnegie Institute of Technology, Pittsburgh, Pa., with the B.S. degree in electrical engineering and received the M.S. degree in electrical engineering from the Massachusetts Institute of Technology, Cambridge. In 1958, he was appointed a Sloane Fellow at the Stanford University Graduate School of Business.

❖

James H. Johnson (M'58) has joined Eitel-McCullough, Inc., as a senior sales engineer, according to a recent announcement. He has been assigned to the company's Los Angeles, Calif., sales office. His sales territory includes southern California, Arizona, and southern Nevada.

J. H. JOHNSON

He was previously Chief Project Engineer with Ling Electronics, a design engineer with Hughes Aircraft Company and the General Electric Company. He received the B.S. degree in electrical engineering from Mississippi State College, in State College.

❖

Allen H. Kaltman (A'60) has been appointed a Senior Member of the Product and Market Planning Staff, Auerbach Electronics Corporation, Philadelphia, Pa., and New York, N. Y. He will conduct his activity from the New York offices of the corporation where he is engaged in studies of competitive developments in data processing utilization in several industry and market segments. He is also working on diversification programs for major manufacturers in the electronics field and is developing marketing requirements for high-speed digital building blocks, input-output equipment, and other electronic data processing equipments.

He has had many years of experience in product and market planning. He was formerly Vice President of Charles Roberts Associates, Inc., an industrial marketing consulting firm, and Research Director of the New York World Telegram and Sun. He has conducted extensive studies involving the needs for supervisory control systems, identifying market opportunities for their utilization. He has performed market evaluations of a wide variety of electrical and electronic devices and components, and has evaluated product lines of sophisticated electronic equipment.

Mr. Kaltman received the Bachelor of Science degree in Economics and Statistics from Rutgers University, New Brunswick, N. J. He is a member of the American Statistical Association and the American Marketing Association.

❖

(Continued on page 140A)

FABRICATING MICROMODULES TO MOBILE ROOMS

Magnetic Shields Custom Fabricated to Any Size or Shape

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- Co-operative design facilities.

Use this SINGLE CONVENIENT SOURCE for all magnetic shielding requirements. Saves you countless design hours . . . helps speed your project . . . lowers your costs.

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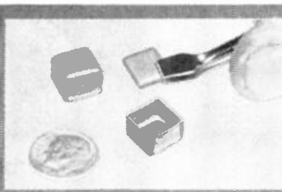
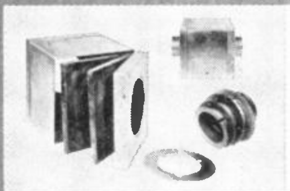


Micro miniature shield and cover, punch press operation.

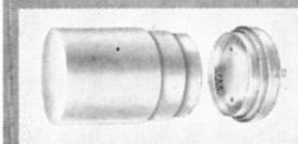
Sequence of shield cans, punch press or spinning.

Complex configuration multi-lamina shield, hydroformed.

CRT shield illustrating combination of hand fabrication, spinning and sizing.

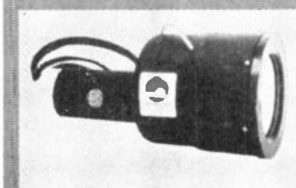


Backward wave tube shield assembly design, involving hand fabrication and hydroform or spinning.

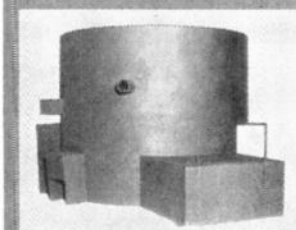


Special purpose shield, hand fabrication (levitated gyro).

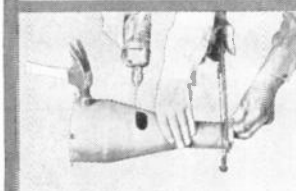
Tape preserver can be spun, hydroformed or punch press fabricated.



Data storage tube shield, hydroform or spinning, plus hand fabrication.



Large fabricated special structure (shaker table shield), approx. 60" dia. and 57" high.

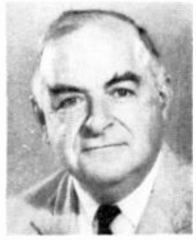


Composite photo demonstrating that magnetic shielding qualities of NETIC alloy material are not affected by vibration, shock, (including dropping), etc.

IRE People

(*Continued from page 138A*)

Frederick R. Lack (A'20–F'37) has been named to the newly created post of Senior Vice President in charge of research of the Sprague Electric Company, it was announced by R. C. Sprague, chairman of the board and chief executive officer.



F. R. Lack

He has been a Sprague director since 1959, soon after his retirement as Vice President and Director of the Western Electric Company, manufacturing arm of the American Telephone and Telegraph Company. He had been associated with the Bell System since 1911.

In June, 1959, he received the Medal of Honor of the Electronic Industries Association for his many contributions to progress in the electronics industry. He has also served as president of the American Standards Association and was director of the Army-Navy Electronics Production Agency for a period during World War II.

Mr. Lack has served two terms as a Director of the IRE. He is also a member of the American Institute of Electrical Engineers, the American Association for the Advancement of Science, the American Physical Society, and the Harvard Engineering Society.

❖

The appointment of **Ralph S. LaMontagne** (M'47–SM'56) as Marketing Manager of the Waltham Laboratories of Sylvania Electronic Systems, a division of Sylvania Electric Products, Inc., has been announced.

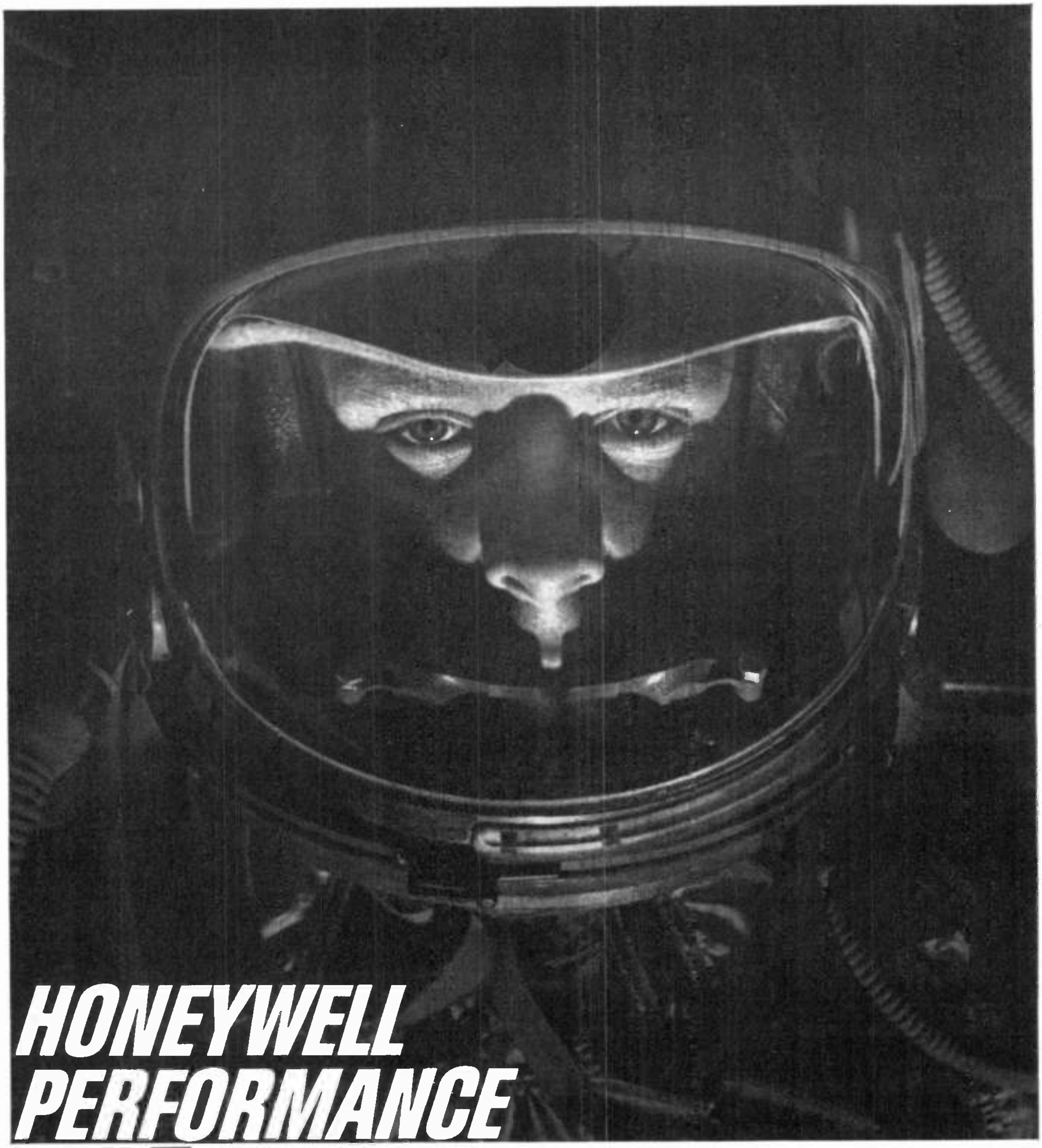
R. S. LaMontagne

Prior to joining Sylvania, he was manager of marketing of the Missile Electronic and Control Division, Radio Corporation of America, Burlington, Mass. Earlier, he served in the same capacity for RCA's Airborne Systems Department, Moorestown, N. J.

The Waltham Laboratories includes four major laboratories—the advanced development laboratory, systems engineering laboratory, microwave & antenna laboratory and equipment engineering laboratory. Programs within the laboratories are in the areas of detection, tracking and defensive missile systems, electronic warfare systems, support systems, and related military electronics equipments, components and study programs.

From 1942 to 1956, Mr. LaMontagne served with the U. S. Air Force. During this period, he was chief of Aircraft Con-

(*Continued on page 142A*)



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Honeywell

H Precision Meters

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(Continued from page 140A)

trol and Warning Branch, Washington, D. C., and earlier deputy chief of staff, Communications and Electronics of the Alaskan Air Command, Anchorage, Alaska.

He is a graduate of the University of Maryland, College Park, with the bachelor's degree in military science. He has also attended Massachusetts Institute of Technology, Cambridge, and Boston University, Boston, Mass. He is a member of the board of governors of the Lexington-Concord (Mass.) chapter of the Armed Forces Communications and Electronics Association, Air Force Association, and Association of the United States Army.

❖

Rese Engineering, Inc., of Philadelphia, Pa., electronic manufacturers supplying digital and magnetics systems and equipments to the EDP market, has recently announced the opening of their new West Coast facility. Located in Gardena, Calif., it encompasses laboratories and offices, and will serve to anchor engineering and customer service activities in West Coast markets.

D. N. Lee

Appointed as Manager of the West Coast operations is **Don N. Lee** (A'53), who will also direct product line development for the firm's new line of fast pulse generators. Recognized as an expert in the data processing and circuit design areas, he brings a solid engineering and management background to his new post.

Formerly Senior Staff Engineer with Computer Products Division of Ampex Corporation of Los Angeles, Calif., he participated in the development of the first high speed transistorized large capacity memory. His earlier experience includes activities with Telemetering Corporation of America, Sepulveda, Calif., as Senior Circuit Design Engineer; with NCR designing memory circuits for airborne computers; with Bendix Aviation's Computer Division designing one of the first memories using transistorized drivers; and, initially, with RCA participating in design of the BIZMAC computer.

A graduate of the University of Southern California, holding the B.S. degree in electrical engineering, Mr. Lee served with the U. S. Air Force as a radio and radar instructor. He is the originator of five patent applications relating to memory and telemetry developments.

❖

The appointment of **James R. Pepper** (M'59) as production engineer for Microwave Electronics Corporation, Palo Alto, Calif., has been announced.

He came to Microwave Electronics from General Electric Company at Milwaukee, Wis., where he was a project engineer in the X-Ray Department. Previously he was a production engineer with Sylvania Electric Products Microwave Tube Laboratory at Mountain View, Calif.

A specialist in the design and production of electronic tubes, he is a graduate of the University of Minnesota, Minneapolis. He has taken advanced studies in microwave radiation and transmission at UCLA and in engineering heat transfer at the University of Wisconsin, Madison.

Mr. Pepper has had military service with the Army Signal Corps and has held a reserve officer commission.

❖

T. L. Taggart, Senior Vice President of Ampex Corporation, has announced the appointment of **Meyer Leifer** (A'46–M'48–SM'50–F'55) as Chief Engineer of the Ampex Instrumentation Products Company, Redwood, Calif., manufacturers of magnetic tape recorders for business, space technology, defense, and scientific research.

M. Leifer

He joins Ampex after a fourteen year association with Sylvania, where his latest position was General Manager of Microwave Device Operations. Credited with a number of engineering innovations, Mr. Leifer is Chairman of the San Francisco Section of the IRE, and a member of Sigma Xi, the Research Society of America, Pi Mu Epsilon, and Pi Sigma Pi.

❖

Max A. Lowy (S'51–A'51–M'56–SM'57) nationally-recognized authority on telemetry systems, has been named Manager of Systems Integration for Gulton Industries, Inc.

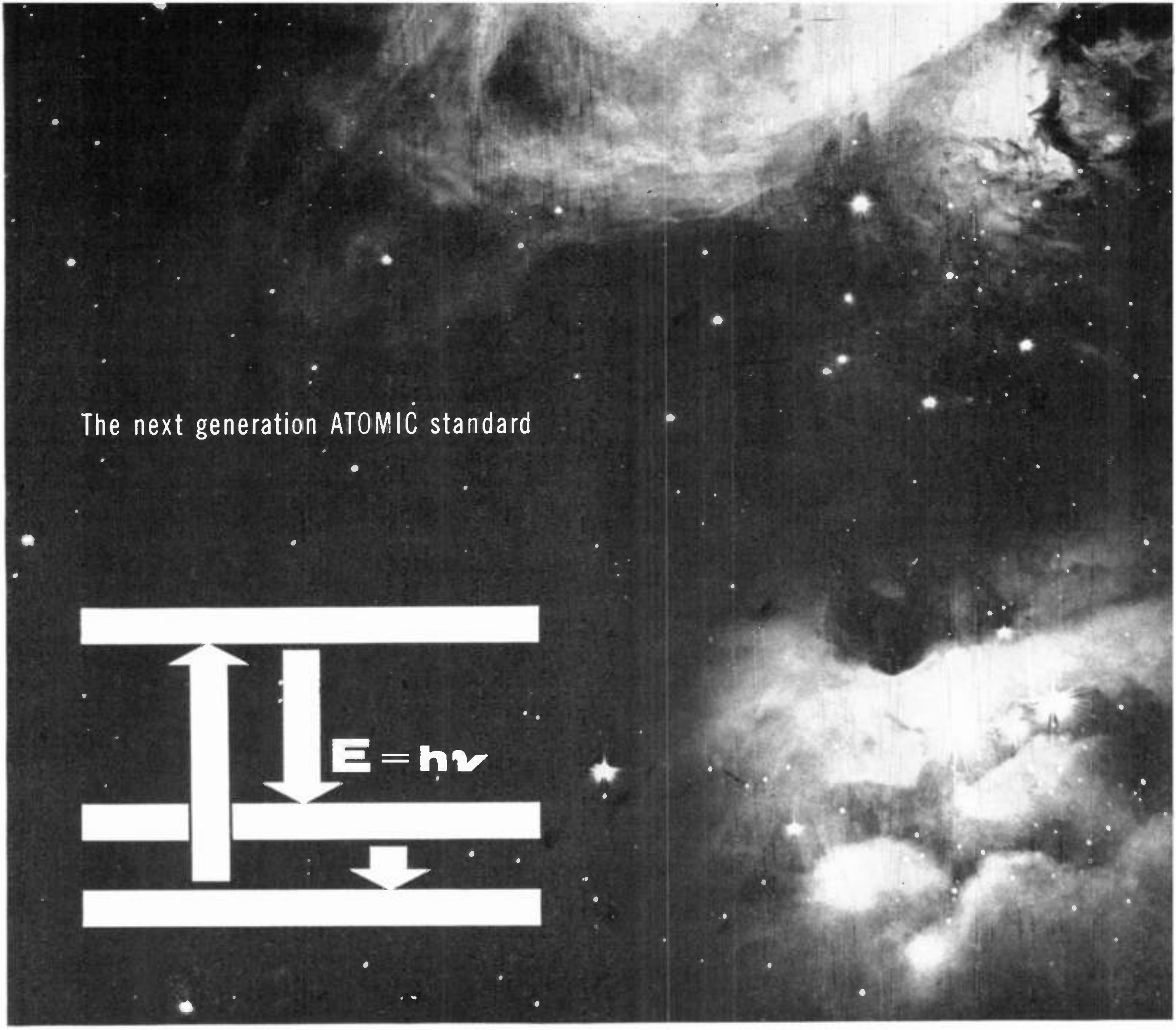
M. A. Lowy

He has broad experience with major missile programs in the areas of instrumentation and telemetry and he will be responsible for coordinating and integrating the work of Gulton's eleven domestic divisions in the engineering of complete systems for the Corporation. He will report to the Vice President of the company.

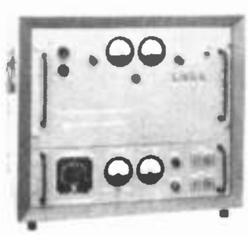
The new group under his direction will produce integrated engineered systems on the basis of the company's technical position in the areas of materials research, component and transducer design, power supply and conversion systems, instrumentation, data reduction, analysis and programming and telemetry systems.

He has been associated with Data Control Systems in Danbury, Conn., Space Technology Laboratories in Los Angeles, Calif., and the Jet Propulsion Laboratory

(Continued on page 146A)



VARIAN RUBIDIUM FREQUENCY STANDARD



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Time . . . accurate to the tenth order of magnitude . . . is now available from a newly practical and reliable source – paving the way for new advances in navigation, tracking, and communications systems. Also, the instrument is a suitable, precise calibration standard for makers of frequency systems and devices.

Working on the principles of optical pumping and transmission monitoring, the rubidium standard is recommended for continuous year-after-year operation. The heart of the system consists of two ultra-reliable elements: a long-life rubidium lamp and an all-glass rubidium vapor absorption cell. Design emphasis is on dependability throughout.

Each absorption cell is manufactured to a customer-selected time scale, i.e. Ephemeris Time (A.1) or the current standard frequency broadcast offset of 150 parts in 10^{10} relative to A.1. Cells at more than one frequency can be supplied with each instrument. Fine tuning affords time scale flexibility and extremely precise time synchronization for navigation or communications systems.

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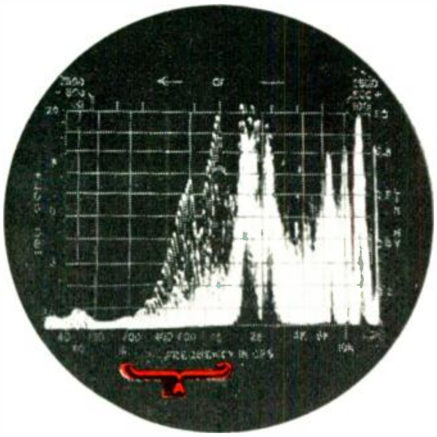
20 cps-22.5 kc.



"Coast-down" analysis of Gyro Motor by Model LP-1a Spectrum Analyzer. Area A shows decreasing fundamental frequency, resonant rise and decay, and vibration components over 60 successive scans in one minute.

The Model LP-1a "quick-look" helps locate and evaluate discrete or random signals *faster* and *easier* by scanning the entire spectrum logarithmically from 40 cps to 20 Kc. Once every second it automatically separates, measures and plots the frequency and voltage of waveform components on the calibrated X and Y axes, respectively, of a long persistence 5" CRT.

For very detailed analysis, linear segments 40 to 5000 cps wide, centerable between 0 and 20 kc, may be magnified on the screen. Amplitude ratios of up to 40 db can be simultaneously measured.

High sampling rate and panoramic displays assure

1. Minimum risk of missing weak signals or spectrum holes.
2. Fast measurements by eliminating slow point by point plots.
3. Simultaneous measurement of signals with widely divergent amplitudes and/or frequencies.
4. Continuous analysis of rapid changes in spectral content or design parameters.

Proved in hundreds of research, design and production installations, the LP-1a is a valuable tool for Noise and Vibration analysis. Harmonic and IM measurements. • General waveform studies. • Power Spectral Density analysis. • Response Curve Tracing.

SUMMARY OF SPECIFICATIONS:

Frequency Range: 20 cps—22.5 Kc.
- (1) Preset linear frequency scans: any segment width of 200, 1000, 5000 cps centerable from 0-20 Kc: Variable from 40 cps to 5000 cps with Auxiliary Function Unit C.
- (2) Preset Log Scan—40 cps to 20 Kc

Frequency Scales: Linear and Log

Center Frequency Control: Calibrated 0-20 Kc (used on lin scan)

Dynamic Range: 60 db

Amplitude Scales: Linear and 2 decade log (Expandable to 60 db)

Sensitivity: 500 μv to 500 v for full scale inear deflection

Voltage Accuracy: Lin Sweep (40 cps—22.5 Kc): ±5% or ±0.5 db. Log Sweep (40 cps—20 Kc): ±10% on lin ampl. scale, ±1.5 db on log ampl. scale.

Scan Rate: 1/sec., internally generated; adjustable with accessory equipments

Resolution: For log scan, automatically optimized. For lin scan, preset 30, 75 and 170 cps at 200, 1000 and 5000 cps sweepwidth, respectively. Variable from 10 cps to 1 Kc with Auxiliary Function Unit C.

HIGHLIGHT FEATURES:

- 1-sec. "quick-look" at entire spectrum (40 cps—20 Kc)
- Magnified analysis on reduced sweep widths
- Direct reading frequency—selective voltmeter
- Exceptionally stable circuitry; better than 5 cps/hr.
- Economical
- Simple Operation

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(Continued from page 142A)

at the California Institute of Technology, Pasadena. He also has considerable experience in industrial systems work.

Mr. Lowy is a member of the executive committee of the National Telemetering Conference and chairman of its FM/FM Standards Subcommittee; a member of the Institute of Aeronautical Sciences' advisory group on instrumentation; a member and past chairman of the American Rocket Society's Communications Committee, and presently a member of its Publications Committee. He holds the B.E.E. degree from Brooklyn Polytechnic Institute, Brooklyn, N. Y.

❖

Dr. Leonard S. Sheingold (S'47–A'51–M'55–SM'55), Director of the Applied Research Laboratory of Sylvania Electric Products Inc., has been named Chief Scientist of the United States Air Force by General T. D. White, Air Force Chief of Staff.

L. S. Sheingold

As Chief Scientist, he will have responsibility for providing technical advice on Air Force plans, programs and requirements. The appointment of a Chief Scientist is made for a one-year period.

He has served in his current position as Director of Sylvania's Applied Research Laboratory since its establishment in 1956. The laboratory is the central research facility for Sylvania Electronic Systems, a major division of the company with over-all responsibility for systems management of GT&E's major government projects.

He has held numerous active and advisory appointments to governmental scientific groups since affiliation with Sylvania in 1952. He has participated in studies involving air defense, penetration tactics, battlefield surveillance, and command and control systems.

In September, 1958, he was appointed a member of the U. S. Air Force Scientific Advisory Board by General White. He was re-appointed to that position in January, 1960, for a period of three years. He is also a member of the Advisory Group on Electronic Warfare in the Office of the Director of Defense Research and Engineering.

Under his direction, the Applied Research Laboratory, located in Waltham, has been engaged in a variety of basic and applied research programs relating to electronic systems. These programs are in the areas of communications, information processing, radio physics, mathematics, engineering, operations research, and advanced system concepts.

At the request of former Defense Secretary C. E. Wilson, Dr. Sheingold served as a member of the Defense Department's Weapons Systems Evaluation Group during 1956, and until recently was a consultant to that group. In March, 1957, he served as chairman of the Ground Environment Committee during a joint U. S.-Canada meeting on penetration tactics at a Weapons Systems Evaluation Group symposium in Washington, D. C.

During the summer of 1958, he was a consultant to the University of Michigan, Ann Arbor, and assisted in the preparation of a long-term combat surveillance research program. He has been a consultant in air defense to the Operations Research Office of the Johns Hopkins University, Baltimore, Md.

He holds the Master of Arts and Doctor of Philosophy degrees in applied physics from Harvard University, Cambridge, Mass. Prior to joining Sylvania, he was an instructor at Syracuse University, Syracuse, N. Y., where he received the Bachelor of Science and Master of Science degrees in electrical engineering.

A native of Boston, he is a member of Sigma Xi and Sigma Pi Sigma honorary societies, the American Physical Society, and the Anti-Submarine Warfare Committee of the National Security Industrial Association. He has authored numerous papers in the areas of electromagnetic theory and microwave techniques.

Dr. Sheingold served with the United States Army during World War II.

❖

William F. Utlaut (SM'55), a specialist in the use of radar techniques to study the ionosphere and upper atmosphere, has been appointed Chief of the National Bureau of Standards' Section on High Frequency and Very High Frequency Research according to an announcement by Dr. F. W. Brown, Director of the NBS Boulder Laboratories. This section is a part of the Radio Systems Division within the NBS Central Radio Propagation Laboratory.

W. F. Utlaut

Since joining NBS in 1952 Mr. Utlaut's studies have been concerned with increasing the understanding of the ionosphere and its effect upon radio waves. His efforts have ranged from the design of special electronic equipment for radio research to theoretical studies of ionospheric propagation.

The wide scope of these studies is indicated by the programs currently being conducted under his direction. These include research on radio systems for long distance communication circuits and for the detection of missiles and nuclear detonations, research on the scattering of VHF signals by one of the lower layers (D region) of the ionosphere, and a study of whether or not lines of force in the earth's magnetic field could be used to bounce signals between the northern and southern hemispheres of the globe.

The third program, being conducted in cooperation with R. M. Gallet of NBS, ac-

(Continued on page 148A)

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.

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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

IRE People

(*Continued from page 146A*)

tually reaches far above the ionosphere into the region known as the exosphere. The first experiment in this program used radar techniques to bounce a signal from Washington, D. C., to a point near the southern tip of South America and the path taken by the signal curved above the earth some six thousand miles into space.

In addition to directing the activities of this section, Mr. Utlaut is the NBS scientific officer for a research program on meteor astronomy being conducted by Harvard University. He is a consultant to the Army Signal Corp and in this position recommends methods of selecting the best locations to use for military communication links overseas. He is also a consultant to the Department of Defense Advanced Research Projects Agency in their program of electromagnetic wave studies for nuclear detection and a consultant to the Defense Atomic Support Agency.

He is Chairman of the local chapter of the IRE Professional Group on Antennas and Propagation and is a senior member of the Research Society of America. He is the co-author of a recent scientific paper on the exospheric studies described above and, also, is author of a paper which appears in the current issue of the *NBS Journal of Radio Propagation*, on the "Effect of Antenna Radiation Angles Upon HF Radio Signals Propagated Over Long Distances." He received both the B.S. and M.S. degrees in electrical engineering from the University of Colorado, Boulder, in 1944 and 1950, respectively, and is currently teaching a course in Electromagnetic Waves for the Extension Division of Colorado University. He also left NBS for one year to head a research project for the University of Colorado on the separation of radioactive isotopes by electronic methods.

❖

William B. Lurie (M'46–SM'48), former senior project engineer and program director of General Precision Laboratories, Inc., has joined Burnell & Co., Inc., of Pelham, N. Y., an independent manufacturer of electronic filters and delay lines, as chief engineer.

W. B. Lurie

He will be in charge of the company's design and engineering administration, and he will also coordinate engineering and research of the company's subsidiaries and its Guillemin Research Laboratory.

At GPL's Pleasantville, N. Y., facility, he was in charge of all phases of the engineering, product design and administration of a $4 million-a-year government program. Earlier, working on Doppler radar navigational equipment for self-contained airborne navigators, he supervised circuit

(*Continued on page 150A*)

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Double Frame Grid wide band pentode

Type 7788... **designed for use as a broadband amplifier in radio and TV relay systems, coaxial telephone lines, radar equipment and oscilloscopes**

OUTSTANDING FEATURES:

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- Both control and screen grids of frame grid construction (an Amperex first!) for reduced microphonics; minimum spread in characteristics; and lower screen grid current which contributes to short and long term stability
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IRE People

(Continued from page 118A)

design and analysis in the low and medium frequency areas.

A 1939 graduate in mathematics of Yale University, New Haven, Conn., he also holds the master's degree from Columbia University, New York, N. Y. During World War II, he was a physicist with the Navy, first in the Bureau of Ships and then in the Naval Ordnance Laboratory. Later he was a physicist at the Picatinny Arsenal, Dover, N. J. Before joining GPL in 1951, he was a design engineer with Machlett Laboratories, Inc., in Springdale, Conn.

He is a member of the American Physical Society, has written numerous papers and articles in the electronics field, and holds several patents.

❖

The promotion of **Bogdan R. Stack** (A'50–M'53–SM'59) to Associate Laboratory Director at ITT Federal Laboratories, Palo Alto, Calif., has been announced. He will head the Communication Laboratory at the West Coast location.

Formerly Executive Engineer, he joined the Laboratories in 1957 from the Stromberg-Carlson Company, Rochester, N. Y., where he was responsible for development of commercial subscriber telephone equipment. At Palo Alto he held responsibility for the administration and supervision of communication projects including circuit design for various communication exchanges.

He has also held engineering positions with Lenkurt Electric Company and Radio Engineering Products, Canada, and was associated with design and development work involving telegraph, cable and radio equipment.

Mr. Stack is a member of the American Institute of Electrical Engineers and the Institute of Electrical Engineers, London. He received the Bachelor and Master degrees in electrical engineering from the University of Bristol, England, and McGill University, Canada, respectively. During World War II he served in the Polish Army under British command.

❖

Eitel-McCullough, Inc., announces the election of **Richard T. Orth** (A'31–M'38–SM'43) as a Director of the company. He is Vice President, Operations, of the San Carlos electronics firm.

R. T. Orth

He was graduated from Purdue University, Lafayette, Ind., in 1930 with the Bachelor of Science degree in electrical engineering. His past employment includes a number of years with the Radio Corporation of America.

❖

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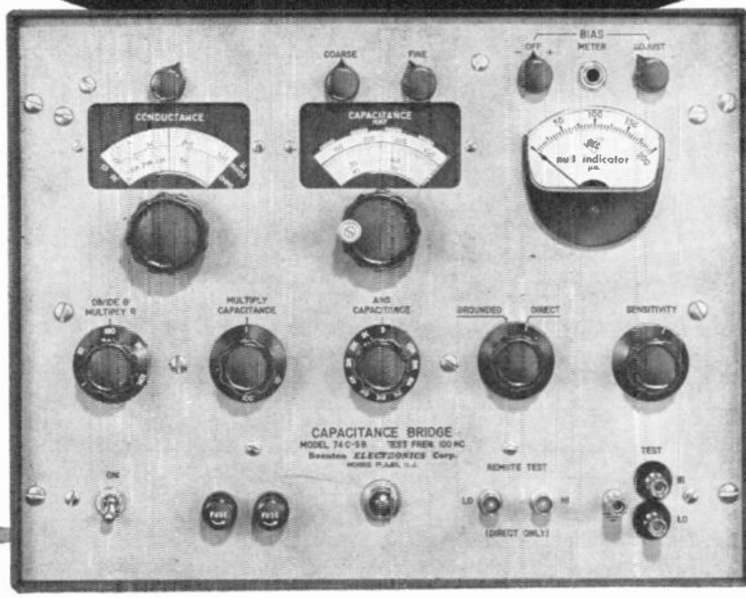

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- 0.0002-11,000 μμf Generally 0.25%
- 1000 ohms to 1000 megohms Shunt Resistance
- 0.001 to 1000 μmhos Conductance

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MODEL 74C-S8 (Shown)
- With −5 to +100V DC Bias for Diode Testing

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MODEL 75A-S8
- With −5 to +100 VDC Bias for Diode Testing

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A revolutionary equipment development patented by Philco. Provides unequalled control over all production phases. Results in an inherent reliability possible only through automated production. To you concerned with reliability this means a *constant* output of reliable transistors.

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A widely acclaimed 5 million dollar transistor reliability assurance program. Surveys quality and reliability from initial design to final release of the finished product. A key factor in this project is the most advanced test equipment found anywhere . . . eliminates all test variables induced by operator techniques or errors. To you concerned with reliability this means the constant checking and eliminating of the factors that result in unreliable transistors.

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Currently, more than two million test hours are being run weekly . . . new equipment being installed will run the logged hours of life tests to 2½ billion by 1963! Provides a history of transistor operation under maximum stress conditions. This is important in determining the effectiveness of present production techniques and in developing new ones. To you concerned with reliability this means that you not only receive the optimum in reliability now, but are assured of it in the future.

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EVERY 4TH EMPLOYEE'S SOLE CONCERN IS RELIABILITY!

Right! 25% of our working force is employed full time at improving our product. But it doesn't stop there. All personnel including operators, engineers, supervisors and managers attend comprehensive training programs designed to educate them with the need for reliability and the steps necessary to achieve it. To you concerned with reliability this means that we are highly concerned with it too!

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THE EXTRA INSURANCE OF 100% TESTING OF ALL PRODUCED UNITS!

Tests and sorts at the incredible speed of 3600 transistors per hour. To you concerned with reliability this is *extra insurance* that you receive unmatched quality and reliability.

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SQUARE WAVES
and PULSES

A Wide Range of Outputs From Two Compact Signal Sources

Type 1210-C Unit R-C Oscillator,* $180

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- Wide Frequency Range: 20 cps to 500 kc in 5 ranges, either sine or square waves. Calibration accuracy: ±3%
- Three Outputs:

Sine Wave, low-impedance output
(for loads of 500 ohms and higher). Maximum open-circuit output is 7v. Output constant within ±1 db to 200 kc; distortion less than 1.5% over entire range; hum down 60 db.

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Square Waves
0 to 30v peak to peak; rise time about ⅓ μsec; overshoot approximately 1%, hum down 60 db.

***Can be converted to a sweep oscillator** with addition of G-R Synchronous Dial Drive. Type 908-P1 Drive sweeps oscillator at a rate of one frequency decade in 70 sec. Price, $32.

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Type 1217-A Unit Pulser, $250

requires Unit Power Supply, $50

- **Repetition Rate:** 30 cps, 60 cps; 100 cps to 100 kc in X1, X2, and X5 steps; with external drive (1210-C Oscillator or equivalent), continuous from 15 cps to 100 kc. (minimum external drive is 10v to 10 kc, 25v to 100 kc)
- **Pulse Duration:** 0.2 μsec to 60,000 μsec.
- **Pulse Shape:** Rise time 0.05 μsec; decay time 0.15 μsec. Pulse top is flat to within 5% of maximum value.
- **Amplitude:** Adjustable from 0 to 20v open circuit for both positive and negative pulses, 50v negative pulse obtainable when positive terminal is grounded.
- **Jitter:** No observable jitter when one full period is displayed on scope.
- **Output Impedance:** 200 ohms for positive pulses; 1500 ohms for negative pulses.

Unit Pulser and Power Supply can be easily rack mounted with the Type 480-P4U3 Adaptor Panel ($12.00)
Same Adaptor-Panel Size accepts Unit R-C Oscillator, and Power Supply

GENERAL RADIO COMPANY

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