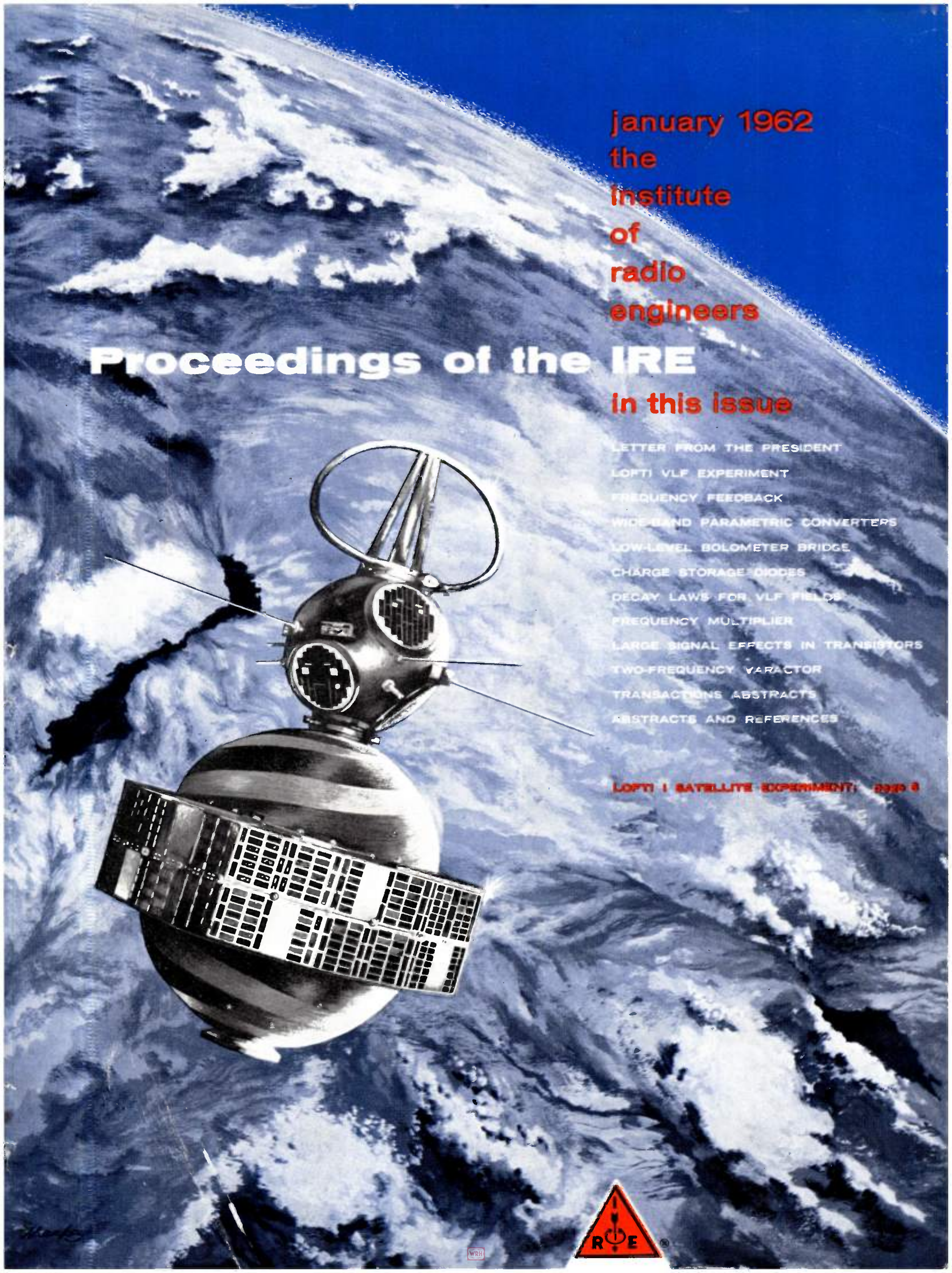

january 1962
the
institute
of
radio
engineers

Proceedings of the IRE

in this issue

LETTER FROM THE PRESIDENT

LOFTI VLF EXPERIMENT

FREQUENCY FEEDBACK

WIDE-BAND PARAMETRIC CONVERTERS

LOW-LEVEL BOLOMETER BRIDGE

CHARGE STORAGE DIODES

DECAY LAWS FOR VLF FIELDS

FREQUENCY MULTIPLIER

LARGE SIGNAL EFFECTS IN TRANSISTORS

TWO-FREQUENCY VARACTOR

TRANSACTIONS ABSTRACTS

ABSTRACTS AND REFERENCES

LOFTI I SATELLITE EXPERIMENT: page 8

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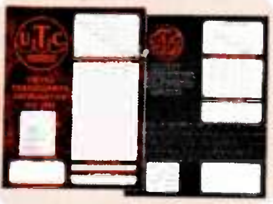

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VOLUME 50, NUMBER 1

January, 1962

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

contents

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CORRESPONDENCE

COVER

A LOFTI satellite mounted atop a TRANSIT satellite was sent aloft last year to investigate VLF propagation phenomena in the ionosphere. The initial results of this experiment are reported on page 6 of this issue.

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Proceedings of the IRE®

continued

REVIEWS

Books:

ABSTRACTS

IRE NEWS AND NOTES

Programs:

DEPARTMENTS

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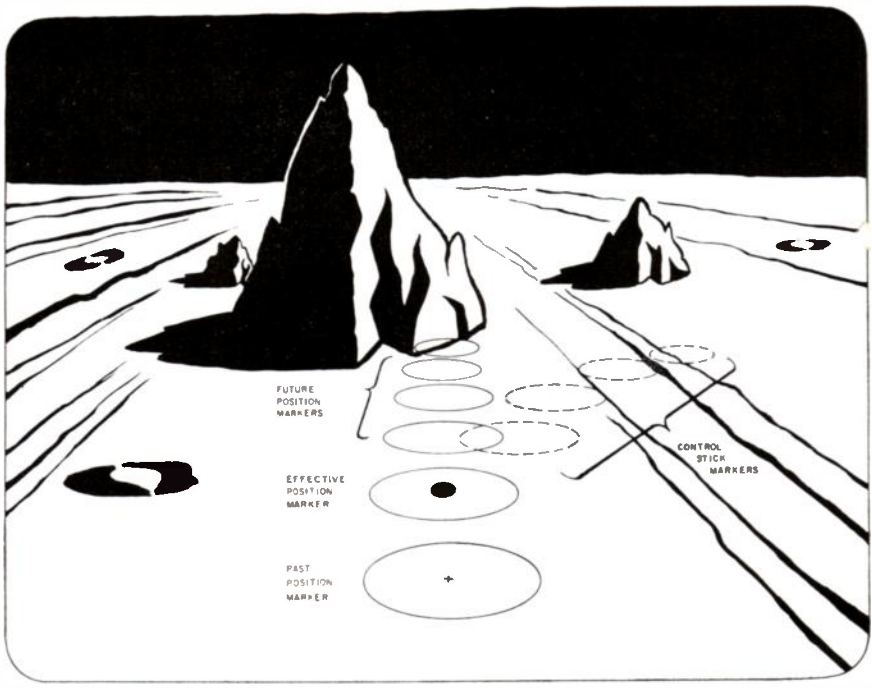
A robot lunar vehicle is probably the simplest and cheapest tool for exploring the lunar surface. Since the first vehicles will probably be guided by commands from an earth-based control center, AIL has investigated the effects of transmission and other delays on the rapidity and precision of vehicle guidance. This month David Cohen of our Advanced Projects Department describes a position predictor system he has developed to minimize the effects of these delays and permit continuous vehicle guidance.

Prediction Display Computer for Remote Lunar Vehicle Guidance



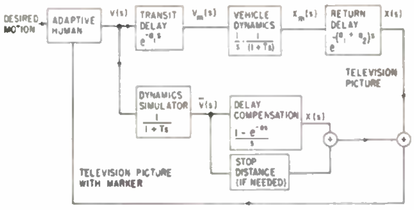
Predicted display

Since the average earth-moon distance is only about 250,000 miles, a lunar robot vehicle exploring the lunar surface can be guided by real-time command signals from an earth-based control center. The guidance signals would be based upon a TV display of the lunar terrain at the earth-based control center as seen by the TV camera mounted on the vehicle. However, the total delay of 3.6 seconds (2.6 seconds for round-trip transmission, 1 second for slow-scan TV) between initiation of a command and perception of vehicle response on the TV display would significantly hinder rapid and accurate vehicle guidance.

To understand the guidance problem and its solution, some time delay concepts must be clarified. The *apparent* vehicle position is the vehicle position as indicated by the TV display at the earth-based control console. This TV display, transmitted from the vehicle-borne TV camera, is the view from the vehicle position 2.3 seconds in the past because of the moon-earth transmission time and the slow-scan TV delay. The *actual* vehicle position is that position which the vehicle does in fact occupy at any given instant of time. The *effective* vehicle position is the vehicle position on the lunar terrain 1.3 seconds (earth-moon transmission time) in the future from its actual position when a guidance command is issued at the control console.

Since the delays are additive, the guidance operator at the control console does not see the results of his guidance commands for 3.6 seconds. Tests using the AIL lunar robot vehicle system simulator demonstrated that the total time delay and consequent discrepancy between the apparent vehicle position as indicated by the TV display and its effective position as determined by earth-moon transmission time results in slow and inaccurate vehicle guidance.

To remedy this, a system has been devised for computing and presenting an effective vehicle position marker on the console TV display of the lunar terrain. The marker responds immediately to the guidance commands issued by the console operator who steers the marker across the displayed terrain as if it were the vehicle. Thus, the marker position represents at all times the position at which the commands issued at the console will be effective.



Transform of control loop for unidirectional motion

The marker position on the console TV display is computed by integrating the results of vehicle guidance commands over the total delay time caused by round-trip transmission time and slow-scan TV. Vehicle dynamic characteristics and the nonlinear terrain scale as seen on the console TV display are included in the computation. The electrical signals representing the computed effective vehicle position on the TV display are used in the console TV display circuitry to generate and position variable sized elliptical effective position markers. An elipse rather than a dot or circle is used to give the marker the appearance of perspective varying with range as seen on the TV display.

A complete bound set of our fifth series of articles is available on request. Write to Harold Hechtman at AIL for your set.

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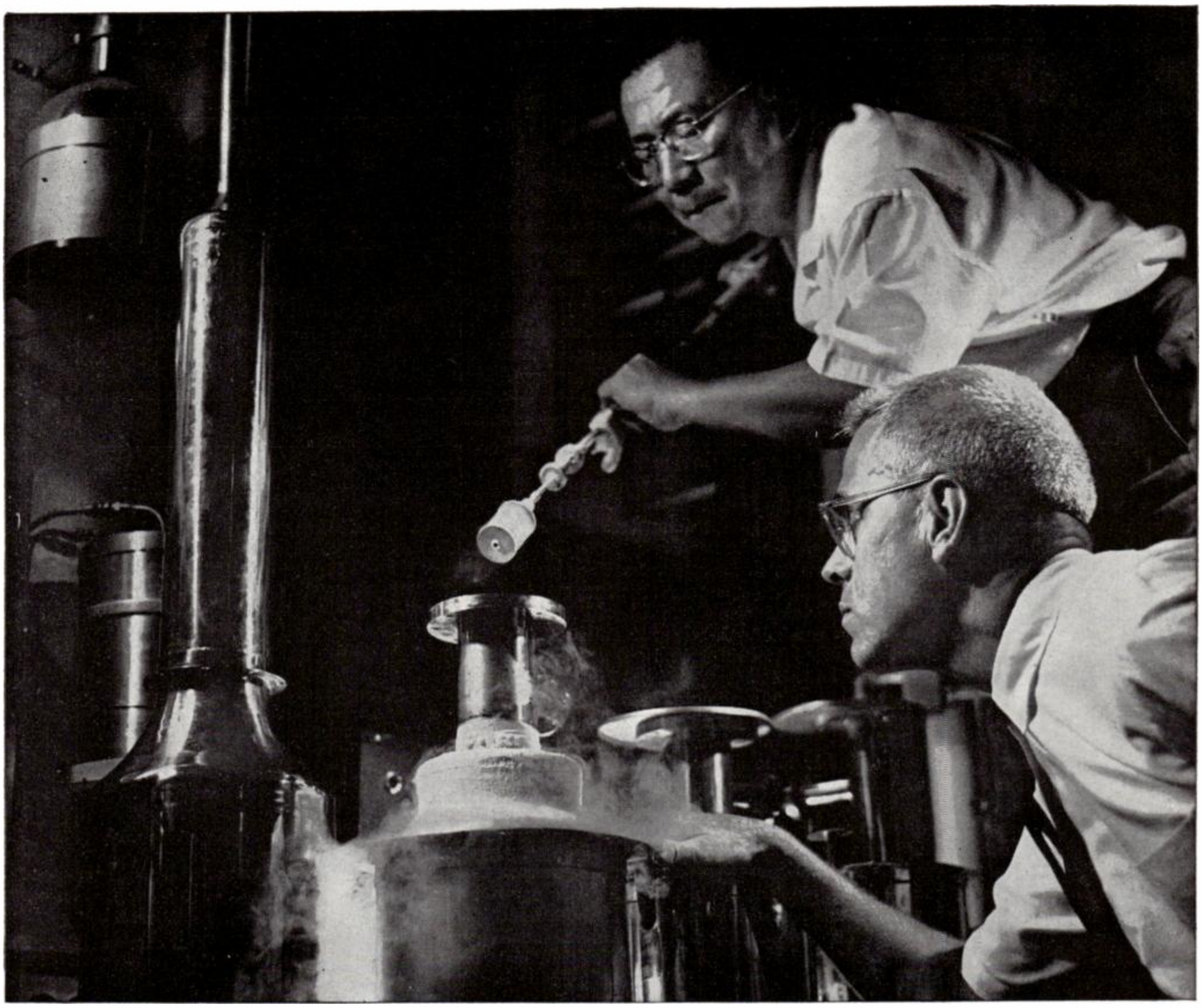

The Making of a Magnet. Bell scientists test new superconducting electromagnet, the small cylindrical object being removed from helium bath at minus 450 degrees F. An early experimental design produced a field strength over 65,000 gauss.

OUT OF SOLID STATE SCIENCE COMES A POWERFUL NEW MAGNET

Bell Telephone Laboratories' creation of a powerful superconducting electromagnet once again illustrates the role of materials research in the advancement of communications.

It has long been known that certain materials called superconductors have a zero electrical resistance at temperatures near absolute zero. A solenoid of superconductive wire carrying a large current should be capable of producing an extremely powerful magnetic field without the bulky power equipment that is needed for conventional electromagnets.

A formidable obstacle blocked the way, however. The strong magnetic field tended to destroy the wire's superconductivity.

Bell Laboratories scientists studying superconductors—as part of their endless search for new materials for communications—were led to the discovery of a number of alloys and compounds having exceptional superconductive properties. One of these materials, a compound of niobium and tin, was found to possess a startling ability to retain its superconductivity in intense magnetic fields of over 100,000 gauss. Bell scientists went on to show how the brittle, intractable material could be made into a wire and hence wound to make an extremely powerful electromagnet.

By finding a low-cost way to create enormously powerful magnetic fields, Bell scientists have brought closer new applications of magnetism in communications. Intense magnetic fields provide an invaluable tool in research, and offer an attractive means for containing hot plasma in thermonuclear experiments.

The new magnet is another example of how Bell Laboratories research not only works to improve Bell System communications but also benefits science on a broad front.

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TL-8D16A	8 kc
TL-10D18A	10 kc
TL-16D25A	16 kc
TL-20D32A	20 kc
TL-30D45A	30 kc
TL-40D55A	40 kc
TL-45D60A	45 kc
TL-32E48C	32 kc
TL-50D85C	50 kc

(suffix "A" denotes 455 kc center frequency;
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Clevite ceramic ladder filters provide more selectivity for their size than any conventional i-f filter. They are fixed tuned and need no alignment—are non-magnetic and non-microphonic. Leading manufacturers now have Clevite ladder filters in their communications equipment. Improve your newest design with these unique filters. Write now for complete specifications—Bulletin 94012, or for selectivity curves available on each stock model. ■ Dimensions: 5/16″ diameter x 1½″ long. ■ Selectivity: 60 db/6db shape factor from 1.3:1 to 2.6:1. ■ Center Frequency Stability: within 0.2% for 5 years, and within 0.2% from −40° to +85° C. ■ Impedance: 1200-1500 ohms. ■ Designed for military environment.

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Meetings with Exhibits

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

Δ

February 7-9, 1962

3rd Winter Convention on Military Electronics, Ambassador Hotel, Los Angeles, Calif.

Exhibits: Mr. Walter E. Peterson, c/o IRE Los Angeles Office, 1435 South La Cienega Blvd., Los Angeles, Calif.

March 1-3, 1962

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

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

March 26-29, 1962

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

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

April 11-13, 1962

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

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

May 1-3, 1962

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

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

May 14-16, 1962

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

Exhibits: IRE Dayton Office, 33 N. Torrence St., Dayton, Ohio

May 23-25, 1962

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

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

June 12-14, 1962

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

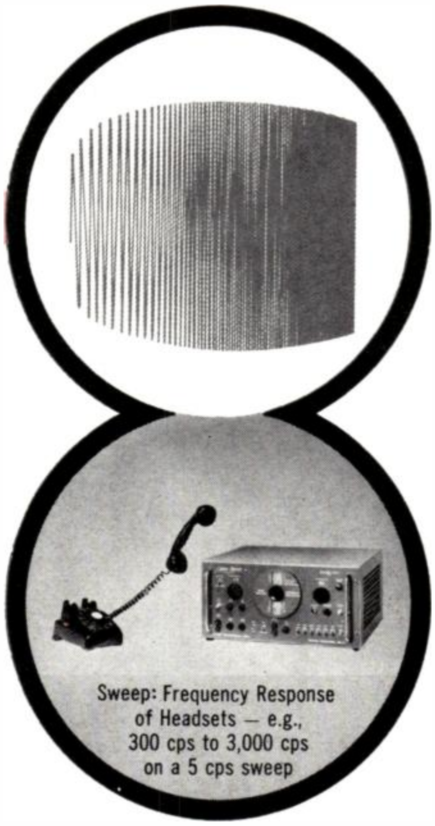
Exhibits: Mr. William C. Copp, 72 W. 45th St., New York 36, N.Y.

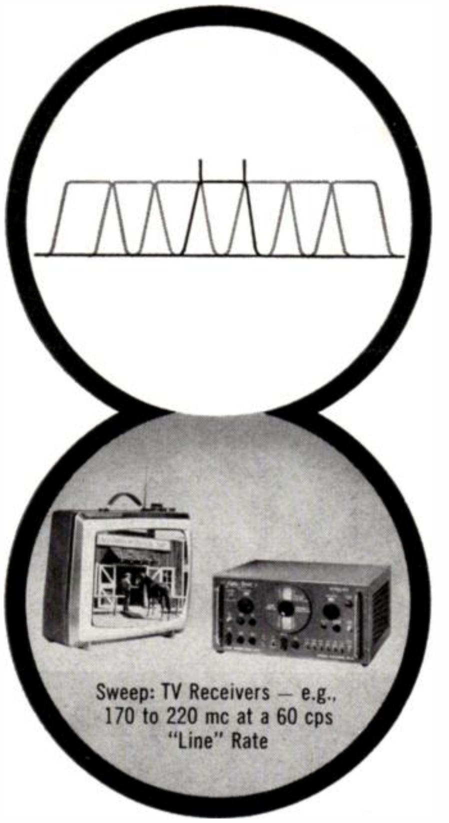
June 25-27, 1962

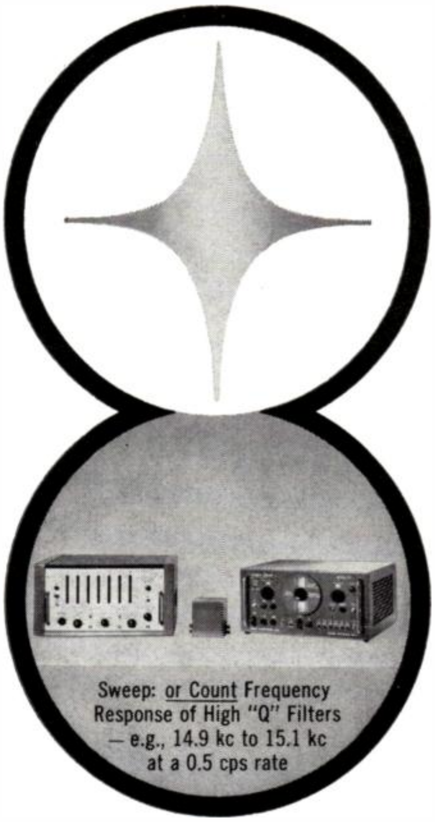
Sixth National Convention on Military Electronics, Shoreham Hotel, Washington, D.C.

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

(Continued on page 10A)


Sweep: Frequency Response of Headsets — e.g., 300 cps to 3,000 cps on a 5 cps sweep


Sweep: TV Receivers — e.g., 170 to 220 mc at a 60 cps "Line" Rate


Sweep: or Count Frequency Response of High "Q" Filters — e.g., 14.9 kc to 15.1 kc at a 0.5 cps rate

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50 CPS TO 220 MC IN 12 BANDS • WIDE RANGE OF SWEEP WIDTHS
VARIABLE REP RATES • MANUAL AND AUTOMATIC OPERATION

- Single wide-sweep video displays from 10 mc down to 1 kc.
- Linear and logarithmic sweeps of 0.2 cps to 30 cps; or sweep locked to line frequency.
- Audio Sweep of 50 cps to 20,000 cps.
- 8 fixed, narrow-band video frequency sweeps for repetitive operations.
- Fundamental frequency 10 mc to 220 mc (widths to 30 mc plus).
- High-level output of 1 V rms into 70 ohms. AGC'd to ± 0.5 db over widest sweep.
- Manually-operated control for varying oscillator frequency.
- Fixed pulse-type markers or variable marker provision.

Price: $1295.00 F.O.B., Factory ($1425.00 F.A.S., New York).

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For high frequency work, the unit provides 9 sweep bands, operating at fundamental frequencies for wide, stable sweeps from 10 to 220 mc. At the low end of the spectrum, an audio frequency sweep from 50 to 20,000 cps is provided. High order stability permits frequency sweeps to as low as 50 cps.

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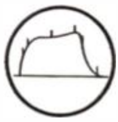
Check Video Bandpass — e.g., 1 kc to 10 mc at 1 cps rate


Check Radar IF's — e.g., 25 to 35 mc at 30 cps


Check Audio Bandpass — e.g., 50 cps to 20 kc at 1 cps Log Sweep

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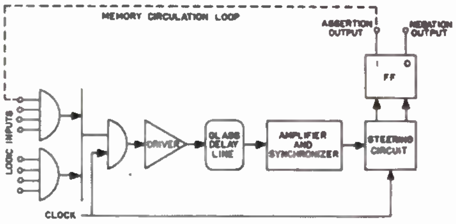

Meetings with Exhibits

(Continued from page 8A)

August 21-24, 1962

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

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

August 29-September 1, 1962

Second International Conference on Information Processing, Munich, Germany

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

October 1-3, 1962

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

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

October 8-10, 1962

National Electronics Conference, Exposition Hall, Chicago, Ill.

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

October 29-31, 1962

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

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

November 13-15, 1962

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

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

December 4-5, 1962

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

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

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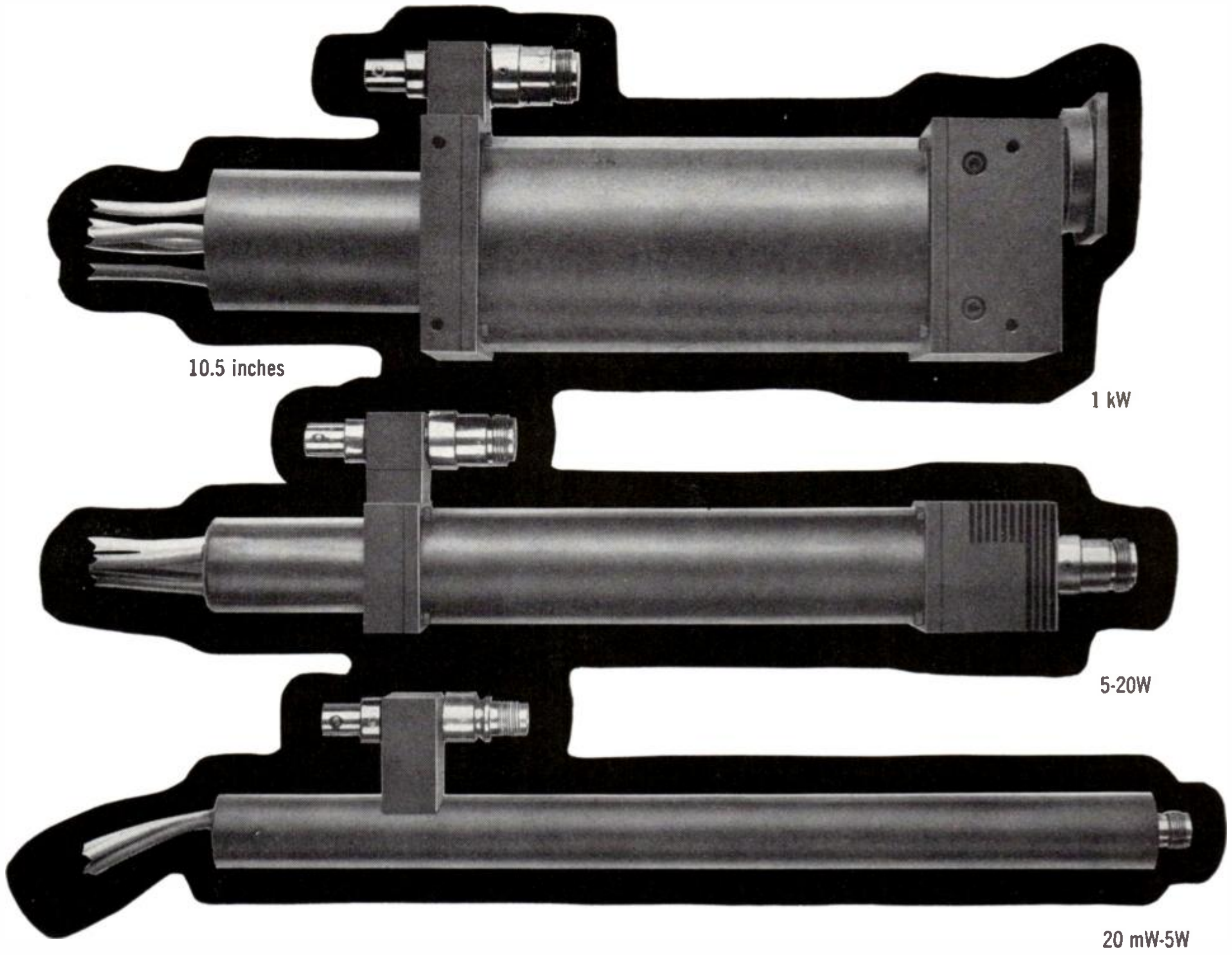
Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings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



NEW LITTON TRAVELING WAVE TUBES

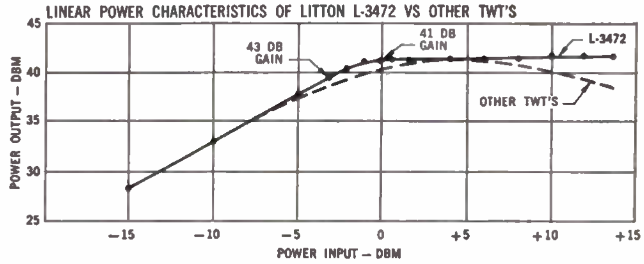
20 mW to 5.5 kW, 400 Mc to 11,000 Mc

Want high performance, high CW power and wide bandwidth from a compact, lightweight TWT? You get just that with the Litton Industries underweight, undersized family of TWT's that are bigger and better in every other way. For instance, the L-3472 gives you 10 watts in a 12-inch, 2.5-pound package. The 2-watt L-3612 is only 11 inches long and weighs just 1.5 pounds.

Most of these tubes are PPM focused. Special helix design and fabrication techniques have raised the saturation gain to within three db of small signal gain. Can be supplied with small signal gains up to 60 db. Input and output circuits are coaxial, providing minimum frequency sensitivity. Bandwidth limited only by combined beam-circuit bandwidth. Optimum voltage for small signal operation same as for saturated operation. Helix isolated from tube body, permitting overload protection, helix modulation, and helix current monitoring.

Guaranteed to withstand Class II military environment. Applications include airborne ECM, power tube driver, radar target repeaters, and voice and data communications systems.

Contact us at San Carlos, California, for more information.

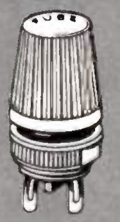


Tube Type Number	Frequency Range Megacycles	Power Output Minimum	Small Signal Gain Minimum	Duty Factor
L-3499	2000-4000	2 W	36 db	CW
L-3663	2000-4000	10 W	33 db	CW
L-3619	2000-4000	20 W	33 db	CW
L-3470	4000-8000	20 mW	36 db	CW
L-3711	4000-8000	1 W	36 db	CW
L-3471	4000-8000	2 W	36 db	CW
L-3657	4000-8000	10 W	33 db	CW
L-3658	4000-8000	20 W	33 db	CW
L-3611	7000-11000	20 mW	36 db	CW
L-3612	7000-11000	2 W	36 db	CW
L-3528	5000-11000	5 W	33 db	CW
L-3472	7000-11000	10 W	36 db	CW
L-3529	7000-11000	20 W	36 db	CW
L-3614	8000-11000	1 kW	36 db	.02
L-3497*	1240-1400	5.5 kW	40 db	.06
L-3674*	400- 450	5 kW	37 db	.06
L-3637*	5900-8400	200 mW	30 db	CW

*In Development

LITTON INDUSTRIES
Electron Tube Division
MICROWAVE TUBES AND DISPLAY DEVICES

GMW TRON FUSE
Diameter: .270 inch
Length of Body: ¼ inch



HWA FUSEHOLDER
Diameter: .500 inch
Length with Knob and Terminals: 15/16 inch

ACTUAL SIZES SHOWN

Another BUSS sub-miniature fuse and holder combination

EXTREME RELIABILITY UNDER HIGH SHOCK AND SEVERE ENVIRONMENTAL CONDITIONS.

Rigid construction of fuse and holder assures extraordinary reliability under high shock and vibration conditions. Fully insulated ceramic body isolates fusible element from effect of dust, corrosion, moisture and vapors.

DESIGNED FOR SPACE-TIGHT APPLICATIONS

Panel Mounted. Holder can be mounted on panel by hand. No special tool required to run down holding nut.

Prong type contacts on fuse make it easy to install or replace.

A knob for the holder may be used to make holder water proof from front of panel.

HOLDER CAN BE MOUNTED IN PRINTED CIRCUITS

Terminals of holder can be inserted into holes and soldered on printed circuit board without additional forming.

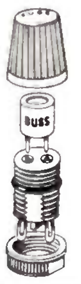
If desired, GMW fuse may be used without holder and mounted directly into printed circuit boards.

AVAILABLE RATINGS FOR GMW FUSES.

Fuses are made in sizes from 1/10 to 5 amperes for use on circuits of 125 volts or less where fault current does not exceed 50 amperes.

Transparent window in end of fuse body permits visual inspection of fusible element.

Before crystallizing your design using sub-miniature fuses be sure to get full data on the Buss GMW fuse and HWA holder combination.



Exploded view of GMW-HWA fuse & holder combination.

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

FUSETRON
TRUSTWORTHY NAMES IN ELECTRICAL PROTECTION
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IRE News and Radio Notes

Current IRE Statistics

(As of October 31, 1961)

Membership—95,572
Sections—111
Subsections—31
Professional Groups—29
Professional Group Chapters—291
Student Branches†—215

† See June, 1961, issue for a list.

Calendar of Coming Events and Author's Deadlines*

1962

Jan. 9–11: 8th Nat'l. Symp. on Reliability and Quality Control, Statler Hilton Hotel, Washington, D. C.

Jan. 28–Feb. 2: AIEE 1962 Winter General Mtg., New York, N. Y.

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

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

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

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

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

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

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

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

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

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

Apr. 24–26: Symp. on the Mathematical Theory of Automata, United Engrg. Ctr., New York, N. Y. (DL*: Jan. 5, 1962, Symp. Committee, Polytech. Inst. Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.)

May 1–3: Spring Joint Computer Conference, Fairmont Hotel, San Francisco, Calif. (DL*: Nov. 10, 1961, R. I. Tanaka, Lockheed Missiles and Space Co., Palo Alto, Calif.)

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

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

IRE ELECTS OFFICERS FOR 1962

Dr. Patrick E. Haggerty (A'47–SM'53–F'58), President of Texas Instruments Inc., Dallas, Texas, has been elected the 1962 President of the Institute of Radio Engineers. He succeeds Lloyd V. Berkner (A'26–M'34–SM'43–F'47), President of Graduate Research Center Inc., Dallas.

T. A. Hunter (A'29–M'45–SM'46–F'53), President of Hunter Manufacturing Co., Inc., Iowa City, Iowa, has been elected Vice-President representing North America for 1962. He succeeds J. F. Byrne (SM'45–F'50) Manager of the Riverside Laboratories, Motorola, Inc., Riverside, Calif. The Vice President representing overseas countries for 1962 will be A. M. Angot (SM'51), Technical Director of Telecommunications Radioelectriques et Telephoniques, Paris, France. He succeeds Franz Ollendorff (SM'52), Research Professor at the Technion-Israel Institute of Technology, Haifa, Israel.

Elected as Directors for the 1962–1963 term are Dr. G. K. Teal (SM'53–F'56), Assistant Vice President of the Research and Engineering Division of Texas Instruments Inc., and G. A. Woonton (A'40–SM'44–F'51), Professor and Director of Eaton Electronics Research Laboratory of McGill University, Montreal, Canada.

Regional Directors for the 1962–1963 term are: Region 1—Dr. F. K. Willenbrock (S'46–A'51–M'55), Associate Dean of Engineering at Harvard University, Cambridge, Mass.; Region 3—M. R. Briggs (A'37–SM'46–F'54), Communications and Control Engineering Manager of Westinghouse Electric Corp., Baltimore, Md.; Region 5—W. B Bruene (A'45–SM'49), Collins Radio Co., Cedar Rapids, Iowa; Region 7—D. K. Reynolds (A'48–SM'51), Professor of Electrical Engineering, at the University of Washington, Seattle, Wash.

SYMPOSIUM ON REDUNDANCY TECHNIQUES FOR COMPUTING SYSTEMS

A Symposium on Redundancy Techniques for Computing Systems will be held on February 6–7, 1962, co-sponsored by the Information Systems Branch, Office of Naval Research, and the Electronics Division of Westinghouse Electric Corp. This Symposium will be held in the Department of the Interior Auditorium, C Street between 18th and 19th Street, N. W., Washington, D. C.

The objective of the Symposium is to focus attention toward new ideas, research and developments, which may lead to the introduction of redundancy techniques into forthcoming computing systems. It is apparent that many information processing systems of the future will be of such size that conventional methods of fabrication and of maintenance will be completely unfeasible. Computers for space applications have reliability requirements which are presently not achievable without use of redundancy. Thus, some form of logical redundancy must be introduced if practical systems are to be operated. Accordingly, it is planned to present in a single symposium a collection of related papers concerning suggested new techniques for the use of redundancy in computing systems. This should permit relative evaluation of such techniques with respect to their probable future utilization in various aspects of the computing field. The program will consist of papers invited from many of the organizations engaged in appropriate research and development activities.

Attendance is open to all interested technical personnel. Further information and a preliminary Symposium program, when available, may be obtained by contacting: Miss. J. Leno, Code 430A, Office of Naval Research, Washington 25, D. C.

The Sixth Annual Symposium of the North Carolina Section held in Greensboro, North Carolina, October 19–20, 1961. *Left to right:* Dr. B. J. Dasher, Director Region III; P. A. McMasters, Chairman, South Carolina Section; A. B. MacIntyre, Chairman, North Carolina Section; and W. A. Bernath, General Chairman of the Symposium.

IRE ANNOUNCES 1962 AWARDS

The Institute of Radio Engineers has announced the names of the recipients of the 1962 IRE Awards. Formal presentation of the awards will take place during the 1962 IRE International Convention, on Wednesday, March 28, 1962 at the Waldorf Astoria Hotel, New York, N. Y.

Edward V. Appleton of Scotland will receive the 1962 Medal of Honor "for his distinguished pioneer work in investigating the ionosphere by means of radio waves."

The 1962 Morris N. Liebmann Memorial Award will be presented to Victor H. Rumsey (SM'50–F'60) of the East Bay Subsection of the San Francisco Section "for basic contributions to the development of frequency independent antennas." Henri B. Smets of France will receive the 1962 Browder J. Thompson Memorial Award for his paper entitled "Analysis and Synthesis of Nonlinear Systems," which appeared in the December, 1960 issue of the TRANSACTIONS OF THE IRE PROFESSIONAL GROUP ON CIRCUIT THEORY.

The 1962 Harry Diamond Memorial Award will be presented to William Culshaw (SM'57) of the San Francisco Section "for outstanding accomplishments in the field of microwave optics and interferometry." George A. Morton (A'39–SM'46–F'51) of the Princeton Section will receive the 1962 Vladimir K. Zworykin Award "for his contribution to electronic television through the development of camera and imaging tubes." The 1962 W. R. G. Baker Award will be presented to Marvin Chodorow (A'43–SM'47–F'54) and Tore Wessel-Berg (A'55) of the San Francisco Section for their paper entitled "A High-Efficiency Klystron with Distributed Interaction," which appeared in the January, 1961 issue of the TRANSACTIONS OF THE IRE PROFESSIONAL GROUP ON ELECTRON DEVICES.

CALL FOR PAPERS COLORADO CONFERENCE

The 1962 International Conference on Precision Electromagnetic Measurements (formerly Conference on Standards and Electronic Measurements) will be held August 14–16 at the Boulder Laboratories of the National Bureau of Standards, Boulder, Colo. This Conference is sponsored by the Radio Standards Laboratory of the NBS, the Professional Group on Instrumentation of the IRE, and the Instrumentation Division of the AIEE. The change of name points up the basic aim of the Conference, which is the advancement of standards and accurate measurement throughout the coherent frequency spectrum. In addition, the 1962 Conference will give more emphasis to the impact of quantum electronics and space physics on precision electromagnetic measurements. In its concern with the strictly technical aspect of measurement, this Conference will not include papers describing the large-scale organization and operation of standards laboratories (as did the conferences of 1958 and 1960). The interest in both fields is so great that the two subjects may not be adequately handled in a single conference. The second field, however, may be covered in a second conference which would closely precede or follow the International Conference on Precision Electromagnetic Measurements.

Original papers in the following areas will be considered for presentation: (1) atomic frequency and time, (2) determination of conductivity and complex (tensor) electric and magnetic susceptibilities, (3) direct current and low frequency measurements, (4) radio frequency and microwave measurements, (5) quantum electronics in precision measurements, (6) electromagnetic measurements for space exploration, and (7) data reduction in precision measurements.

Titles and summaries will be accepted in English or French but papers will be published in English. As before, it is planned to publish a conference record by including the conference papers in one issue of the IRE TRANSACTIONS ON INSTRUMENTATION. Papers should be submitted to the Program Chairman, Dr. George Birnbaum, Hughes Research Laboratory, Malibu, Calif. These should be in the form of 500 to 1,000 word summaries.

The deadline for submission of summaries is March 15, 1962. Notifications of acceptance or rejection will be mailed by April 15, 1962. The full technical program will be announced about June 1, 1962. This, with the registration brochure, will be mailed to those attending the 1960 Conference, members of IRE-PGI, IRE-PGMTT, and AIEE Instrumentation Division. Further details relating to the Conference, including registration forms (available about June 1), can be obtained by writing: J. F. Brockman, National Bureau of Standards, Boulder, Colo.

SYMPOSIUM ON SCINTILLATION AND SEMICONDUCTOR COUNTER

The Eighth Scintillation and Semiconductor Counter Symposium will be held March 1–3, 1962 at the Hotel Shoreham, Washington, D. C. These Symposia have been instrumental in bringing together those interested in these nuclear radiation detectors either as users or designers. Both the theoretical and practical aspects of the newer techniques, equipments, and components will be covered in invited and contributed papers by workers in the field.

The Symposium will consist of five daytime sessions covering the following topics: scintillators, photomultipliers and image tubes, counter systems and applications, semiconductor detectors and scintillation track imaging and spark chambers. There will be an evening discussion session, preceded by a banquet, on March 1. A limited exhibition will display some of the latest equipment available in the scintillation and semiconductor counter field.

The AEC, AIEE, IRE, and NBS are the sponsors of the Symposium. Further information can be obtained from G. A. Morton, Chairman, Scintillation and Semiconductor Counter Symposium Committee, RCA Laboratories, Princeton, N. J.

Calendar of Coming Events and Author's Deadlines*

(*Continued from page 14A*)

May 4–5: Bay Area Symp. on Reliability and Quality Control, U. S. Naval Postgraduate School, Monterey, Calif. (DL*: Dec. 15, 1961, F. B. Durand, 553 Connemara Way, Sunnyvale, Calif.)

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

May 14–16: NAECON, Dayton, Ohio.

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

May 22–24: Nat'l. Microwave Theory and Techniques Symp., Boulder, Colo. (DL*: Dec. 18, 1961, R. W. Beatty, Nat'l. Bur. Standards, Boulder, Colo.)

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

May 23–25: Nat'l. Telemetering Conf., Sheraton Park Hotel, Washington, D. C. (DL*: Dec. 15, 1961, D. G. Mazur, Goddard Space Flight Center, Greenbelt, Md.)

May 24–26: IRE 7th Region Conf., Seattle, Wash.

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

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

June 25–30: Symp. on Electromagnetic Theory and Antennas, Tech. Univ. of Denmark, Copenhagen. (DL*: Dec. 1, 1961, H. L. Knudson, Øster Volgade 10 G, Copenhagen K, Denmark)

June 27–29: Joint Automatic Control Conf., New York Univ., New Yrok, N. Y. (DL*: Oct. 15, 1961 (abstracts), Nov. 15, 1961 (papers), H. J. Hornfeck, Bailey Meter Co., 1050 Ivanhoe Rd., Cleveland 10, Ohio)

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

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

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

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

Sept. 3–7: Internat'l. Symp. on Information Theory, Brussels, Belgium (DL*: Jan. 15, 1962, Dr. F. L. Stumpers, Philips Research Labs., Eindhoven, Netherlands)

* DL = Deadline for submitting abstracts.

Call for Papers

6th National Convention on Military Electronics (MIL-E-CON 1962)

June 25–27, 1961

Shoreham Hotel, Washington, D.C.

The MIL-E-CON 1962 Convention is sponsored by the IRE Professional Group on Military Electronics. Ralph I. Cole, Director, Government Project Service, Westinghouse Air Brake Co., Falls Church, Va., is Convention President for MIL-E-CON 1962.

Papers presenting original work in military electronics are invited for this meeting. Suggested topics include, but are not limited to the following:

- Current Problems of Space Technology
- Space Electronics
- Ranging and Tracking
- Electronic Propulsion
- Data Handling Systems
- Molecular Electronics and Systems
- Guidance and Control
- Inertial Systems
- Reconnaissance Systems
- Communication Systems
- Operational Analysis
- Reliability

The technical program will include both classified (confidential) and unclassified sessions, with the Air Force Systems Command sponsoring the classified sessions. An unclassified and bound *Proceedings* of the Convention will be available at the meeting.

Prospective authors are requested to furnish the following information *not later than February 1, 1962:* three copies of a 750-word, unclassified abstracts of the proposed paper plus the name and address of the author(s) and name and address of company or affiliation; three copies of biographical sketch of the author(s), including education, experience, memberships in professional societies, honors, publications. Each author must obtain the appropriate military and company clearances for his abstract.

Send abstracts to J. J. Slattery, F316, The Martin Co., Baltimore, Md.

WESCON EXECUTIVE COMMITTEE NAMED

Leadership of WESCON for 1962 was assigned recently to four prominent Southern California electronics executives. WESCON will be held in the Los Angeles Memorial Sports Arena on August 21–24, 1962. Assignments to WESCON's executive committee were made by the board of directors during its annual meeting.

Donald C. Duncan, President of Duncan Electronics, will serve as Chairman of the 1962 WESCON Board, and Bruce S. Angwin, Western Regional Manager of General Electric's receiving tube department, as Chairman of the Executive Committee. Edward C. Bertolet, Vice President of Behlman-Invar Electronics is Convention Director, and S. H. Bellue, Vice President of Osborne Electronics, is Show Director. Don Larson, WESCON Manager, is the fifth member of the executive committee.

Bay Area members of the eight man WESCON Board include John A. Chartz, Dalmo Victor, and Meyer Leifer, Ampex Instrumentation Products, all newly elected to membership, and Dr. John V. N. Granger, President of Granger Associates, and Calvin K. Townsend, Chairman of the Board of Jennings Radio Manufacturing, both former members of the Board.

Plans for WESCON in Los Angeles include a 1200-booth technical exhibit, 200 more booths than in 1960 when WESCON was last held in Los Angeles. A 94,000-square foot pavilion will supplement exhibit areas within the permanent Sports Arena facilities.

JOINT COMPUTER CONFERENCE COMMITTEES NAMED

Executive and committee responsibilities for the 1962 Spring Joint Computer Conference and Exhibition in San Francisco May 1–3, 1962 have been announced. The technical program and exhibits will be at the Fairmont Hotel. Sponsorship of the national meeting is by the American Federation of Information Processing Societies, representing the IRE, the AIEE and the Association for Computing Machinery.

General Chairman is G. A. Barnard of Philco Western Development Laboratories, Palo Alto. Vice Chairman is Dr. H. D. Crane of Stanford Research Institute and Secretary-Treasurer is R. A. Isaacs of Philco Western Development Laboratories.

Heading the technical program is Dr. R. I. Tanaka of Lockheed Missiles and Space Co., Palo Alto. Program Vice Chairman is Dr. R. C. Minnick of Stanford Research Institute, and Associate Chairmen for special sessions are J. E. Sherman of Lockheed Missiles and Space Co. and R. J. Andrews of International Business Machines, San Jose.

G. A. Barnard (left), General Chairman, and Dr. H. D. Crane (right), Vice-Chairman, of 1962 Spring Joint Computer Conference.

J. W. Ball of Pacific Telephone Co., Sacramento is Exhibits Chairman. Arrangements Chairman is R. G. Glaser of McKinsey and Co., San Francisco, and Vice Chairman of arrangements is R. Smith of Smith-Corona Marchant, Oakland.

Printing and mailing are being handled by W. O. Hamlin, Chairman, and R. Johnston, Vice Chairman, both of Fairchild Semiconductor, Mountain View. Publications Chairman is E. T. Lincoln of IBM, San Jose, who is being assisted by J. J. McNulty, Vice Chairman, also of IBM, San Jose. Chairman and Vice Chairman of the Registration committee are D. E. Eliezer and L. Rinsler, both of IBM, San Jose. N. S. Jones of Friden, Inc., San Leandro, is Chairman of the Public Relations committee and R. D. Painter of Fairchild Semiconductor is Vice Chairman.

A committee headed by Miss M. G. Conley, Chairman, and Miss A. Nemi, Vice Chairman, both with IBM in San Francisco, will plan a social program for women in San Francisco in connection with the conference.

Consulting assistance to the conference committees is being given by R. E. Teasdale of the law firm Morrison, Foerster, Holloway, Shuman and Clark, San Francisco, for legal matters; John L. Whitlock of John Leslie Whitlock Associates, Oakton, Va., for exhibits; and William C. Estler of Palo Alto, for public relations.

WINTER CONVENTION ON MILITARY ELECTRONICS

The 1962 Winter Convention on Military Electronics to be held in Los Angeles, February 7–9, was announced recently by M. D. McFarlane (Robertshaw-Fulton), Chairman.

An opening panel session will present J. Rubel, Assistant Secretary of Defense for Research and Engineering; Dr. F. J. Larsen, Assistant Secretary of the Army, Research and Development; Dr. B. McMillan, Assistant Secretary of the Air Force, Research and Development. Maj. Gen. K. P. Bergquist, Commander of the Air Force Systems Command Electronic Systems Division, will deliver the convention's banquet address on Wednesday night, February 8.

A technical program, under the direction of M. E. Brady, Space Technology Laboratories, and H. L. Richter, Jr., Electro-Optical Systems, will run in concurrent sessions throughout the three-day event. Mr. Brady has singled out communications systems as a subject for special attention in the program. Seventy-five exhibit booths will display new military electronic components, instruments, and systems produced by the military services. A program for women has also been arranged.

The Professional Group on Military Electronics, and the Los Angeles Section, IRE, are sponsors of the convention. Information may be obtained from the IRE Business Office, 1435 S. La Cienega Blvd., Los Angeles 35, Calif.

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LA 40-05B	20-105 VDC	10 V	0- 4 AMP	495
LA 80-05B	20-105 VDC	10 V	0- 8 AMP	780
LA 8-08B	75-330 VDC	30 V	0- 0.8 AMP	395
LA 15-08B	75-330 VDC	30 V	0- 1.5 AMP	560
LA 30-08B	75-330 VDC	30 V	0- 3 AMP	860

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Ripple and Noise.......Less than 1 millivolt rms with either terminal grounded.

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AIEE PACIFIC ENERGY CONVERSION CONFERENCE

A new technical Conference on the subject of Energy Conversion will be sponsored by the San Francisco Section of the AIEE. The Conference will be held in San Francisco on August 13–16, 1962.

This meeting will be devoted to reports on the state of the art and panel discussions an engineering aspects of applications in the field of energy conversion. Included will be discussions on thermionics, photovoltaics, fuel cells, energy storage, thermo electrics and magnetohydrodynamics.

Authors interested in submitting papers are requested to furnish three copies of a brief summary by January 15, 1962. Submission should be made to H. N. Skow, General Program Committee Chairman, Westinghouse Electric Corporation, 410 Bush St., San Francisco, Calif.

CONFERENCE ON SELF-ORGANIZING SYSTEMS

A conference on Self-Organizing Systems will be held on May 22–24, 1962 at the Museum of Science and Industry, Chicago, Ill., co-sponsored by the Information Systems Branch of the Office of Naval Research and the Armour Research Foundation.

The objective of the Conference is to bring together research workers who are concerned with the evolution of self-organizing information systems. While improved understanding and modelling of cognitive, learning and growth processes is clearly of interest, this Conference is primarily concerned with these fields only insofar as they interact with the major objective. Since there has been greatly increased emphasis placed on self-organizing systems, it would seem that this might be an appropriate time to evaluate recent progress and to consider the future directions of research. Therefore, this topic will be examined in depth, with particular emphasis placed upon the more salient research of the past three years.

Attendance is open to all interested technical personnel. Further information and a preliminary Conference program, when available, may be obtained by contacting G. T. Jacobi, COSOS Conference Secretary, Armour Research Foundation, 10 West 35 St., Chicago 16, Ill.

CONVENTION RECORDS FOR SEVENTH NATIONAL COMMUNICATIONS SYMPOSIUM

The National Communications Symposium Convention Record which was presented to attendees as part of the technical program at the Symposium, is available in a limited number of copies containing the complete papers which were summarized in the Symposium technical sessions as follows:

Seventh National Communications Convention Record (1961) $4.50 postage paid.

Sixth National Communications Convention Record (1960) $4.50 postage paid.

Payments may be made to the IRE National Communications Symposium in care of IRE Secretariat, Room 223B, Mohawk Valley Technical Institute, Armory Drive, Utica, N. Y. Payments are to be made *in advance* of shipment and, in the case of foreign orders, are to be made payable in United States currency.

SYMPOSIUM ON PLASMA PHYSICS

There will be a Symposium on Interaction Phenomena in Plasmas, July 2–4, 1962, at the Chalmers University of Technology in Gothenburg, Sweden. The Symposium will deal with some aspects of plasma physics which will not be discussed extensively during the Copenhagen Symposium on Electronmagnetic Theory and Antennas. Registration forms for the Gothenburg Symposium can be obtained from the Secretary General, Dr. H. Wilhelmsson, Research Laboratory of Electronics, Chalmers University of Technology, Gothenburg, Sweden.

AIR FORCE MARS ANNOUNCES SCHEDULE

The Air Force MARS Eastern Technical Net broadcasting on 3295, 7540, and 15,715 kc announces the following schedule of broadcasts on Sundays, 2–4 P.M. EST:

Jan. 7—"New Electron Tubes for the Modern Era," Dr. J. E. Beggs, Research Associate, General Electric Research Laboratory.

Jan. 14—"Thermionic Integrated Micro-Module Circuits for High Temperature Environments," A. P. Haase, Manager, Receiving Tube Department, General Electric Company.

Jan. 21—"Applications of Transistors," F. B. Smith, RCA Laboratories.

Jan. 28—"General Discussion of Semiconductor Devices," E. O. Johnson, Research Engineer, RCA Laboratories.

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

Dr. Constantine D. J. Generales, Jr. (SM'59) and Dr. Wernher von Braun are shown above receiving honorary certificates for original research in space science and space medicine from the Greek government through the Hellenic Astronautical Society. *Left to right:* Mr. Triantaphyllides of the H.A.S., Consul General Dr. Basil Vitsaxis of Greece, Mrs. von Braun, Dr. von Braun, and Dr. Generales.

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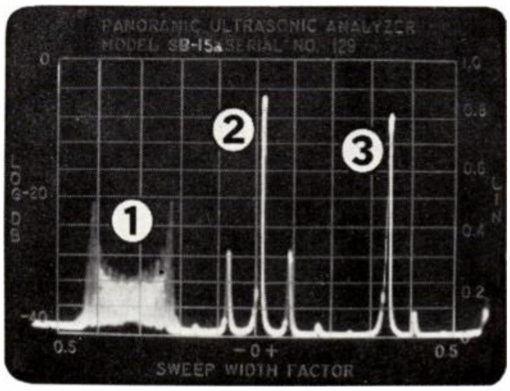

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IF Bandwidth: Variable, 100 cps to 4 kc.

Sweep Rate: Variable, 1 cps to 60 cps.

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Obituaries

Mario Tchou (M'54), who was Director of the Laboratorio Ricerche Elettroniche Olivetti, Milan, Italy, died suddenly on November 9, 1961.

He was born on June 26, 1924. From 1942 to 1945 he attended the University of Rome. In 1947 he received the B.S. degree in electrical engineering from Catholic University of America, Washington, D. C., and in 1949 he received the M.S. degree in physics from Polytechnic Institute of Brooklyn, N. Y.

From 1947 to 1952 he was an Instructor of Electrical Engineering at Manhattan College, New York, N. Y. From 1952 to 1954 he was an Associate Professor of Electrical Engineering at Columbia University, New York, N. Y.

Mr. Tchou was responsible, to a large extent, for the formation of the Italy Section of the IRE and had long been active in IRE affairs. At the time of his death he was Secretary-Treasurer of the Italy Section.

Richard W. Johnston (SM'56), who was Chairman of the IRE Nuclear Techniques Committee from 1958 to 1960, died in Bethesda, Md., on November 17, 1961.

R. W. Johnston

He was born on May 29, 1917 in Boston, Mass. In 1947, he graduated from the Massachusetts Institute of Technology, where he majored in electrical engineering. During World War II, he was assigned to the special engineering detachment of the Manhattan Engineering District, Oak Ridge, Tenn. In 1946, he resumed his studies and worked as an electronic engineer for the Stevens-Arnold Co., Boston, Mass., doing research and development work on radar equipment.

Since 1948, he had been with the Atomic Energy Commission. In 1952 he was named Chief of the technical coordination section of the radiological physics and instruments branch, and in 1957 he became Head of the entire branch.

Mr. Johnston was a recipient of the Arthur S. Flemming Award for outstanding government. He was Chairman of ASA Sectional Committee on Nuclear Instrumentation, a member of the Administrative Committee of the IRE Professional Group on Nuclear Science, the Instrument Society of America, the Electronics Industries Association, and the American Standards Society.

Gordon W. Olive (A'29–VA'39–F'51) who was Director General of Engineering at the Canadian Broadcasting Corporation, Gananoque, Ontario, died on October 22, 1961.

He was born on March 18, 1898 in Montreal. From 1919 to 1922 he attended McGill University. He served overseas during World War I with the McGill Siege Battery. In 1923, he built and operated radio broadcasting station CFCO for Semmelack and Dickson Company, while he was manager of the radio department of that firm. In 1924, he joined the staff of the radio department of Canadian National Railways as a radio engineer. He was appointed Technical Assistant to the Director of radio in 1927.

Mr. Olive was a member of the Frequency Modulation Committee of the IRE from 1946 to 1947, and a member of the Modulation Systems Committee from 1947 to 1948.

Call for Papers For Chicago Conference

The 1962 Chicago Spring Conference on Broadcast and Television Receivers is planned for June 18 and 19, and will again be held at the O'Hare Inn.

Papers are sought that would be of particular interest to those in the home entertainment radio and television industry, especially papers dealing with new products, and with new concepts of component and circuit design. The deadline for receipt of papers is February 15, 1962. At their earliest convenience, potential authors are asked to submit in triplicate a fifty to one hundred (50–100) word summary including title of the paper and name of author(s), together with company affiliation and position. Papers should be limited to approximately 2500 words, and the presentation to twenty minutes.

Authors will be notified of the selection of papers early in March . Biographical information will be requested at that time, but complete papers will not be required until the time of presentation. Since it is hoped that most of the papers will be published in the PGBTR Transactions, authors will be expected to supply, before presentation, a transcript suitable for such submission. Papers should be submitted to: A. Cotsworth, Papers Committee, Zenith Radio Corp., 6001 West Dickens Ave., Chicago 39, Ill.

Dr. Kalmus Receives Army's Gold Medal Award

Dr. Henry P. Kalmus (A'39–SM'45–F'56), Associate Technical Director of the Diamond Ordnance Fuze Laboratories, Washington, D. C. was presented with the Army's Gold Medal Award for Exceptional Service by Dr. Finn Larsen, Assistant Secretary of the Army (R&D). This is the highest award the Army bestows upon a civilian. Dr. Kalmus was cited for "his numerous inventions applicable to missiles, nuclear weapons, and radar and target detection systems (which) have contributed immeasurably to developments in modern weaponry."

A member of the Washington Academy of Sciences, he received the Dipl. Engr. degree from the Technical University of Vienna in 1930 and his Doctorate from the same University in 1960. After an association of eight years with the Orion Radio Corporation in Europe, Dr. Kalmus came to the United States and was employed in a professional capacity with the Emerson Radio Corporation, New York, N. Y., and with the Zenith Corporation in Chicago, Ill., prior to joining the staff of the National Bureau of Standards (and consequently the Diamond Ordnance Fuze Laboratories) in 1948. Dr. Kalmus holds more than thirty patents in the field of electronics and has published a great number of scientific papers and articles. Among the many honors he has received is the Department of Commerce Gold Medal for Exceptional Service (1954).

Dr. Finn Larsen (*left*) congratulates Dr. Henry P. Kalmus (*right*) upon presentation of the Army's Gold Medal Award for Exceptional Service to Dr. Kalmus.

BREAKTHROUGH!

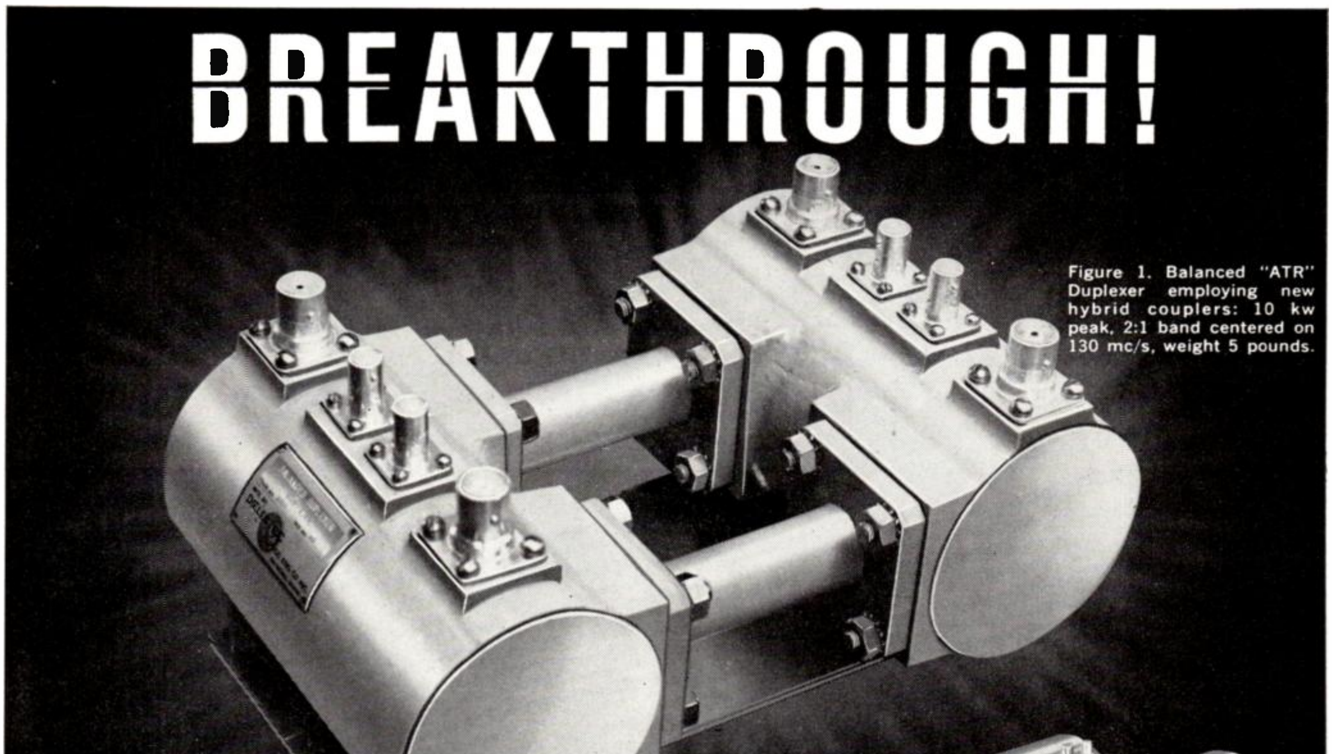

Figure 1. Balanced "ATR" Duplexer employing new hybrid couplers: 10 kw peak, 2:1 band centered on 130 mc/s, weight 5 pounds.

Figure 2. Balanced "TR" Duplexer: specifications similar to those of Figure 1.

A new, HIGH-POWER, 3 db, crossover HYBRID COUPLER*

HYBRID COUPLER* that makes possible duplexers of a size and weight heretofore unknown. The crossover feature which allows two outputs on one side provides the maximum flexibility. Unlike many high-power microwave components, this coupler lends itself to design centers as low as 15 mc/s or lower without becoming unreasonably large. Otherwise it has the properties of the classical $\pi/2$ hybrid.

As an example, Figure 1 illustrates application to a balanced "ATR" duplexer centered at 130 mc/s. Over a 2:1 bandwidth, receiver insertion loss is 0.3 db nominal and transmitter insertion loss is less than 0.5 db. Isolation is nominally 30 db. Using tubes especially developed for these duplexers by Tucor, Inc., recovery time is 100 microseconds. This particular unit was furnished for a 10 kw peak, 200 watts average power application; but these ratings can be extended considerably.

Note the size with relation to the connectors. The four larger connectors on outside corners are type C for r-f connections; the four inner are BNC for d-c returns, bypassed, through the r-f circuitry for keep-alive voltages.

In Figure 2 is a balanced "TR" duplexer with type LC connectors for transmitter and antenna. Hidden are type N connectors for receiver and termination. The BNC's on top are for keep-alive voltages. This particular unit has a center frequency of 130 mc/s, and will handle 10 kw peak power, 600 watts average.

Figure 3 illustrates the basic coupler, the heart of these and similar devices. Power in port 1 is split equally between 2 and 3 but 90° out of phase; power 4 is also split equally but 90° out of phase in opposite sign. A similar situation holds for ports 1 and 4 when fed from 2 or 3.

When power components 90° out of phase are fed to ports 2 and 3, the combined output will be from port 1 or 4 depending on sign of the phase difference. A similar situation holds with ports 2 and 3 when power components are fed to ports 1 and 4.

When ports 2 and 3, or 1 and 4, are terminated, a nominal 30 db isolation may be expected between the other two ports over a 2:1 frequency band. Note that the angular orientation between ports 1 & 4 and 2 & 3 is unimportant. This fact greatly eases the packaging problem.

For high power capability in a small, light-weight package, specify *Dielectric*. Further information may be obtained from our Proposal Engineering Department.

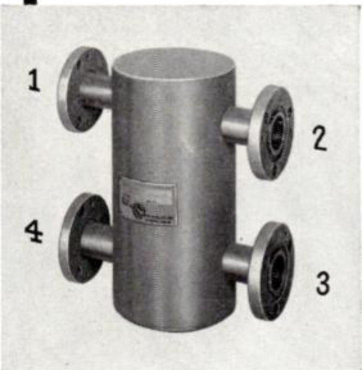

Figure 3. Basic 3 db coupler: orientation between port pairs 1 & 4 and 2 & 3 may be any angle between 90° and 180°. The crossover feature providing two outputs on the same side facilitates application.

*Patent pending

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Eighth National Symposium on Reliability and Quality Control

STATLER HILTON HOTEL, WASHINGTON, D. C., JANUARY 9–11, 1962

The Eighth National Symposium on Reliability and Quality Control will be held January 9–11, at the Statler Hilton Hotel, Washington, D. C. Sponsoring bodies are the IRE Professional Group on Reliability and Quality Control, the AIEE, the EIA, and the ASQC. The formal opening of the Symposium will be Tuesday morning, January 9. Senator Henry M. Jackson will be the banquet speaker. Some 200 papers were nominated as candidates for presentation at this Symposium, fifty per cent more than have ever been submitted in previous years. Registration will be held Monday evening, January 8, from 5:00 P.M. to 10:00 P.M., and Tuesday, January 9, from 8:00 A.M. to 4:30 P.M. Registration fees, including Banquet and Proceedings, are $20.00 for advance payment, and $24,00 for door payment for members of IRE, AIEE, and ASQC, and $22.00 and $24.00 for all others. The Proceedings of past Symposia, as well as this Symposium may be obtained at the Symposium or from the IRE, Editorial Department, 1 East 79 Street, New York, N. Y., at a price of $5.00 each. M. P. Smith, Honeywell, St. Petersburg, Fla. is the General Chairman of the Symposium, and E. F. Jahr, IBM Corporation, Owego, N. Y. is the Program Chairman.

The program of the conference will be as follows:

Tuesday Morning, January 9

Session 1—Keynote Address

"A Management Appraisal of the Reliability Challenge," *C. F. Horne, President, General Dynamics, Pomona.*

Session 2A—Statistical Techniques

Moderator: *Dr. M. Sobel, University of Minnesota.*

"Statistical Independence in Reliability Equations," *A. M. Breipohl, Sandia Corp.*

"Truncation of Sequential Life Tests," *T. L. Burnett, IBM Corp.*

"Confidence Limits for System Reliability from Components Tests," *Dr. W. S. Connor, Research Triangle Inst.*

Session 2B—Program Management

Moderator: *R. H. Dewitt, Office of Electronics, ODDR & E.*

"Reliability Program Requirements for Aerospace Systems and Subsystems," *Major V. J. Bracha, U.S.A.F.*

"PERT-PEP Controls for Reliability and Costs," *Dr. H. G. Romig, Operations Research Inc.*

"Management of Equipment Reliability Activities at USASRDL," *T. J. Edwards, U. S. Army Signal Research and Development Lab.*

Tuesday Afternoon

Session 3A—Statistical Techniques

Moderator: *W. R. Allen, Consultant.*

"Life and Reliability Testing Based on the Weibull Distribution," *Prof. H. P. Goode, Cornell Univ.*

"Screening for Critical Variables Affecting Reliability," *Dr. C. H. Li, General Instrument Corp.*

"Factorial Experiments in Reliability Analysis," *C. L. Barron, Honeywell.*

"Statistical Methods as Design Tools," *E. F. Grey, U.S.A.F.*

Session 3B—Quality Control Techniques

Moderator: *L. J. Jacobson, International Resistance Co.*

"Process Acceptance vs Lot Acceptance," *C. R. Clark, Sandia Corp.*

"Martin-Denver Auto QC," *R. A. Bennett, The Martin Co.*

"Precision through Variations Flow Analysis," *N. L. Enrick, Univ. of Virginia.*

"QC Management for Time-Compressed R & D Programs," *F. A. Barta, Hughes Aircraft Co.*

Tuesday Evening

Session 4—Panel Discussion

Moderator: *Dr. R. C. Seamans, Jr., Associate Administrator, NASA.*

"Reliability—The Key to Space Operations." A panel of four leaders from Government and Industry will discuss the challenge the United States faces in space exploration and the key roll reliability plays in this activity.

Wednesday Morning

Session 5A—Reliability Testing

Moderator: *F. E. Wenger, USAF-AFSC.*

"Analysis Testing for Improved Circuit Reliability," *B. J. Grinnell, IBM Corp.*

"Optimum Reliability Testing for Space Systems," *Mrs. C. D. Hock, NASA.*

"Failure Analysis," *E. W. Kimball, The Martin Co.*

"Estimating Guided Missile System Reliability from R & D Systems Firing Tests," *A. Steinberg, U. S. Army Ordnance Missile Command.*

Session 5B—Mathematics and Reliability Models

Moderator: *Dr. B. Epstein, Consultant.*

"Methodology for System Reliability Analysis," *B. Tiger, RCA.*

"Equipment Self-Repair Techniques," J. Klion, *Rome Air Development Center.*

"Mathematical Analysis of Redundant Matrices," *G. D. Weinstock, ITT Labs.*

"Intertial Guidance System Component Reliability," *L. A. Weaver, Honeywell.*

Session 5C—Training Session

Moderator: *L. J. Paddison, Sandia Corp.*

"Reliability Management by Objectives and Results," *Dr. L. W. Ball, Boeing Co.*

"Practical Statistical Tools for Reliability Measurement and Acceptance," *C. M. Ryerson, EL-TEK Corp.*

"Full Design Definition Is Essential to Reliability," *W. L. Hurd, Jr., Lockheed Missile and Space Co.*

"Contractural Reliability—An Education Program," *R. H. Johnson, The Martin Co.*

Wednesday Afternoon

Session 6A—Reliability Testing

Moderator: *P. D. Darnell, Bell Telephone Labs.*

"CAMESA Reliability Test Program," *A. P. Harris, Canadian Agency.*

"Practical Aspects of a High Reliability Specification," *T. W. Gross, Lockheed Missile and Space Co.*

"Reliability Evaluation by Application Derating," *K. W. Davidson, Texas Instruments Inc.*

"Acceleration Factor Determination on Glass Capacitors," *L. D. Hines, Corning Glass Works.*

Session 6B—Reliability and Quality Control Education

Moderator: *H. C. Jones, Univ. of Maryland.*

"Effective Reliability—Through Education," *Dr. P. H. Zorger, The Martin Co.*

"AGREE in Action," *G. W. Lindsay, U.S.A.F.*

"A Reliability Slide Rule," *B. Ellison, International Electric Corp.*

"A Comprehensive Education and Training Program for Design Reliability," *F. E. Marsh, Boeing Co.*

Session 6C—Training Session

Moderator: *Dr. L. W. Ball, Boeing Co.*

"The Problem of Human Initiated Failures," *D. Meister, General Dynamics—Astronautics.*

"Reliability Through Process Capability Studies," *G. O. Hawley, Sandia Corp.*

"A Theory of Component Part Life Expectancies," *D. R. Earles, AVCO Corp.*

"A Missile Reliability Reporting System," *I. R. Whiteman, C.E.I.R., Inc.,*

Wednesday Evening

Banquet

"A Senator Looks at the Problem of Reliability in National Survival," *Henry M. Jackson, U. S. Senator, State of Washington.*

1961 National Reliability Award—presented by *M. P. Smith, General Chairman.*

1961 IRE Professional Group on Reliability and Quality Control Award—presented by *L. J. Paddison, Chairman, PGRQC.*

1961–1962 ASQC Electronics Division Award—presented by *H. D. Voegtlen, Division Chairman.*

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Thursday Morning

Session 8A—Reliability and Quality Control Programs

Moderator: *D. A. Hill, Hughes Aircraft.*

"Reliability Program Based on Bomb-Nav. System Experience," *F. W. Blackburn, Sperry Rand Corp.*

"Reliability Research on Components for Space," *E. E. Brueschke, Hughes Aircraft Co.*

"Quality Assurance in the Field of Nuclear Ordnance," *C. E. Smith, Jr., Sandia Corp.*

"Reliability Emphasis on Production Contracts," *J. A. Marshik, Honeywell.*

Session 8B—Maintainability

Moderator: *E. Harwood, Office of Electronics, ODDR & E.*

"Simulation of Aircraft Electronics Maintenance," *D. S. Ellis, Hughes Aircraft Co.*

"The Human Element in the Maintenance Package," *Dr. H. W. Page, IBM Corp.*

"Maintainability Prediction and Measurement," *B. L. Retterer, RCA Service Co.*

"Some Considerations of Scheduled Maintenance," *R. Meyers, International Electric Corp.*

Thursday Afternoon

Session 9A—Reliability Prediction and Measurement

Moderator: *Dr. N. E. Golovin, National Aeronautics and Space Administration.*

"Operational Reliability of the Athena Computer," *G. P. Anderson, Sperry Rand Corp.*

"Reliability Predictions for Multi-Mode Electronic Systems," *J. A. Connor, Astro Reliability Corp.*

"Correlation of AGREE Testing with Operational Reliability," *G. W. Lindsay, U.S.A.F.*

"Development of New Prediction Techniques," *I. Bosinoff, Sylvania Electric Systems.*

Session 9B—Reliability Design Techniques

Moderator: *A. W. Rogers, U.S.A.S.C.—R & D.*

"General Usage Assemblies for Navy Electronic Equipment," *T. E. McDuffie, N. Y. Naval Shipyard.*

"Increasing Reliability through the QUAD Technique," *R. M. Fasano, Sylvania Electric Products, Inc.*

"Reliability and Forced Air Cooling," *J. L. Keller, Sperry Rand Corp.*

"Determination and Use of Failure Patterns," *R. L. Horn, Boeing Co.*

Third National Winter Convention on Military Electronics

Ambassador Hotel, Los Angeles Calif., February 7–9, 1962

The Third National Winter Convention on Military Electronics will be held at the Ambassador Hotel, Los Angeles, Calif., on February 7–9, 1962. The convention is sponsored by the IRE Professional Group on Military Electronics and the Los Angeles Section of the IRE. Included in the three day program will be technical sessions, up to 75 military systems exhibits by such firms as Lockheed, Autonetics and Westinghouse, field trips, banquet, luncheon, reception, and a ladies program.

The banquet speaker will be Major General Kenneth P. Bergquist, Commander, Electronic Systems Division, Air Force Systems Command. The luncheon speaker will be Vice-Admiral Robert B. Pirie, Deputy Chief, Naval Operations.

The program of the Convention will be as follows:

Wednesday, February 7

Session I—Panel Discussion

"Trends in Weapon System Development."

Moderator: *J. H. Rubel, Asst. Secretary of Defense for Research and Engineering DOD.*

Panel Members: *Dr. F. J. Larsen, Asst. Secretary, Research and Development, U. S. Army; J. H. Wkelin, Jr., Assist. Secretary, Research and Development, U. S. Navy; Dr. B. McMillan, Asst. Secretary, Research and Development, U. S. Air Force.*

Session IIA—Undersea Warfare and Sonar (Confidential)

"Sonar Equipment Design Problems Peculiar to Advanced ASN Vehicles," (Confidential), *Dr. D. E. Andrews, Jr., Head Acoustics Div., U. S. Navy Electronics Lab., San Diego, Calif.*

"Adjustable Doppler Tolerance Waveform for Correlation Detection," (Confidential), *Dr. P. R. Kan, Space Technology Labs., Los Angeles, Calif.*

"On the Possibility of All-Weather Passive Nonacoustic Submarine Detector," (Confidential), *C. R. Hill, Military-Space Div., Electro-Optical Systems, Inc., Pasadena, Calif.*

"Improving the Sonar Search Capabilities of Ocean Escorts," (Confidential), *D. L. Heebner, Hughes Aircraft Co., Fullerton, Calif.*

"Hydro-Acoustic Transducer," (Confidential), *S. Wisotsky.*

"Deep Water Transponder Using High Pressure Components," (Unclassified), *A. E. Johanson, Heavy Military Electronics Dept., General Electric Co., Syracuse, N. Y.*

Session IIB—Maintainability (Unclassified).

Session IIC—Maintainability (Unclassified).

"The Design of Complex Systems for Automated Testability," *R. A. Kirkman and M. C. Peterson, Space Technology Labs., Inc.*

"Design of Missile Systems for Maintainability," *M. V. Ratynski, Electronic Systems Div., Air Force Systems Command, Bedford, Mass.*

"Reliability Inputs to Maintainability Design Decisions," *R. E. Redfern, Space Guidance Center, IBM, Owego, N. Y.*

"Theory of Integrated Maintainability and Reliability Analysis for Modular Equipment," *Dr. E. J. Althaus, Hughes Aircraft Co., Culver City, Calif.*

Session IID—Military Equipment Design (Unclassified).

"Human Engineering and Inertial Guidance Systems," *G. J. Ferwerda, AC Spark Plug Div., General Motors Corp., Milwaukee, Wis.*

"Device 3G36," *E. F. Moretta, U. S. Naval Training Device Center, Port Washington, N. Y.*

"The Malfunction Simulation System Data-Flow Class of Trainer," *G. B. Rozran, Burroughs Labs., Burroughs Corp., Paoli, Pa.*

"Configuration of an Inertial Guidance Computer for an Advanced Ballistic Missile," *D. K. Richardson, Advanced Armaments Lab., Hughes Aircraft Co., Culver City, Calif.*

Thursday, February 8

Session IIIA—Radar I (Confidential).

"BMEWS: Its Present and Future in Missile and Satellite Surveillance," (Confidential), *A. D. Korbin, Defense Electronic Products, RCA, Moorestown, N. J.*

"A Technique to Discriminate Nose Cones During Midcourse," (Confidential), *G. A. Cato, Electro-Optical Systems, Inc., Pasadena, Calif.*

"A Microwave Missile Detection Technique," (Confidential), *Capt. P. Tokareff, Jr., Reconnaissance Lab., Aeronautical Systems Div., Wright-Patterson AFB, Ohio.*

"Improved Automatic Detection for Radar," (Confidential), *J. W. Caspers, U. S. Navy Electronics Lab., San Diego, Calif.*

Session IIIB—Tracking, Telemetry and Command (Confidential).

Session IIIC—System and Technical Management (Unclassified).

"An Appraisal of Scientific Techniques in *R* and *D* Management," *Dr. R. J. Freeman, Researnalysis Co., Santa Monica, Calif.*

"The Evaluation of a Proposal for the Design of an Automatic Monorial System for the City of Los Angeles." The remaining two hours will be used for a project which will illustrate how proposals are generated and evaluated.

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Features include built-in leveler and narrow band symmetrical sweep, adjustable frequency markers, and Quick Look readout. Drift, less than ±.02% per hour; residual FM, less than ±.0025% peak.

These units save valuable engineering time by providing constant RF power input to microwave tubes and other components under test. With constant input, variations in output indicate directly component transfer characteristics. Microwave properties are examined over continuous spectrum, assuring accuracy during broad or narrow band testing. Prices: Model 620 series, $2,890-3,340; Model 605, $1,750; Generator Heads, $1,500-1,990; Model 6001B-6004C, $7,250-8,690.

MICROWAVE LEVELERS

Ideal instruments for maintaining constant power output (±1 db) with long term stability. CW or square wave modulation. Fast response, very high gain assure flat RF output from BWO's. Also available: Model 700 leveler amplifier, which may be used with external RF components. Prices: Microwave Levelers, $875-990; Model 700, $575.

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Alfred manufactures microwave amplifiers for virtually every requirement — general purpose, medium power, high gain, and low to medium noise figure. All models feature low spurious modulation and stable operation. General purpose amplifiers provide 30 db gain and 10 mw power output with 25 db noise figure. Medium power models offer power up to 10 watts. High power amplifiers provide up to 1 Kw pulse power. Low noise amplifiers provide noise figure from 6 to 15 db at frequencies from .5 to 12 kmc.

In addition to standard amplifiers, Alfred offers periodic or permanent magnet focused amplifiers where light weight and low input power is required as well as amplifiers designed specifically for phase modulation. Prices: General purpose amplifiers, $1,490-1,690; Medium power amplifiers, $1,550-3,590; Low noise amplifiers, $3,150-4,990.

POWER SUPPLIES

Alfred furnishes four basic types of power supplies. Model 250 Traveling Wave Tube Supply operates low and moderate power traveling wave amplifier and oscillator tubes. This instrument provides all normal sources — helix, collector, four separate anodes, grid, heater, solenoid and blower—from one compact unit.

Microwave Power Supplies Model 252, with the unregulated and regulated solenoid supplies Model 253 and 254, operate all presently known low and medium noise figure TW tubes. Model 252 furnishes electrode and heater power for permanent magnet focused TW tubes. Electromagnet focused TW tubes require in addition either Model 253 or 254 solenoid supply. Alfred's Sweeping Power Supplies serve as general purpose sources for either electronically swept or fixed frequency operation of voltage tuned magnetrons, BWO's and similar microwave tubes. Alfred's floating high voltage supplies provide extremely stable and highly regulated DC voltages, featuring wide voltage and current ranges, very small ripple and accurate voltage adjustment. Prices: Model 250, $1,990; Models 252, $890; 253, $300; 254, $400; Sweeping Power Supplies, $1,650-1,690; High Voltage Supplies, $690-1,090.

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Session IIID—Displays (Unclassified).

Thursday Afternoon

Session IVA—Missile and Space Systems (Confidential).

Session IVB—Information and Data Handling Systems (Unclassified).

Session IVC—System Engineering (Unclassified).

"A New Communication System Design Method Responsive to User Requirements," *G. S. Mauksch and H. Rapaport, International Electric Corp., Paramus, N. J.*

"A Modular Approach to Air Defense and Air Control Systems," *R. Stuckelman, Litton Systems, Inc., Canoga Park, Calif.*

"Use of Finite Markov Chains in the Reliability Analysis of Systems," *K. Hall, and K. Zendere, Sylvania Electronic Systems, Reconnaissance Systems Lab., Mountain View, Calif.*

"Combined Analog—Digital Simulation of Space Vehicle Systems," *W. R. Metzger and R. P. Schuk, General Electric Co., Syracuse, N. Y.*

"Determination of the Maximum Allowable Weapon System Error," *J. D. Palmer and J. A. Nickel, Research Institute, University of Oklahoma, Norman, Okla.*

Session IVD—Navigation (Unclassified).

"Aircraft Navigation by Doppler in the North Atlantic," *R. White, Trans-World Airlines, Kansas City, Mo.*

"Inertial Navigator for Aircraft," *D. J. Atwood Jr., Boston Research and Development Labs., AC Spark Plug Div., General Motors, Co., Wakefield, Mass.*

"Geodesy on the Transit Program," *Dr. R. B. Kushner, Applied Physics Lab., Johns Hopkins University, Silver Spring, Md.*

"The Determination of Position and Velocity Vector of a Navigator Using a Satellite for Reference," *Dr. I. Kantor, RCA, Moorestown, N. J.*

"Damp-Transit, the First Operational use of the Transit System," *L. Farkas, RCA, Moorestown, N. J.*

Friday, February 9

Session VA—Reconnaissance and Electronic Warfare (Confidential).

"Military Weapons Systems in a Countermeasures Environment," (Confidential), *S. Adelman and S. M. Shinners, Sperry Gyroscope Co., Great Neck, N. Y.*

"Antijam Communication," (Confidential), *R. E. Hechert, Aeronutronic, Newport Beach, Calif.*

To be announced.

"Spectrum Surveillance System," (Unclassified), *A. J. Valluzi, Pan American, and J. C. Rote, ITT Federal Labs., Nutley N. J.*

Session VB—Communication Systems (Unclassified).

"Counting the Effects of Decreasing Solar Activity," *E. I. Finke, U.S.N., Office of Chief of Naval Operations, Washington, D. C.*

"Some Preliminary Results from an Oblique Incidence Step Sounder Link," *W. R. Vincent and R. A. Rack, Stanford Research Institute, Menlo Park, Calif.*

"The Echo Control Problem in Satellite Communications," *F. C. Hunter and J. A. Stewart, General Telephone and Electronic Labs., Menlo Park, Calif.*

"Large Communication Systems Analysis by Computer," *Mrs. E. M. Young and R. A. Rack, Stanford Research Institute, Menlo Park, Calif.*

"On Experimental Communication Circuit Utilizing Echo I," *G. T. Bergemann, and C. M. Beamer, Collins Radio Co., Cedar Rapids, Iowa.*

Friday, February 9

Session VA—Reconnaissance and Electronic Warfare (Confidential).

Session VB—Communications Systems (Unclassified).

Session VC—Aerospace Ground Equipment (Unclassified).

"Launch Control for Ballistic Missile Weapon Systems," *W. C. Knoblock, Space Technology Labs., Los Angeles, Calif.*

"Advanced Concepts in Spacecraft Evaluations," *W. A. Schanbacher, Space Technology Labs., Los Angeles, Calif.*

"An Automatic System for Real-Time Reduction of Telemetry Data," *L. D. Healy and M. S. McFarland, Missiles and Space Div., Lockheed Aircraft Corp., Sunnyvale, Calif.*

"A Programmable Digital Computer for Automatic Checkout Systems," *A. J. Monow and R. W. King, Sperry Gyroscope Co., Great Neck, N. Y.*

Session VD—Reliability (Unclassified).

"Reliability vs Cost Tradeoff," *E. S. Winlund, Manager, Reliability and Quality Control, General Dynamics Electronics, San Diego, Calif.*

"The Game of Reliability—Amateur or Professional," *R. P. Bosche, Electronics and Ordnance Div., Avco Corp., Cincinnati, Ohio.*

"Efficiency in System Reliability Programs," *D. Rubenstein, Advanced Electronics Center, General Electric Co., Ithaca, N. Y.*

"Reliability vs Value Engineering," *Dr. E. A. Keller, Military Electronics Div., Motorola Inc.*

"Reliability Assurance of Military Electronics Equipment Through Effective Reliability Control," *S. A. Rosenthal, Arma Div., American Bosch Arma Corp.*

Friday Afternoon

Session VIA—Radar II (Confidential).

To be announced.

"A Microwave Receiver for Instantaneous Polarization Measurement," (Confidential), *J. C. de Brockert and M. Crane, Electronics Labs., Stanford, Calif.*

"A New High Performance Doppler Radar Frequency Tracker," (Unclassified), *R. M. Pincus, GPL Div., General Precision, Inc., Pleasantville, N. Y.*

"A New Optical Radar," (Unclassified), *R. E. Johnson, J. E. Ticen and A. P. Sheppard, The Martin Co., Orlando, Fla.*

Session VIB—Missile and Space Power Systems (Unclassified).

"Requirements and State of the Art of Electric Systems for Missile and Space Vehicles," *Dr. A. Krausy, Space Technology Labs., Los Angeles, Calif.*

"Battery Experience on the Ballistic Missile Programs," *E. M. Morse and D. L. Lawson, Eagle-Picher.*

"Solar Energy Conversion Systems for Long-Duration Space Missions," *R. L. Robinson, Space Technology Labs., Los Angeles, Calif.*

"Power Supply Problems in Manned Space Missions," *D. L. Ken, MSVD, General Electric Co., Valley Forge, Pa.*

Session VIC—Electron Devices (Unclassified).

"Wide-band Single-Diode Parametric Amplifiers of Specified Gain, Bandwidth and Signal Load Impedance," *A. Mayer, Electronic Defense Labs., Sylvania Electric Products Inc., Mountain View, Calif.*

"Application of Diode Parametric Amlifiers to Military Radar Systems," *H. D. Tenny, Melabs, Palo Alto, Calif.*

"A Reliable Wide-Band Transmitting Amplifier for Space Applications," *H. R. Mathwick and J. D. Kiesling, Astro-Electronics Div., RCA Princeton, N. J.*

"Laser Ranging on Single Photon Events," *W. R. Sooy and M. L. Wiederhold, Research and Development Div., Hughes Aircraft Co., Culver City, Calif.*

"The Laser," *Dr. T. Maiman, Quantatron Inc., Santa Monica, Calif.*

Session VID—Communications Systems and Equipment (Unclassified).

"Monitor and Control Concepts for Technical Management of Military Communications," *R. A. Skene and R. L. Tate, Hoffman Electronics Corp., Los Angeles, Calif.*

"Characteristics and Development of the Communication Equipment on the Mercury Capsules," *L. Leopold, Space Task Group, NASA, Langley Field, Va.*

"Tanlock: A Phase Lock Loop of Extended Tracking Capability," *W. S. Pope, Space and Information Systems Div., North American Aviation, Inc., Downey, Calif.*

"A High Sensitivity Panoramic Receiver Using Pulse Compression Techniques," *F. J. Mueller, R. L. Goodwin and I. M. Saffitz, Airborne Instruments Lab., Melville, N. Y.*

"Design Approach for Intermediate-Frequency Amplifiers," *J. I. Hanon, Surface Armament Division, Sperry Gyroscope Co., Great Neck, N. Y.*

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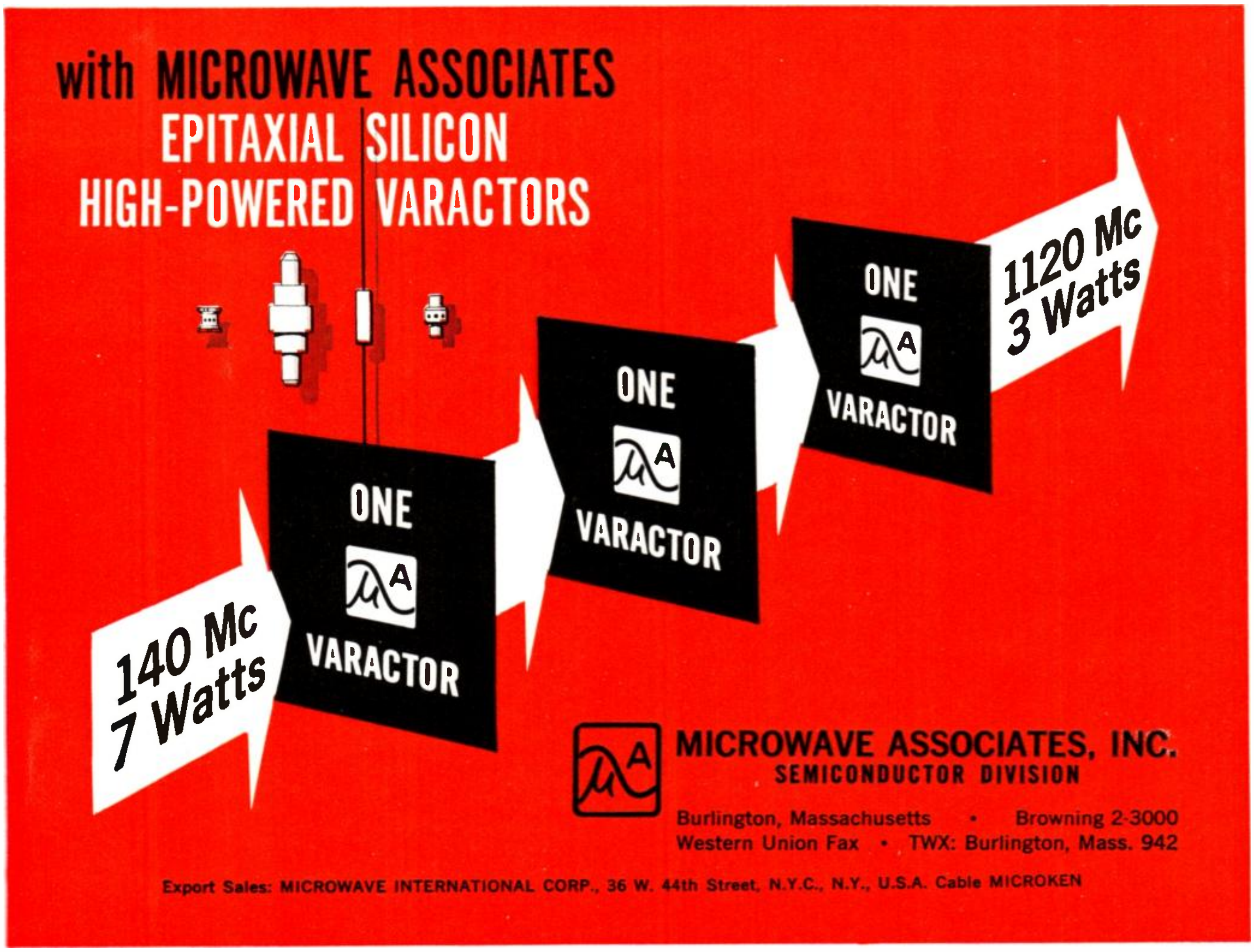

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1962 International Solid-State Circuits Conference

UNIVERSITY OF PENNSYLVANIA AND SHERATON HOTEL, PHILADELPHIA, PA., FEBRUARY 14–16, 1962

The 1962 International Solid-State Circuits Conference, the ninth annual meeting, will feature a 46-paper program during ten daytime sessions at the University of Pennsylvania. Eleven informal evening discussion sessions will also be held at the Sheraton Hotel. The Conference is sponsored by the IRE Professional Group on Circuit Theory, AIEE Committee on Electronics, IRE and AIEE Philadelphia Sections, and the University of Pennsylvania. Illustrated condensations of all the talks will be published in a *Digest of Technical Papers*. A 112-page letterpress printed edition—volume 5 in the series—will be distributed free to all registrants at the meeting.

The program of the Conference will be as follows:

Wednesday, February 14

Session I—Logic

Chairman: *A. W. Lo, IBM Corp., Poughkeepsie, N. Y.*

"Nonlinear Devices for Threshold Logic," *E. P. Stabler, Electronics Lab., General Electric Co., Syracuse, N. Y.*

"New Forms of All-Transistor Logic," *R. H. Beeson and H. W. Ruegg, Fairchild Semiconductor, Div. Fairchild Camera-Instr. Corp., Palo Alto, Calif.*

"A Tunnel-Diode Adaptive Logic Net," *E. N. Powers and D. R. Vernot, Philco Corp., Blue Bell, Pa.*

"The Technology and Application of Optoelectronic Circuits," *J. G. Van Santen, Phillips Research Labs., Eindhoven, Netherlands.*

Session II—Microwave Parametric Circuits

Chairman: *J. B. Angell, Stanford University, Stanford, Calif.*

"High-Order Harmonic Generation with Varactor Diodes," *F. P. Storke, Philco Corp., Palo Alto, Calif.*

"High-Power Operation of Varactor Devices," *R. P. Rafuse, MIT, Cambridge, Mass.*

"Design Techniques for Broadband Nondegenerate Parametric Amplifiers," *B. T. Vincent and K. M. Johnson, Texas Instruments Inc., Dallas, Tex.*

"Special Problems in Microwave Harmonic Generator Chains," *M. E. Hines, Microwave Associates Inc., Burlington, Mass.*

Wednesday Afternoon

Formal Opening of Conference

Introductory Comments, *J. J. Suran, General Electric Co.*, Chairman of Conference

Welcoming Remarks, *G. P. Harnwell, President, University of Pennsylvania.*

1961 Conference Awards, *T. R. Finch, Bell Telephone Labs., Inc.*

"The Principles and the Promise of Active Homogeneous Semiconductor Devices," *P. Aigrain, University of Paris, Paris, France, Defense Research for France.*

Session III—New Devices and Device Characterization

Chairman: *S. K. Ghandhi, Philco Corp., Blue Bell, Pa.,*

"Generalized Gain and Speed Limitations of Linear Amplifiers," *R. D. Thornton, MIT, Cambridge, Mass.*

"Ultrasonic Traveling-Wave Amplifier," *D. L. White, Bell Telephone Labs., Inc., Whippany, N. J.*

"Charge Definition of Transistor Properties," *A. R. Boothroyd, Imperial College, London, England.*

"Evaporated Circuits Incorporating a Thin-Film Transistor," *P. K. Weimer, RCA Labs., Princeton, N. J.*

"Applications of the Surface-Potential Controlled Transistor Tetrodes," *H. Z. Bogert, D. A. Tremere and C. T. Sah, Fairchild Semiconductor, Div., Fairchild Camera-Instrument Corp., Palo Alto, Calif.*

Session IV—Memory

Chairman: *R. H. Baker, MIT Lincoln Lab., Lexington, Mass.*

"Magnetoresistive Readout of Thin-Film Memories," *P. Huijer, Philips Research Labs., Eindhoven, Netherlands.*

"A High-Speed Thin Magnetic Film Memory," *W. E. Proebster and G. Kohn, IBM Corp., Zurich, Switzerland.*

"The Engineering of a Fast-Word Organized Tunnel-Diode Store," *A. Cole, G. B. B. Chaplin and P. M. Thompson, Plessey Co., Ltd., Hants, England.*

"A Magnetic Disk Serializer-Deserializer," *J. G. Simek, IBM Corp., Endicott, N. Y.*

"Improved MDRO Modes for Magnetic Film Memories," *K. Chu, IBM Corp., Yorktown Heights, N. Y.*

Wednesday Evening

Informal Discussion Sessions

"Distributed Logic"

Moderator: *E. E. Loebner, Hewlett-Packard Co., Palo Alto, Calif.*

Panel Members: *H. D. Block, Cornell University, Ithaca, N. Y.; M. Hirscher, RCA, Camden, N. J.; R. J. Lee, Adaptronics, Springfield, Va.; P. D. Low, Stanford University, Stanford, Calif.; E. E. Newhall, Bell Telephone Labs., Inc., Murray Hill, N. J.; R. D. Vernot, Philco Corp., Blue Bell, Pa.*

"The Interconnection Problem"

Moderator: *T. R. Finch, Bell Telephone Labs., Inc., Murray Hill, N. J.*

Panel Members: *E. Koenjian, American Bosh Arma Corp., Garden City, N. Y.; J. Kilby, Texas Instruments Inc., Dallas, Tex.; W. J. Means, Bell Telephone Labs., Inc., Murray Hill, N. J.; R. Rice, IBM Corp., Yorktown Heights, N. Y.; A. Coleman, RCA, Camden, N. J.; R. Gerhold, Hexagon, Signal Corps Lab., Ft. Monmouth, N. J.*

"Microwave Power Sources"

Moderator: *J. B. Angell, Stanford University, Stanford, Calif.*

Panel Members: *A. Uhlir, Jr., Microwave Associates Inc., Burlington, Mass.; W. B. Hauer, General Telephone and Electronics, Corp., Bayside, N. Y.; R. T. McCoy, Micromega Corp., Venice, Calif.; R. W. Roberts, Melabs, Inc., Palo Alto, Calif.; F. Sterzer, RCA Labs., Princeton, N. J.; F. Storke, Philco Corp., Palo Alto, Calif.*

"Magnetic Thin-Film Memories"

Moderator: *A. K. Rapp, Philco Corp., Blue Bell, Pa.*

Panel Members: *E. E. Bittman, Burroughs Corp., Paoli, Pa.; R. Petschauer, Remington Rand Univac, St. Paul, Minn.; J. I. Raffel, MIT Lincoln Labs., Lexington, Mass.; J. A. Rajchmann, RCA Labs., Princeton, N. J.*

"Noisemanship"

Moderator: *R. B. Adler, MIT, Cambridge, Mass.*

Panel Members: *H. A. Haus, MIT, Cambridge, Mass.; J. C. S. Kim, Electronics Lab., General Electric Co., Syracuse, N. Y.; A. van der Ziel, University of Minnesota, Minneapolis, Minn.; W. W. Mumford, Bell Telephone Labs., Inc., Whippany, N. J.*

"Charge-Control Characterization of Semiconductor Devices"

Moderator: *G. H. Goldstick, National Cash Register Co., Hawthorne, Calif.*

Panel Members: *J. L. Moll, Stanford Electronics Lab., Stanford University, Stanford, Calif.; J. J. Sparkes, British Telecommunications Research, Ltd., Berkshire, England; C. E. Simmons, Philco Corp., Lansdale, Pa.; A. R. Boothroyd, Imperial College, London, England; R. D. Lohman, RCA, Somerville, N. J.; C. L. Hegedius, IBM Corp., Poughkeepsie, N. Y.*

Thursday Morning, February 15

Session V—High-Speed Switching

Chairman: *V. H. Grinich, Fairchild Semiconductor, Div. Fairchild Camera and Instrument Corp., Palo Alto, Calif.*

"High-Speed Nonsaturating Switching Circuits Using a Novel Coupling Technique," *D. W. Murphy, IBM Corp., Poughkeepsie, N. Y.*

"Kilomegacycle Tunnel-Diode Logic Circuits," *B. E. Sear, Remington Rand Univac, Div. Sperry Rand Corp., Whitpain, Pa.*

"Computer Circuitry for 500 Mc," *W. Peil and R. Marolf, Electronics Lab., General Electric Co., Syracuse, N. Y.*

"Tunnel-Diode Balanced-Pair Switching Characteristics," *J. J. Gibson, G. B. Herzog, H. S. Miller and R. A. Powlus, RCA Labs., Princeton, N. J.*

Session VI—Low Noise Amplification

Chairman: *E. G. Nielson, Electronics Lab., General Electric Co., Syracuse, N. Y.*

"Tunable Millimeter Traveling-Wave Maser Operation," *F. Arams and B. Peyton, Airborne Instruments Lab., Deer Park, N. Y.*

"Packaged Tunable S-Band Traveling-Wave Maser Amplifier System," *S. Ikwit and J. G. Smith, Airborne Instruments Lab., Deer Park, N. Y.*

"An Extremely Low-Noise 6-Gc Nondegenerate Parametric Amplifier," *M. Ueno-*

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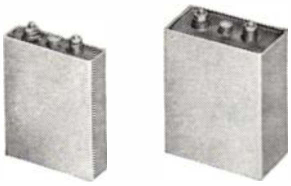
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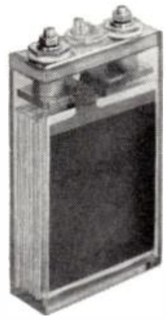
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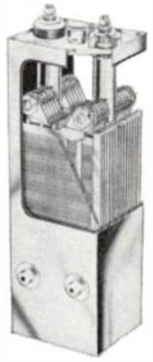
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- Microwave
- Engine Starting
- Lanterns

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

MARINE
- Engine Starting
- Radio and Alarms
- Life Boats

hara, Bell Telephone Labs., Inc., Murray Hill, N. J.

"Field-Effect Transistors as Low-Noise Amplifiers," *P. P. Lauritzen and O. Leistiko, Fairchild Semiconductor, Div. Farichild Camera and Instrument Corp., Palo Alto, Calif.*

"The Cryotron as a Small-Signal HF Amplifier," *N. D. Richards, Semiconductor Div., Mullard Ltd., London, England*

Thursday Afternoon

Session VII—Tunnel Diode Applications

Chairman: *A. E. Stern, The Martin Co., Baltimore, Md.*

"The Place of the Tunnel Diode in Solid-State Circuits," *P. M. Thompson, Plessey Co. Ltd., Hants, England.*

"A 4-mw, 6-kMc Tunnel-Diode Oscillator," *W. B. Hauer, General Telephone and Electronics Labs., Bayside, N. Y.*

"Tunnel Diode FM Transmitter for Medical Research," *W. Ko, W. Thompson and E. You, Case Institute of Technology, Cleveland, Ohio.*

"A Low-Noise X-Band Esaki-Diode Amplifier," *R. Trambarulo, Bell Telephone Labs., Inc., Holmdel, N. J.*

"Sensitive Tunnel-Diode Pressure Transducers," *M. E. Sikorski, Bell Telephone Labs., Inc., Murray Hill, N. J.*

Session VIII—Digital Transmission

Chairman: *J. S. Mayo, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"A Drift-Free Hyperbolic Encoder," *N. De Polo, Telettra S.P.A., Milan, Italy.*

"Carrier Pulses at Microwave and Millimeter-Wave Frequencies," *A. F. Dietrich and W. M. Sharpless, Bell Telephone Labs., Inc., Holmdel, N. J.*

"Variable-Width Fractional-Nanosecond Pulse Generators," *D. L. Berry, Electronics Lab., General Electric Co., Syracuse, N. Y.*

"The Semiconductor Delay Line and Related Devices," *J. J. Sein and S. N. Levine, RCA, New York, N. Y.*

Thursday Evening

Informal Discussions Sessions

"Nanosecond Circuitry"

Moderator: *E. P. Stabler, Electronics Lab., General Electric Co., Syracuse, N. Y.*

Panel Members: *B. J. Lechner, RCA, Camden, N. J.; W. M. Goodall, Bell Telephone Labs., Inc., Holmdel, N. J.; J. R. Turnbull, IBM Corp., Poughkeepsie, N. Y.; W. F. Chow, Remington Rand, Sperry Rand Corp., Philadelphia, Pa.; W. Peil, Electronics Lab., General Electric Co., Syracuse, N. Y.; C. N. Winningstad, Tektronix, Inc., Beaverton, Ore.; D. Berry, Electronics Lab., General Electric Co., Syracuse, N. Y.; M. M. Kaufman, RCA, Camden, N. J.*

"Optical Masers"

Moderator: *R. H. Kingston, MIT Lincoln Lab., Lexington, Mass.*

Panel Members: *P. A. Franken, University of Michigan, Ann Arbor, Mich.; C. G. B. Garrett, Bell Telephone Labs., Inc., Murray Hill, N. J.; H. Scharfman, Raytheon Co., Waltham, Mass.; M. L. Stich, Hughes Aircraft Co., Malibu, Calif.; A. E. Siegman, Stanford Electronics Lab., Stanford University, Stanford, Calif.*

"Thin-Film Magnetic and Supecronductive Techniques in Future Computers"

Moderator: *R. E. Hayes, Plessey Co., Ltd., Hants, England.*

Panel Members: *To be announced.*

"Low-Level Signal Processing"

Moderator: *D. Hilbiber, Fairchild Semiconductor, Div., Fairchild Camera and Instrument Corp., Palo Alto, Calif.*

Panel Members: *R. E. Middlebrook, California Institute of Technology, Pasadena, Calif.; G. S. Bahrs, Vidar Corp., Sunnyvale, Calif.; B. W. Fuller, Redcor Development Corp., Canoga Park, Calif.; J. A. Ekiss, Philco Corp., Lansdale, Pa.*

"High-Power, High-Speed Switching"

Moderator: *R. Bieselle, Shockley Transistor, Palo Alto, Calif.*

Panel Members: *H. Hurd, Stanford Radiation Lab., Stanford University, Stanford, Calif.; J. M. Goldey, Bell Telephone Labs., Inc., Murray Hill, N. J.; C. E. Smith, Texas Instruments Inc., Dallas, Tex.; C. Fisher, Signal Corps Lab., Ft. Monmouth, N. J.*

Friday Morning, February 16

Session IX—Functional Components

Chairman: *T. D. Stanley, RCA Labs., Princeton, N. J.*

"Photoconductive-Electroluminescent Display Panels on Fotoform Glass," *Z. Szepesi, Westinghouse Electric Corp., Elmira, N. Y.*

"Development of an Implantable Cardiac Pacemaker," *H. Raillard, Electronics Lab., General Electric Co., Syracuse, N. Y.*

"Circuit Parameters and Design in Thin-Film Technology," *F. F. Jenny, IBM Corp., Kingston, N. Y.*

"Components for Trainable Systems," *P. R. Low, IBM Corp., and Stanford University, Stanford, Calif. M. E. Hoff, H. S. Crafts and J. B. Angell, Stanford University, Stanford, Calif.*

"Germanium Photodiode Circuits Suitable for Measuring Extremely High Temperatures and Extremely Small Temperature Changes," *M. J. O. Strutt, Swiss Federal Institute of Technology, Zurich Switzerland.*

Session X—Design Applications

Chairman: *R. B. Adler, MIT, Cambridge, Mass.*

"Recent Advances in Wideband FM Receiver Design," *W. E. Ballentine, V. R. Saari, L. F. Willey, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"A 100-Watt Wideband Amplifier for the 1–12 Mc Frequency Range," *C. A. Franklin, Defense Research Telecommunications Establishment, Ottawa, Canada.*

"A Unique Silicon Controlled Rectifier High-Power Inverter With Sine Wave Output Voltage," *R. R. Ott, Electro-Mechanical Research Inc., Sarasota, Fla., L. A. Schlabach, Westinghouse Electric Corp., Pittsburgh, Pa.*

"A 300-kw Semiconductor Modulator," *F. A. Gateka and M. L. Embree, Bell Telephone Labs., Inc., Murray Hill, N. J.*

Professional Groups*

Aerospace & Navigational Electronics (G-11)—G. M. Kirkpatrick, Electronics Equipment and Systems Lab., GE Co., Syracuse, N. Y.; H. R. Mimno, Cruft Lab., Harvard Univ., Cambridge 38, Mass.

Antennas & Propagation (G-3)—Dr. H. Fine, Applied Propagation Branch, Technical Research Div., FCC, Washington, D. C.; S. A. Bowhill, Pennsylvania State Univ., University Park, Pa.

Audio (G-1)—C. M. Harris, Electronics Res. Labs., Columbia Univ., New York 27, N. Y.; M. Camras, Armour Res. Foundation, Tech. Ctr., Chicago 16, Ill.

Automatic Control (G-23)—J. M. Salzer, Ramo-Wooldridge, 5500 E. Segunda, Hawthorne, Calif.; G. S. Axelby, Westinghouse Air Arm Div., Friendship Airport, Baltimore 3, Md.

Bio-Medical Electronics (G-18)—G. N. Webb, Dept. of Medicine, Biophysical Div., Johns Hopkins Hospital, Baltimore 5, Md.; L. B. Lusted, Dept. of Radiology, Univ. of Rochester, Rochester 20, N. Y.

Broadcast & Television Receivers (G-8)—R. R. Thalner, Sylvania Home Electronics, 700 Ellicott St., Batavia, N. Y.; C. W. Sall, RCA, Princeton, N. J.

Broadcasting (G-2)—R. F. Guy, 264 Franklin St., Haworth, N. J.; W. L. Hughes, School of Elec. Engrg., Oklahoma State University, Stillwater, Okla.

Circuit Theory (G-4)—Dr. J. H. Mulligan, Jr., College of Engrg., New York, Univ., University Heights, New York 53, N. Y.; M. E. Van Valkenburg, Dept. of E.E., Univ. of Illinois, Urbana, Ill.

Communications Systems (G-19)—R. L.

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* Names listed are Group Chairmen and TRANSACTIONS Editors.

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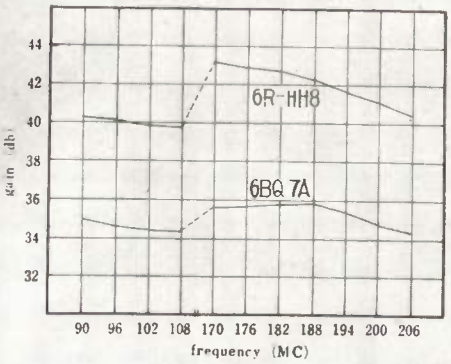
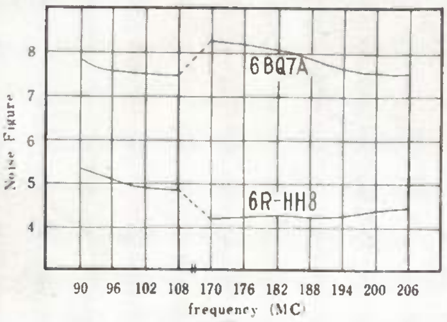
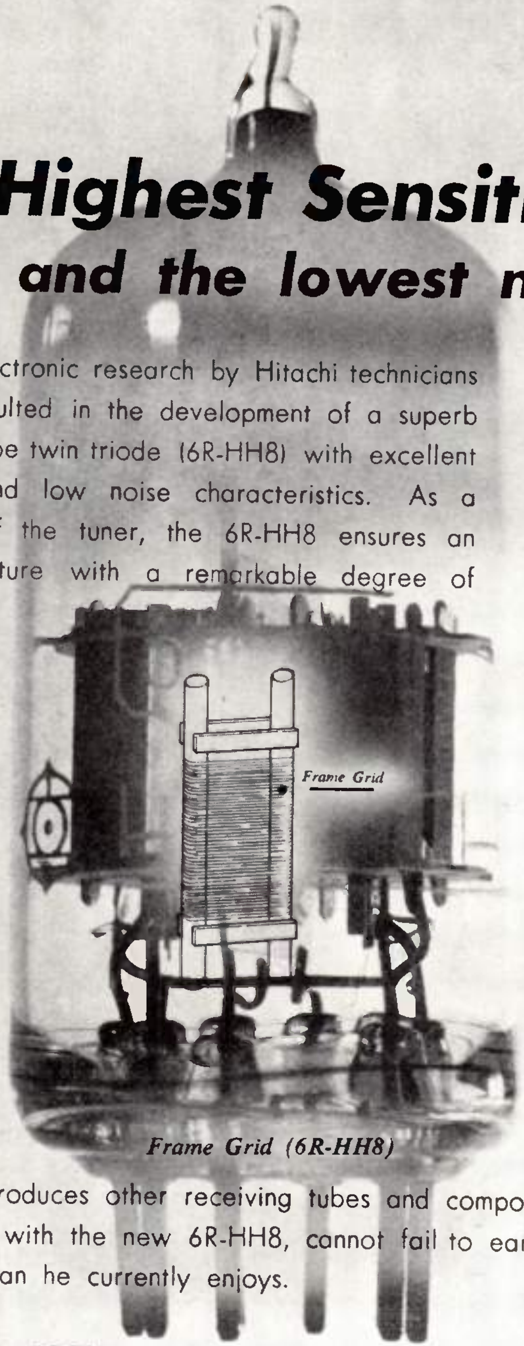


fig. 2 Noise characteristics





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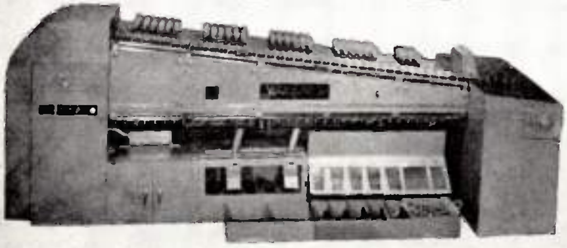

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Component Parts (G-21)—F. E. Wenger, Headquarters ARDC, Andrews AFB, Washington 25, D. C.; G. Shapiro, Engineering Electronics Sec., Div. 1.6, NBS, Connecticut Ave. and Van Ness St., Washington 25, D.C.

Education (G-25)—G. E. Moore, Westinghouse Elec. Corp., East Pittsburgh, Pa.; W. R. LePage, Dept. of E.E., Syracuse Univ., Syracuse 10, N. Y.

Electron Devices (G-15)—W. A. Adcock, Texas Instruments, 6000 Lemmon Ave., Dallas 9, Tex.; G. Wade, Spencer Lab., Raytheon Co., Burlington, Mass.

Electronic Computers (G-16)—A. A. Cohen, Remington Rand Univac, St. Paul 16, Minn.; Prof. N. R. Scott, Dept. of Elec. Engrg., University of Michigan, Ann Arbor, Mich.

Engineering Management (G-14)—T. W. Jarmie, Engineered Electronics Co., 1441 E. Chestnut Ave., Santa Ana, Calif.; Northwestern Univ., Evanston, Ill.

Engineering Writing and Speech (G-16)—J. M. Kimm, Jr., *IBM Journal*, 17th Floor, 545 Madison Ave., New York, N. Y.; H. B. Michaelson, IBM Res. Ctr., Box 218, Yorktown Heights, N. Y.

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Radio Frequency Interference (G-27)—H. E. Dinger, Naval Research Lab., Washington 25, D. C.; O. P. Schreiber, Technical Wire Prods., Inc., 129 Dermondy St., Cranford, N. J.

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* Numerals in parentheses following sections designate region number. First name designates Chairman; second name Secretary.

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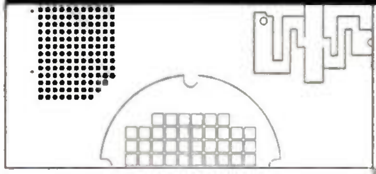

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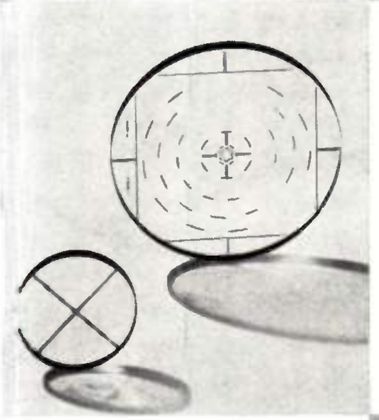
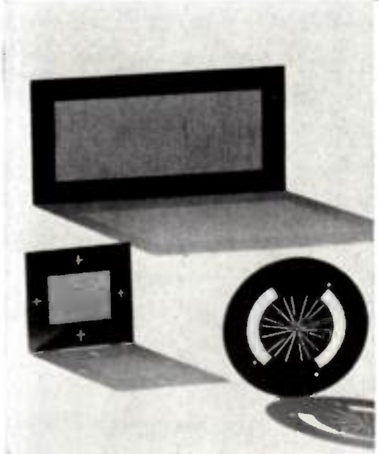


GLASS ETCHING DATA

Calibrated dials, straight and cross line rulings, scales, special purpose reticles and prisms suggest a few of the items produced. Glass may be coated or lines filled permanently with a choice of compounds.

DESIGN VERSATILITY	GLASS TYPES	LINEAR & ANGULAR TOLERANCES	LINE WIDTH TOLERANCES	FILLING MEDIA
Fire Control Reticles	Polished Ordinary Plate	Up to 3" ±.0001	.0005 to .001 ±.0001	Titanium Dioxide
Test Fixture Reticles	Water White Plate	Up to 6" ±.0005	.001 to .002 ±.0002	Ferric Oxide Red
Concave or Convex Reticles	Optical Crown	Up to 12" ±.002	.002 to .005 ±.0004	Ferric Oxide Black
Reticles for Guidance Systems Resolution Targets	Boro-Silicate Crown	Up to 24" ±.004	.005 to .010 ±.0005	Ferric Oxide Yellow
Calibrated Dials	Baryta Flint	Angles ±0° 0' 30"	.010 to .050 ±.001	Conductive Metallic Silver
Light Chopper Disks	Pyrex			



METAL ETCHING DATA

Etching does not change the molecular structure, leaves no burrs to remove. Round, square, oval or slotted holes may be etched. Patterns are etched through metal of these thicknesses: Copper to .030, nickel and alloys to .020, stainless steel to .025, molybdenum to .008. Surface etching on any thickness.

DESIGNS AND TYPES	METALS	DIMENSIONAL TOLERANCES
Fire Control Reticles	Copper and Copper Alloys	Linear to 3" ±.0002
Reticles for Optical Instruments	Nickel and Nickel Alloys	Linear to 12" ±.001
Scales	Cupro Nickel	Linear to 24" ±.003
Calibrated Dials	Spring Steel	Angular ±0° 0' 30"
Reticles for Guidance Systems	Stainless Steel	Line Widths to .002 ±.0001
Micro Component Parts for Control Systems	Aluminum	Line Widths to .005 ±.0005
Curved Surface Etched Parts	Cold Rolled Steel	Line Widths to .010 ±.001
Cylindrical Surface Etching	Molybdenum	Line Widths over .010 15% of Metal Thickness
Mechanical Parts Too Complicated for Die Stamping	Tungsten, Lead, Inconel	

ELECTRO FORMED MESH DATA

Evaporation masks for mesa transistors in miniature and subminiature sizes. Micro mesh sieves and screens are standard production. For all work a special ruling engine draws up to 2,000 lines per inch (straight or cross line, calibrated dials, scales and reticles), to linear and concentric circle accuracies of ±.000039.

MESH LINES PER INCH	SIZE in INCHES	MESH LINES PER INCH	SIZE in INCHES
1,000	4½ x 4½	120	4½ x 5
750	4 1/16 x 4 1/16	100	9 x 9⅜
500	5¼ x 5¼	100	6 x 6⅛
500	8 x 8	100	4½ x 4½
400	8 x 8	100 x 30	13⅝ x 16¼
300	8½ x 8½	80	11½ Dia
300	3¼ x 3¼	80	5⅛ x 5⅛
250	10 x 10	80	4 x 3⅞
250	9½ x 9½	50	4¾ x 4½
200	10 x 11	50	6 x 5⅞
200	5 x 5	40	5⅞ x 5⅞
200 x 60	2½ x 3¼	40	4⅛ x 4⅛
150 H.T.	14⅜ x 16⅝	20	4¾ x 4¾
120	6½ x 6⅜	5	6 x 5⅞
120	5 x 5		

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Lancaster (3)—Y. Uyeda, 1924 Pine Dr., Lancaster, Pa.; R. M. Matheson, 2728 Brookfield Rd., Lancaster, Pa.

Las Cruces-White Sands Proving Ground (6)—H. Coleman, Rte. 1, Box 4B, Las Cruces, N. M.; Secretary to be advised.

Lehigh Valley (3)—J. H. Volk, 411 Grant St., Easton, Pa.; A. I. Larky, Lehigh Univ., Dept. of Elec. Engrg., Bethlehem, Pa.

Memphis (3)—C. Ray, Dept. of Neurosurgery, Baptist Memorial Hospital, Memphis 3, Tenn.; Brother I. J. Haas, Christian Brothers College, Memphis 4, Tenn.

Merrimac Valley (1)—D. D. Sagaser, Bell Telephone Labs., 1600 Osgood St., North Andover, Mass.; W. Banton, 21 Walnut St., North Andover, Mass.

Mid-Hudson (2)—R. H. Domenico, IBM Research Lab., Poughkeepsie, N. Y.; W. Cadden, 67 Round Hill Rd., Poughkeepsie, N.Y.

Monmouth (2)—P. E. Griffith, 557 Cedar Ave., W. Long Branch, N. J.; V. E Reilly, 25 Lenox Rd., Eatontown, N. J.

Nashville (3)—G. P. McAllister, 2923 Twin Lawn Dr., Nashville 14, Tenn.; W. B. Kincaid, Jr., 210 Graeme Dr., Nashville 14, Tenn.

New Hampshire (1)—R. O. Goodwin, 130 Colgate Rd., Nashua, N. H.; J. Butler, Groton Rd., R.F.D. 2, Nashua, N. H.

Northern Vermont (1)—R. A. Marcotte, Valleyview Dr., RD 1, Essex Junction, Vt.; Secretary to be advised.

Orange Belt (7)—J. F. McElwee, 455 N. Live Oak Ave., Glendora, Calif.; A. G. Holtum, Jr., 941 E. Marylind, Claremont, Calif.

Palm Beach (3)—C. E. Cronin, 544 Ebb Tide Dr., North Palm Beach Fla.; A. S. Baran, 301 Bayside Rd., Lake Worth, Fla.

Panama City (3)—J. F. Ault, 1305 Cornell Dr., Panama City, Fla.; R. C. Lowry, 2342 Pretty Bayou Dr., Panama City, Fla.

Pasadena (7)—F. L. Mosely, 700 Fleutridge, Pasadena 5, Calif.; J. C. Crosby, 9018 Youngdale, San Gabriel, Calif.

Pikes Peak (6)—B. J. Bittner, Box 3217, North End Station, Colorado Springs, Colo.; A. O. Behnke, 204 Westcott Ave., Colorado Springs, Colo.

Reading (3)—H. S. Hauck, 216 Jameson Pl., Reading, Pa.; R. M. LeLacheur, 1637 Dauphin Ave., Wyomissing, Pa.

Richland (7)—R. A. Harvey, 657 Cottonwood Dr., Richland, Wash.; D. D. Bisby, 703 Catskill, Richland, Wash.

San Fernando Valley (7)—R. L. Halpern, 3315 Longridge Ave., Sherman Oaks, Calif.; J. L. Brown, 17408 Mayflower Dr., Granada Hills, Calif.

Santa Ana (7)—J. E. Elms, 112 Kings Pl., Newport Beach, Calif.; K. Goodman, 2567 Columbia Dr., Costa Mesa, Calif.

Santa Barbara (7)—J. E. Hacke, Box 535, Santa Barbara, Calif.; S. R. Boyle, Defense Electronics Div., General Electric Co., 735 State St., Santa Barbara, Calif.

Southwestern Ontario (8)—R. A. Clark, 2558 Byng Rd., Windsor, Ont., Canada; Secretary to be advised.

Victoria (8)—C. L. Madill, 2786 Murray Dr., Victoria, B. C., Canada; A. M. Baxter, 620 Rockland Place, Victoria, B. C., Canada.

Westchester County (2)—M. Ziserman, 26 Edgewood Rd., Hartsdale, N. Y.; M. L. Brailey, 12 Dupont Ave., White Plains, N. Y.

Western North Carolina (3)—J. I. Barron, Southern Bell T&T Co., Box 240, Charlotte, N. C.; T. C. Livingston, 926 N. Sharon Amity Rd., Charlotte, N. C.

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Typical Noise Factor 8·5 dB

With voltages fixed for broad band operation 2·5 to 4·1 Gc/s

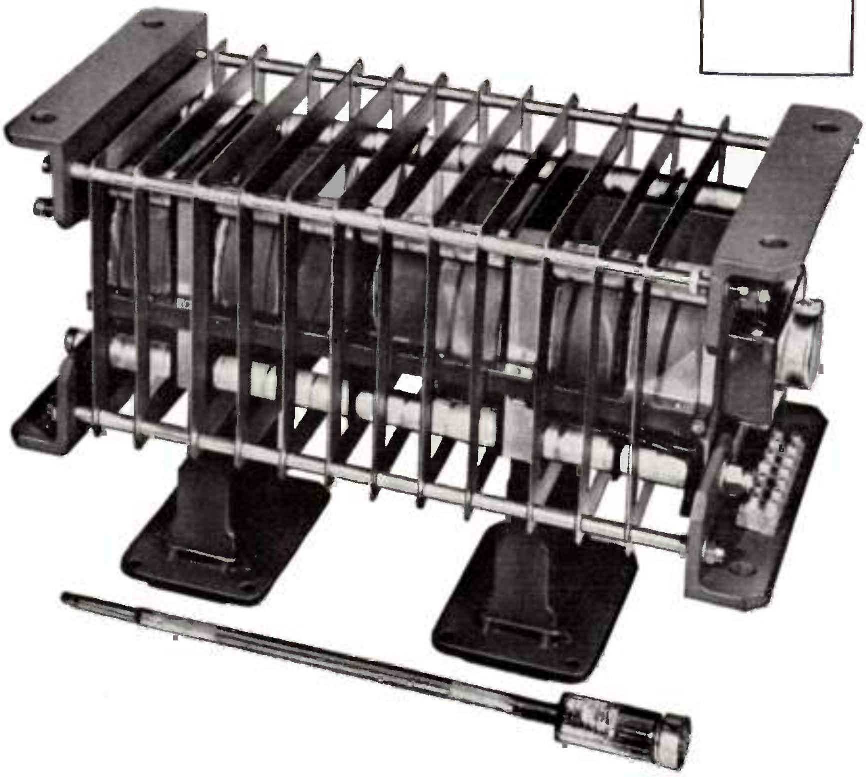
Typical Gain 40 dB

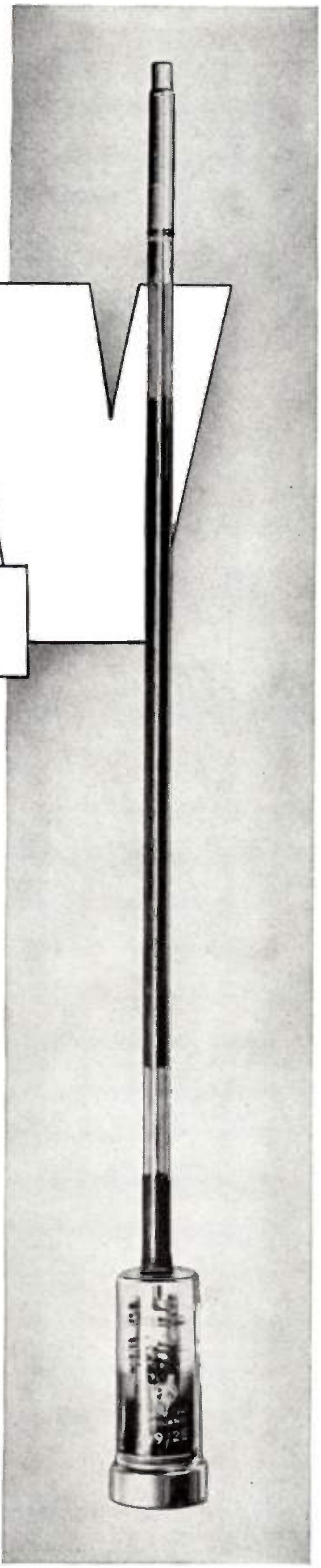
W10/3E

Typical Noise Factor 6·8 dB

With voltages optimised for a spot frequency between 2·7 and 3·3 Gc/s (circuit 495—LVA—003) (illustrated with antimagnetic screening) or between 2·8 and 3·8 Gc/s (circuit 495—LVA—006)

Typical Gain 23 dB





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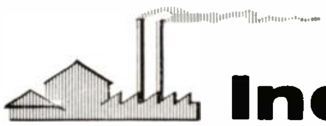

Industrial Engineering Notes[*]

Association Activities

A highlight of the gathering, as in every year since 1941, was presentation of the Radio Fall Meeting Plaque, signifying distinguished achievement for scientific and technological advancement in entertainment electronics. The 1961 recipient of the honor was announced at the meeting dinner as William T. Wintringham, head of the Visual Systems Research Department of the Bell Telephone Laboratories at Murray Hill, N. J. Presentation of the plaque to Mr. Wintringham for his "many technical contributions to the electronics industry in the field of video techniques" was made by Virgil M. Graham, Associate Director of the EIA Engineering Department.

Governmental and Legislative

The Federal Communications Commission has granted a 60-day extension of the December 4 deadline for filing of comments on overall policy and technical proposals for expansion of UHF television. The proceeding is contained in Docket 14229, announced July 26. The extension does not apply to separate proceedings involving deintermixture proposals which would convert eight television markets to UHF and provide shorter-spaced VHF assignments in eight other markets.

Military and Space

A series of tests slated for next year may well solve technical problems involving the use of nuclear engines in rockets, Harold B. Finger, Manager of AEC-NASA Space Nuclear Propulsion Office, said last week. If this breakthrough is achieved, the Project Rover manager said, the way will be clear for the use of nuclear engines in "early" manned lunar operations. The upcoming test series involves the properties of materials used in various reactor designs and control of the designs. The outcome will provide a reactor for the NERVA nuclear rocket engine. Mr. Finger said early flight depends on successful development of the NERVA engine and a flight-test program called RIFT, a program that will lead to nuclear stages which can be used as upper stages of chemical-boost rockets now being considered for carrying an astronaut to the moon.

Industry Marketing Data

September was the best month of the year for production of television and radio receivers alike with totals exceeding August output by 179,906 TV sets and 663,597 radios, it was disclosed in monthly figures released last week by the EIA Marketing Data Department. Output of TV sets in September totaled 694,580, compared with 514,674 in August. TV production for the nine months of this year jumped ahead of the total for the same period last year by 20,684 sets to a high of 4,393,768. Cumulative sales through August exceeded last year's by only about 5000. A total of 2,048,698 radios were sold at the factory in September, against 1,385,101 sold the month before. Cumulative sales through September were below those for the nine-month period in 1960 by 663,597 sets. The totals were 12,001,488 this year, compared with 12,408,377 in 1960. . . . **The EIA Military Marketing Data Committee has published two new marketing aids, one an analysis of the Atomic Energy Commission's fiscal year 1962 budget and the other a summary of the proposed National Meteorological Satellites program.** The analysis of AEC funds, based on the report to Congress by the Joint Committee on Atomic Energy, places special emphasis on electronic content. It breaks down $2,359,400,000 funded for operating expenses and $268,750,000 earmarked for plant acquisition and construction. The meteorological satellites plan was developed by a special panel on Operational Meteorological Satellites, a part of the National Coordinating Committee for Aviation Meteorology, and endorsed by the participating agencies—the National Aeronautics and Space Administration, Weather Bureau, Defense Department, and Federal Aviation Agency. The summary contains a five-year projection of funding requirements. The FY 1962 portion has been approved. A limited number of copies of both documents is available to EIA member-companies from Office of Information, EIA Headquarters, 1721 DeSales Street, N.W., Washington 6, D.C.

Engineering

A wide array of scientific and technical information in libraries and information centers scattered across the country would be made available to researchers faster under an Air Force proposal to establish a "materials information processing capability." The proposed super library is an outgrowth of a modern paradox facing scientific communities everywhere: Scientists and engineers want more reliable information and want it faster, yet do not have enough time to read and evaluate all the information they presently receive. The new information processing capability would ease this dilemma by using computers to sift descriptive information and

(Continued on page 72A)

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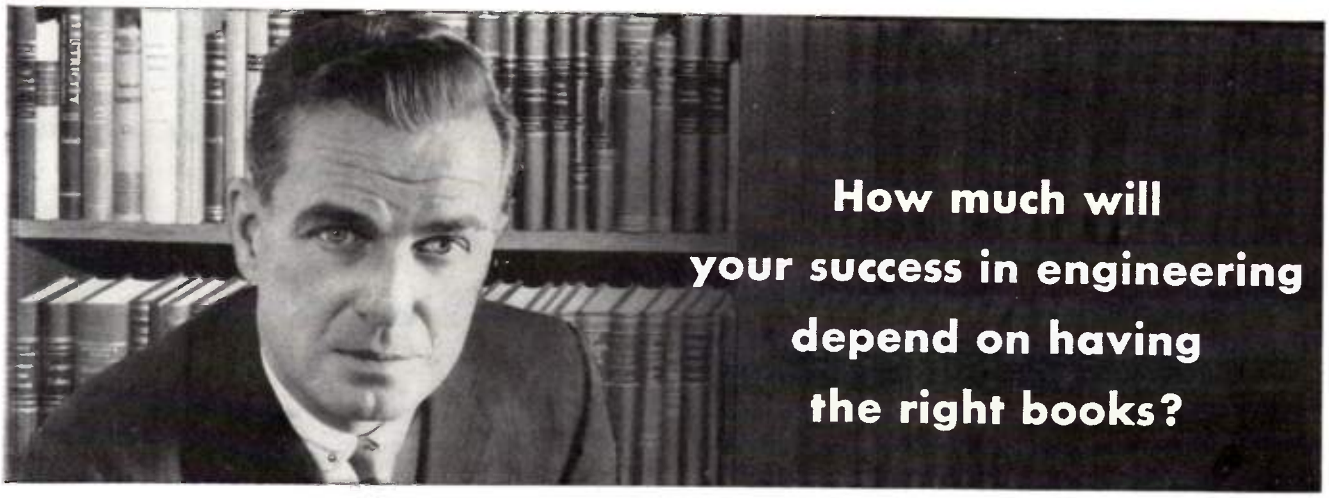
[*] The data on which these Notes are based were selected by permission from *Weekly Report*, issues of November 6, November 13, and November 20, 1961, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.



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ELECTRONIC DESIGNERS' HANDBOOK

By R. W. Landee, D. C. Davis, and A. P. Albrecht

Provides fundamentals and data to help you in the design of all types of electronic equipment. A large number of the circuits used in many different applications are covered, together with theoretical and technical discussions and explanations, design examples to show application of theory, and graphical and tabular data needed in day-to-day design work.

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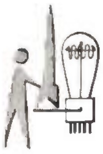
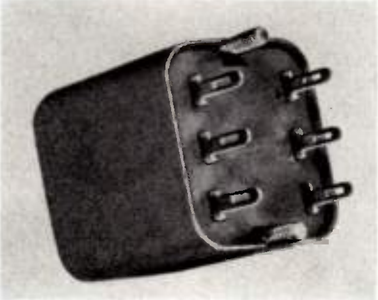


NEWS New Products

IRE

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

Wideband Transformer



North Hills Electronics, Inc., Alexander Pl., Glen Cove, L. I., N. Y., has announced a new wideband transformer. Designated as the type 1900, this transformer provides idolated coupling between 75Ω unbalanced, to 1200Ω balanced, over a frequency range of 200 kc to 50 mc. The insertion loss is less than 1 db. The secondary balance is better than 40 db.

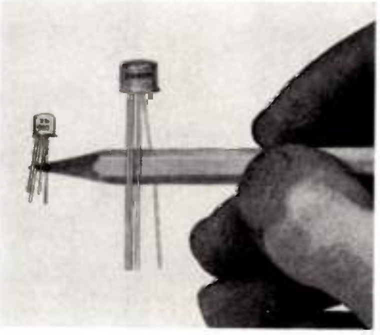
Three case styles are offered. Series AA models are potted in metal cases with a terminal board on the bottom and two 6-32 spade mounting bolts. Series BA are hermetically sealed with glass terminals at the case ends, and two 4-40 studs on the bottom for mounting. Series BB is similar to BA, except that the unbalanced connection is provided by a BNC connector.

A great variety of different requirements is being met by special designs. Your specifications for custom wideband transformers may be submitted for quotation.

Complete data may be obtained from the firm.

Universal Communications Transistors

A low cost, universal communications transistor, the 2N987, has been announced by **Amperex Electronic Corp., Semiconductor and Special Purpose Tube Div.,** 230 Duffy Ave., Hicksville, L. I., N. Y. It is mounted in a subminiature 4-pin, TO-18 case. The same transistor, packaged in a TO-33 case, is available as the type 2N2084.



PADT Germanium alloy-mesa transistors, the 2N987 and the 2N2084 provide the best characteristics of RF front and IF types, and feature the ideal combination of high breakdown voltage, high beta and high frequency. As such, they are suited for use in RF and IF amplifiers in the HF and VHF bands in mobile and airborne communications, instrumentation and radar IF applications.

The subminiature 2N987 is particularly suited for portable paging systems, airborne, mobile communications and other miniture equipment where size, weight and space are important.

The 2N987 is priced at $1.55 each and the 2N2084 at $1.25 each in 1000 lots.

These components provide a typical power gain of 14 db at 100 mc, typical beta (h_{fe}) of 140, breakdown voltage (BV_{cbo}) of 40 V, and an output capacitance (C_{ob}) of 2 μμf.

Both types are suitable for printed circuits and will replace such types as the 2N499, 2N1224, 2N1395, 2N1396, 2N1745 and 2N1866.

Test Procedures Reference Manual for Coated Magnet Wire

A new manual of test procedures for all film coated magnet wire types is available on request from **Hudson Wire Co.** The reference manual is a compilation and summary of all the different NEMA and MIL-W-583B specification testing techniques that apply to these products. While all of the test procedures described are covered in part by various individual specifications, this is the first time that they have been published in a single source.

The manual does not contain data on test results that must be obtained for the different types of film coated magnet wire to be judged acceptable. Because these results vary considerably for each different type, they will be covered in individual technical bulletins on each product that will be published over the next few months.

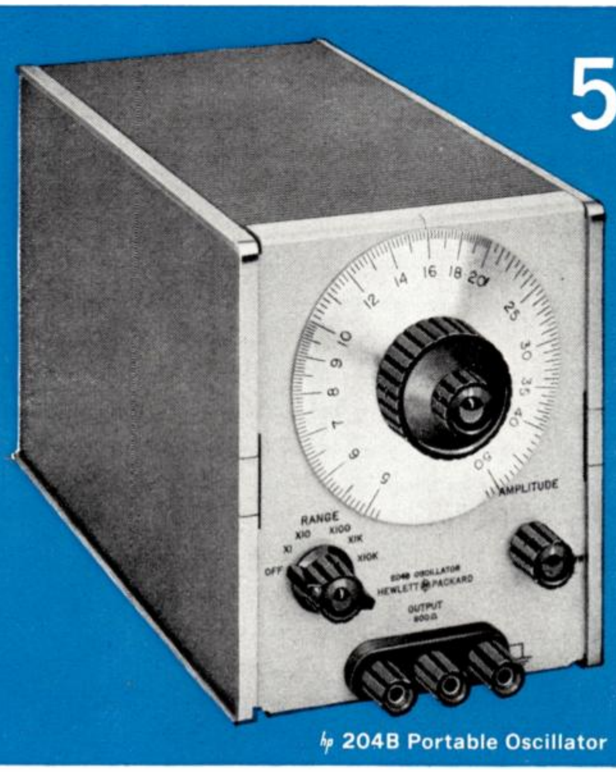
Hudson's magnet wire divisions are established suppliers of magnet wire insulated with plain enamel, liquid nylon, polyurethane, formvar, isonel, silicone, ML and various dual coating. They also produce fabric and film/fabric insulated magnet wires.

Copies of the manual are available free from **Hudson Wire Co., Magnet Wire Division,** at either Winsted, Conn. or Cassopolis, Mich.

(Continued on page 42A)

Use it on the bench—carry it anywhere!

New hp High Stability Oscillator



hp 204B Portable Oscillator

5 cps to 500 KC

Unique frequency, amplitude stability in this compact, light weight, solid state oscillator!

The unusual stability of the new hp 204B combines with solid state design and battery operation to give you one of the most useful, reliable, versatile oscillators available today. Because the 204B is fully transistorized, internal heat is small and warmup drift is negligible. In addition to battery operation, the 204B is operable on ac, with an ac power pack available at extra cost.

Frequency stability over the entire 5 cps to 500 KC range is better than ±0.03%/°C from 0° to 55°C. Amplitude stability over rated frequency range and output levels is better than ±0.1% over 8 hours of operation at constant line voltage and temperature*; better than ±0.2% for line voltage changes of ±10%; better than ±0.1%/°C, 0-55°C.

Output of the 204B is fully floating, isolated from both power line ground and chassis. Balanced and unbalanced loads, and loads referenced either above or below ground, can be driven by this versatile oscillator. Low impedance circuits drive the 600 ohm output, effectively isolating the oscillator stage.

SPECIFICATIONS

Frequency Range:	5 cps to 500 KC, 5 ranges, 5% overlap, vernier control
Dial Accuracy:	± 3%
Frequency Response:	± 3% with rated load
Output Impedance:	600 ohms
Output:	10 mw (2.5 v rms) into 600 ohms; 5 v rms open circuit
Output Control:	Continuously variable bridged "T" attenuator with at least 40 db range
Distortion:	Less than 1%
Noise:	Less than 0.05%
Power Source:	4 battery cells at 6.75 v each, 7 ma drain, life at least 300 hours
Power Accessory Available:	AC power supply can be installed in place of batteries. Optional at extra cost.
Dimensions:	6-3/32" x 5-1/8" x 8". 6 lbs.
Price:	hp 204B, $275.00

7063

Designed in the new hp instrument module packaging, the 204B is only 6-3/32" high, 5⅛" wide and 8" deep; weighs just 6 pounds! A new rack mount adapter holds three 204B oscillators or other hp instruments of the new modular design.

---

*On battery operation, stability −1% during battery depolarization, less than 30 minutes.

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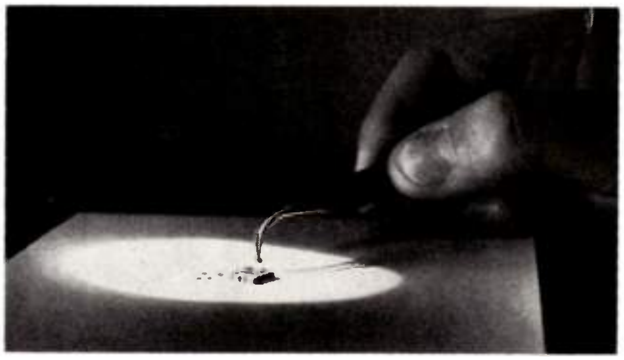
NEWS New Products

(Continued from page 40A)

Electromechanical Gyrator

Piezoelectric and piezomagnetic transducers have been combined to form a device—low frequency gyrator—which violates the reciprocity principle, according to an announcement by **Clevite Electronic Research Center** personnel at the 17th Annual National Electronics Conference.



In a technical paper authored by J. H. Silverman, J. D. Schoeffler and D. R. Curran it was divulged that as an antireciprocal device, the electromechanical (EM) gyrator transforms voltage into current, an inductor into a capacitor and a short circuit into an open circuit.

One important potential application of the experimental device would utilize a gyrator in conjunction with a neutralizing circuit to transmit signals in one direction only, in the form of an isolator.

Possible use is in telecommunications to eliminate cross-talk.

According to Clevite Corporation's Electronic Research Center, Cleveland, Ohio the ideal gyrator, like the resistor, inductor, capacitor and ideal transformer, is a separate independent network element. The current flowing through one pair of a gyrator's terminals is directly related to the voltage existing across the other pair of terminals.

Consequently, the impedance seen at the input of the gyrator is related to the admittance of the load, network elements transformed through the gyrator must be represented by their reciprocal duals, and transformers cascaded with the gyrator completely lose their identities.

A distinguishing feature of the gyrator is its antireciprocal network behavior, characterized by transfer parameters which are equal in magnitude but opposite in sign (a 180° phase displacement) for the two directions of signal transmission. Similarly, any network combination consisting of a reciprocal system cascaded with a gyrator exhibits antireciprocal behavior.

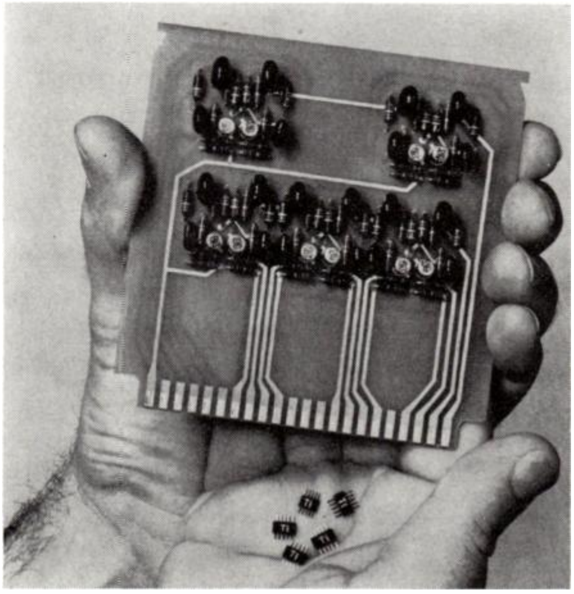
Networks placed in series or in parallel with an antireciprocal system result in a nonreciprocal combination. The transmission characteristics in the two directions will then differ in magnitude as well as phase. By proper choice of the auxiliary networks it is possible to produce an isolator, that is transmission in one direction only.

Clevite Electronic Research Center developed the experimental EM gyrator under a U. S. Air Force contract. It combines lead titanate zirconate piezoelectric ceramic and nickel cobalt magnetostrictive ferrite (with permanent magnet biasing) transducers bonded together with unity mechanical coupling to form a single resonant structure.

Two types of construction—composite beam and butt joint—suffice to cover the frequency range from 1 to 1000 kc. Typical EM gyrators have band widths, at the three db point, of one to three per cent and minimum power insertion losses of one to three db.

Solid Circuit Semiconductor Networks

A new line of Solid Circuit® semiconductor networks—Series 51—was introduced by **Texas Instruments Incorporated,** 13500 N. Central Expy., Dallas, Texas.



The Series 51 line includes six different digital circuit modules processed with the most modern of semiconductor production techniques and embodying the most advanced concepts of circuit design. All are in production and immediately available. The devices will handle 90% of the circuit functions of digital equipments, particularly for military computers, programmers, and other information processing application functions.

Each silicon network is contained in a hermetically sealed package measuring $\frac{1}{4}$ by $\frac{1}{8}$ by $\frac{1}{32}$ of an inch.

TI Series 51 networks perform flip-flop, counter, NOR gate, NAND gate, and exclusive OR gate functions.

Series 51 networks drain but $\frac{1}{10}$ as much power as many ad-

(*Continued on page 45A*)

TODAY'S DEFENSE SYSTEMS:

Worth more & why

In today's defense arsenal, things cost more. They cost more
because they are worth more. They do jobs undreamed
of a few years ago—with reliability never believed possible.
Jet fighters are flying higher and faster.
New atomic-powered ships carry enormous strike power.
Missiles are thrusting farther with greater accuracy.
And new electronic systems, like the Hughes Tactical Attack
System, are making "impossible" control jobs a daily routine.

Hughes systems, developed to meet the most advanced fighter mission problems, have paced airborne systems development. They have flown with the F-86, the F-89, the F-102, the F-106, and other U. S. Air Force, U. S. Navy and Canadian aircraft.

The new Hughes Tactical Attack System in its own right achieves new levels of capability, reliability and over-all economy. Highly flexible in design, this new system will be as modern in the late 1960's as it is today. It integrates three major functions—high resolution radar, weapon control auxiliaries and a navigational subsystem. Each function incorporates the latest state-of-the-art developments.

Coolant Problems which could be encountered in aircraft flying at multiples of sonic speeds and at high altitudes have been solved. New techniques are being applied which provide for more efficient cooling of the thousands of electronic parts and assemblies, promising greater reliability.

New "encapsulation" packaging techniques permit more work to be done in smaller space, saving weight and, again, vastly increasing reliability. This technique packages complex circuits in small "blocks" of plastic material. Each contains scores of diodes, transistors, relays and other electronic components. These units are vibration-free. They can be plugged in and out as easily as a toaster—facilitating trouble-shooting and cutting maintenance time.

These kinds of advancements are the "norm" at Hughes. For they are the result of over 16 years of prime experience in the design and building of airborne systems. The new Hughes Tactical Attack System has grown out of a special, company-funded program which has been active for three years.

Creating a new world with electronics

HUGHES

HUGHES AIRCRAFT COMPANY

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

(Continued from page 43A)

vanced types of conventional circuits. Integrated fabrication techniques are utilized to minimize internal connections and increase potential reliability. Produced to meet the requirements of military applications, they achieve design compatibility with present-day digital circuitry, and operate over a temperature range of −55 to +125°C (−67 to +257°C).

The development of Solid Circuit® semiconductor networks was announced by Texas Instruments early in 1959. At the beginning of 1960, a catalog unit was made available for evaluation by the industry. In addition to these products, an extensive custom design program is available.

Advanced semiconductor integrated design and fabrication technique are used in producing its networks. All of the units in the Series 51 are derived from one standard silicon wafer. After all electrical paths have been formed in the slice, it is then designated for one of the six Series 51 circuits, and appropriate connections are made. The standardization of fabrication steps for this line will allow an effective manufacturing volume to be quickly realized with its attendant economies. Each silicon circuit is protected with oxide coating and deposited leads are used for point-to-point connections on the silicon bar. The completed silicon bars then are mounted in an hermetically sealed package.

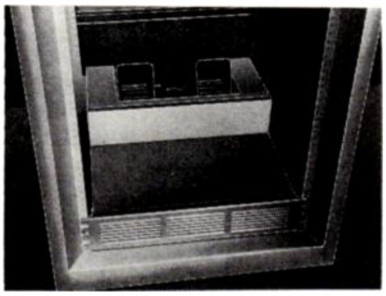
The new semiconductor network circuit line includes types SN 510 and SN 511 flip-flop/counter circuits, types SN 513 NOR/NAND gates, types SN 514 two independent NOR/NAND gates in a single package and type SN 515 exclusive OR gate network. Both the SN 511 and SN 513 have emitter follower outputs in addition to the regular outputs to allow for greater driving capability to following logic stages. The SN 510 flip-flop and SN 512–514 NOR/NAND gates have power drain of 2 mw, while the power drain of the emitter follower units, SN 511 and 513, is about mw for Vcc = 3 volts.

Units of Series 51 sell in the $50–$65 range for quantities of 1000. For sample quantities, SN 510, SN 512, SN 514 and SN 515 sell for $95 each in quantities of 1 to 25 and for $76 each in quantities of 26 and up. The SN 511 and SN 513 sell for $115 each for 1 to 25 units and $92 each for 26 and up.

Blower

A new multi-purpose all-angle blower from **Western Devices, Inc.,** 600 W. Florence Ave., Inglewood 1, Calif., requires 3½″ of panel space, and is claimed to be so versatile that one unit in stock can eliminate procurement of several single-purpose models to satisfy variable requirements.

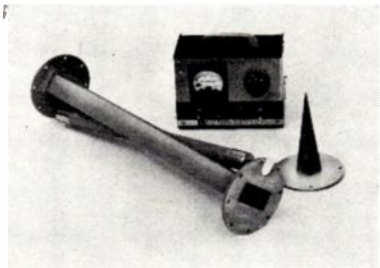
May be used for supply or exhaust, or one port for supply, one for exhaust. Twin



scrolls can be rotated and set to any angle through 230° for accurate flow control, or the motor-rotor assembly may be reversed end-for-end in its housing to provide new, equally diverse flow patterns. Blower assembly is recessed to allow for extra usable chassis or storage space. MIL quality heavy duty construction. Easy maintenance without removal from cabinet. Cushion mounted for quiet operation. Motor bearings are permanently lubricated. Cleanable or disposable filter. Contact the firm directly for details.

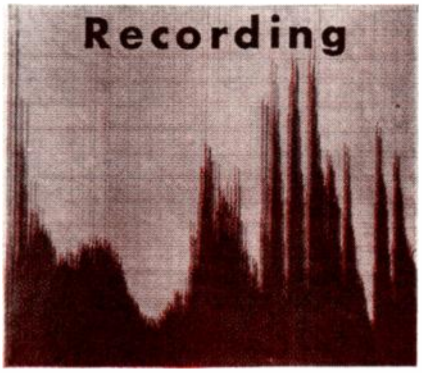
Random Noise Generators

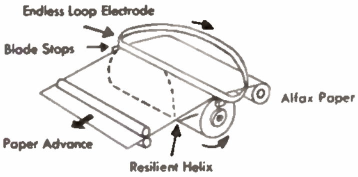
For use in the laboratory or in field systems requiring the measurement of noise figure and receiver sensitivity, Model 501 Random Noise Generators manufactured by **General Microwave Corp.,** 47 Gazza Blvd., Farmingdale, L. I., N. Y., are available in six waveguide sizes covering the frequency range from 2.6 to 18.0 Gc. All models in the series feature the use of argon gas discharge tubes whose output is independent of ambient temperature. Each of the waveguide models in the Model 501 Series Random Noise Generators is supplied with a matched termination (VSWR less than 1.03) as standard equipment, so designed that its inclusion does not add to the normal length of the waveguide mount. The matched termination is readily removable, leaving available a standard waveguide flange for connection to other waveguide components or signal sources.

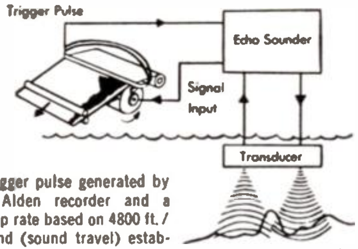


The nominal excess noise power ratio, which varies from 15.2 to 16.0 db, depending upon waveguide size, is constant over the waveguide operating frequency range and uniform from instrument to instrument within ±0.2 db. Each of the waveguide noise sources in the 501 series is designed to operate with Model 301 Universal Power Supply which contains provisions for controlling and monitoring the lamp current. Price range for the Model 501 Noise Generator—$175.00 to $250.00 each (depending upon waveguide size), Model 301 Power Supply—$125.00. Delivery is from stock.

(Continued on page 111A)







IRE People

Appointment of **Richard W. Armil** (S'52–A'50) as Manager of Sales Engineering of Rochester Instrument Systems, Inc. was announced recently. He will direct all sales activities of the company which manufactures annunciators, electronic trips, process control systems and nuclear reactor instrumentation. Previously, he was Regional Sales Manager for General Dynamics/Electronics, Rochester, N. Y. He was also formerly Regional Sales Manager for Hathaway Instruments, Inc., Denver, Colo.


R. W. Armil

Mr. Armil is a member of the American Institute of Electrical Engineers and is a Registered Professional Engineer.

◆

The appointment of **R. L. Biesele, Jr.** (SM'58) as Manager of Operations for the Shockley Transistor unit of Clevite Transistor in Palo Alto, Calif. has been announced. He joined Shockley in 1958 as Manager of Research and Development. With the Shockley unit, he has directed a large R & D staff in the development of advanced semi-conductor devices and in the conduct of research programs for the armed forces and industry. For the present he will continue his work as manager of R & D in addition to his new duties.


R. L. Biesele

In 1936 he graduated from the University of Texas with the B.S. degree in electrical engineering. For the first seven years after graduation he worked with Westinghouse Electric as a lighting application engineer and field lighting specialist. In 1943 he was appointed assistant professor of electrical engineering at Southern Methodist University. Two years later, he achieved the rank of full professor and was made chairman of the department, a position he held until 1952. In 1952, he became vice-president and chief engineer of Toff Corporation of Dallas, Tex. At Toff, he completed the development of a photoelectric grading instrument. After completing his work at Toff, Biesele joined the faculty of the University of Oklahoma. He was director of the digital computer laboratory during his last year at the University prior to joining Fenwal, Inc., Ashland, Mass. At Fenwal, as chief electronics engineer, he participated in advanced work on temperature measurement, monitoring, and control equipment for airborne and industrial applications.

Mr. Biesele is a member of Tau Beta Pi, the AIEE and the National Society of Professional Engineers.

◆

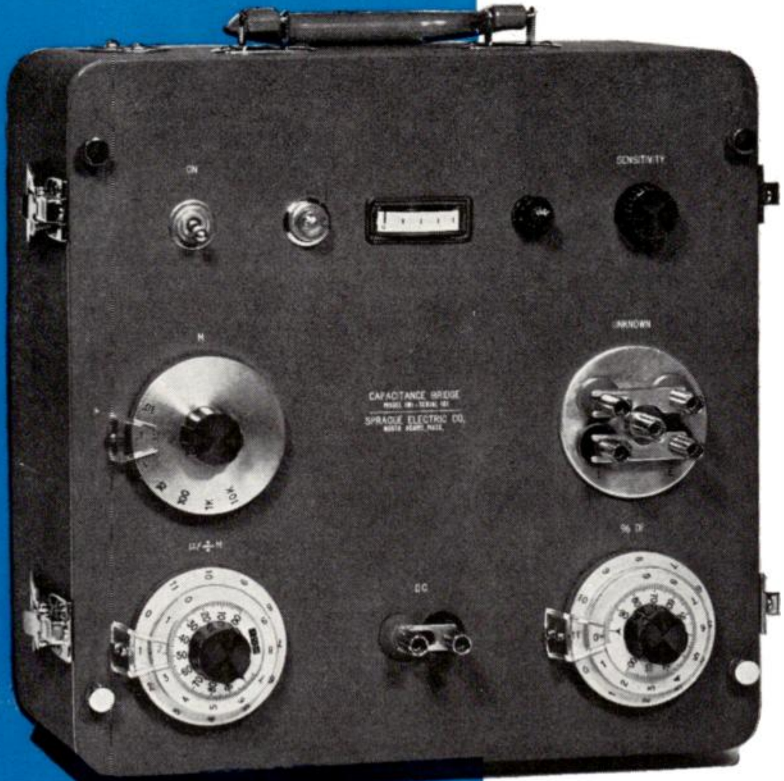
Dr. Leo L. Beranek (S'36–A'41–SM'45–F'52), President of Bolt, Beranek and Newman, Inc., Cambridge, Mass., received the Wallace Clement Sabine Award of the Acoustical Society of America on November 10, at the 62nd Meeting of the Acoustical Society in the Netherland Hilton Hotel, Cincinnati, Ohio. He received the award for service in the field of architectural acoustics as scientist, teacher and supervisor of student research, author of outstanding books and consultant to architects.

Recipient of the Doctor of Science degree in physics from Harvard University, Cambridge, Mass., he served as Director of the Harvard Electro-Acoustic Laboratory from 1940 to 1946 and organized studies on airborne acoustics during World War II. He also formed the Systems Research Laboratory at Harvard during the war to combat the threat from Japanese Kamikaze planes. From 1947 to 1954 he was Technical Director of the Acoustics Laboratory at the Massachusetts Institute of Technology. He was an Instructor and Assistant Professor at Harvard, an Associate Professor at M.I.T. where he is now a Lecturer, and a Visiting Professor at the University of Buenos Aires. He has lectured on acoustics in London, Zurich, Prague, Warsaw, Moscow, Finland, and Switzerland, and for several summers directed a special program on noise reduction at M.I.T. He helped to form the IRE Professional Group on Audio and became its first chairman. He has served on committees of the National Academy of Sciences-National Research Council, and the Department of Defense. He has served also as President of A.S.A., helped organize and edit its magazine "Noise Control," has served on the editorial board of its journal, and has contributed widely to the society's program on standardization and technical development. His work in acoustics includes authorship of five books including "Acoustics," "Acoustic Measurements," and "Noise Reduction." A fifth volume, "Acoustics of Halls of Music," will be published in 1962.

Dr. Beranek is a Fellow of the A.S.A., the American Academy of Arts and Sciences, the New York Academy of Sciences, American Physical Society, and the American Association for the Advance-

(Continued on page 48A)

Now! *For the first time...*



TEST ELECTROLYTIC CAPACITORS

Safely...
Accurately...
Simply!

SPECIFICATIONS

Capacitance
Range: 0 to 120,000μF at 120 cps
Accuracy: ±(1% of reading +10μμF)
Sensitivity: ±(0.1% of reading +10μμF)
Dissipation Factor
Range: 0 to 120% at 120 cps
Accuracy: ±(2% of reading +0.1% DF)
Sensitivity: ±(0.2% of reading +0.05% DF)
Maximum Voltage to Unknown
A-C: 0.5v RMS at 120 cps
D-C: 0-600v (external)
Null Detection
Built-in Galvanometer to Indicate Bridge Balance
Power Input
105-125v, 60 cps, 15w
Case
Sturdy Aluminum Cabinet with Blue Textured Finish, Grey Panel
Dimensions
12" Wide x 12" High x 6" Deep

Note: Five other models, with variations in power inputs and test frequencies, are also available.

■ The Sprague Model 1W1 Capacitance Bridge introduces a new concept in bridge design. Built by capacitor engineers for capacitor users, it incorporates the best features of bridges used for many years in Sprague laboratories and production facilities.

■ The internal generator of the 1W1 Bridge is a line-driven frequency converter, and detection is obtained from an internal tuned transistor amplifier-null detector, whose sensitivity increases as the balance point is approached. It has provision for 2-terminal, 3-terminal, and 4-terminal measurements, which are essential for accurate measurement of capacitors with medium, low, and high capacitance values, respectively.

■ The Model 1W1 Capacitance Bridge will not cause degradation or failure in capacitors during test, as is the case in many conventional bridges and test circuits. The 120 cycle a-c voltage, applied to capacitors under test from a built-in source, never exceeds 0.5 volt! It is usually unnecessary to apply d-c polarizing voltage to electrolytic capacitors because of this safe, low voltage.

For complete technical data on this precision instrument, write for Engineering Bulletin 90,010 to Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

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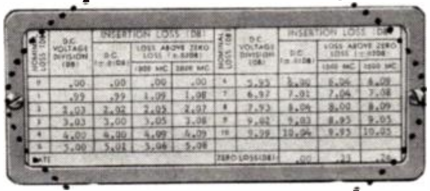

IRE People

(Continued from page 46A)

ment of Science. He received the Presidential Certificate of Merit for his World War II contributions, the honorary D.Sc. degree and a special alumni citation from Cornell College of Iowa, and is a member of Sigma Xi and Eta Kappa Nu.

❖

John A. Buckbee (A'41–M'47) has joined Image Instruments, Inc., Newton, Mass., it was announced recently. As a senior staff member, he will supervise the development of certain types of automatic recognition equipment and advanced information display systems. He is widely known for his work at Raytheon during the past several years as chief application engineer on electrostatic storage tubes. In his earlier work he designed equipment for the test of color tubes, carcinatrons, cathode-ray tubes, magnetrons and klystrons. While a senior engineer at Farnsworth, he developed color television systems, television cameras, and radar systems.

❖

W. G. Chaney

William G. Chaney (A'51–M'56) has been appointed manager of Mobile Telephone Systems engineering and Personal Communications engineering for General Electric's Communication Products Department, Lynchburg, Va. In his new position, he will be in charge of engineering for G-E's full simultaneous duplex mobile dial telephone units and other MTS products. He also has design responsibility for miniaturized personal communication equipment including the company's Voice Commander and Voice Director systems. He joined General Electric in 1960, after 13 years of telephone engineering with Bell Telephone Co. of Pennsylvania and AT&T in New York.

In 1943 he graduated from the University of Pittsburgh, Pittsburgh, Pa. with the B.S. degree in physics and engineering. From 1943 to 1947 he was with the Army Signal Corps as a radar and communications officer. From 1947 to 1950 he served as a radio instructor at the University of Pittsburgh Evening School. He joined the Bell Telephone Co. in Pennsylvania in 1947 and was active in Mobile Radio Systems Design with that company until 1955. From 1951 to 1955 he was an instructor in ac theory and ac Machinery and Transmission Lines at Drexel Institute of Technology, Philadelphia, Pa. From 1955 to 1959 he was engaged as an engineer in Mobile Telephone Systems Design for AT&T in New York.

Mr. Chaney was secretary of the Philadelphia Chapter of the Professional Group on Audio in 1955. In 1956 he became secretary of the New York Chapter of the Professional Group on Vehicular Communications. He was named chairman of the New York PGVC in 1958.

❖

G. F. Corcoran

Professor George F. Corcoran (A'43–SM'53–F'56), chairman of the University of Maryland department of electrical engineering, has been named the 1961 recipient of the Medal in Electrical Engineering Education by the American Institute of Electrical Engineers.

Established by the institute in 1956, the award is made annually to a teacher of electrical engineering based upon his excellence in teaching, his ability to inspire students to give higher achievements, and his contributions to the teaching of electrical engineering by textbooks and by writings on engineering education.

He received the B.S. degree in 1923 from South Dakota State College, Brookings, S. D. and the M.S. degree in 1926 from the University of Minnesota, Minneapolis, Minn. He was a teaching fellow in electrical engineering at the University of Minnesota in 1925, an instructor at Kansas State College in 1927, and associate professor and professor at the State University of Iowa from 1931 to 1941. He has been chairman of the University of Maryland's department of electrical engineering since 1941. He is a registered professional engineer and has served as a consultant to industry, to state and federal agencies.

Among his books are: "Introduction to Electrical Transients" (with E. B. Kurtz), "Alternating Current Circuits" (with R. M. Kerchner), "Basis Electrical Engineering," "Electrical Communications Experiments" (with H. R. Reed and T. C. G. Wagner), "Electronics" (with H. W. Price), and "Introductory Electrical Engineering" (with H. R. Reed).

Professor Corcoran is a member of Eta Kappa Nu, Sigma Tau, Tau Beta Pi, Phi Kappa Phi, and Sigma Xi, the American Society for Engineering Education, the American Institute of Electrical Engineers, and the American Association of University Professors.

❖

(Continued on page 50A)

FROM ELECTRONIC MEASUREMENTS

0–1500 V DC POWER SUPPLIES

Model 238A
With Rear Terminals

CONSTANT-VOLTAGE

In addition to covering a complete voltage range—0 to 1500 V—this programmable, regulated power supply can deliver rated current of 100 ma at any output voltage . . . even at less than a volt.

But an even more useful feature is its continuously variable current-limiting control. Here's a feature that's finding more application with every passing day.

When it comes to regulation, the Model 238A is truly flexible. Standard regulation is 0.1% line or load. However, 0.01% regulation is available by means of an optional plug-in chopper unit.

The chart below lists just a few of the many Electronic Measurements off-the-shelf, constant-voltage power supplies. For complete information request Catalog BR1961.

MODEL	RANGE VOLTS DC	 CURRENT
238A	0-1500 V DC	0-100 MA
239A	0-600 V DC	0-1 AMP
235A	0-600 V DC	0-500 MA
234A	0-300 V DC	0-500 MA
230A	0-200 V DC	0-1 AMP
228A	0-150 V DC	0-1 AMP
229A	0-150 V DC	0-300 MA
218A	0-100 V DC	0-3 AMP
221A	0-100 V DC	0-500 MA
225A	0-75 V DC	0-2 AMP
215A	0-50 V DC	0-3 AMP

Model C638A
With Rear Terminals

CONSTANT-CURRENT

Here's the newest in Electronic Measurements' line of Constant Current Power Supplies . . . a full 1500 V compliance at constant current outputs from less than a ma to 100 ma. Like all Electronic Measurements Power Supplies, the Model C638A has a lot of built-in extras too. For example, there's a modulation input . . . there are also provisions for remote programming . . . and an adjustable voltage compliance control that in effect provides for voltage-limiting.

Specifications include a high output impedance of 10^4 megohms at 0.5 μa to 0.5 megohm at 100 ma. Above 2.2 μa, regulation is better than 0.15%, line or load. Ripple is less than 0.01% +1 μa.

The following listing shows some of the constant current models available. For complete information, request Specification Sheet 3072B.

MODEL	RANGE MIN.	 MAX.	VOLTAGE COMPLIANCE (MINIMUM)
C614A	10 μa	1 AMP	100 V
C613A	10 μa	1 AMP	50 V
C621A	5 μa	500 MA	100 V
C620A	5 μa	500 MA	50 V
C633A	2.2 μa	300 MA	300 V
C629A	2.2 μa	300 MA	50 V
C632A	2.2 μa	220 MA	300 V
C624A	2.2 μa	220 MA	100 V
C638A	1 μa	100 MA	1500 V
C631A	1 μa	100 MA	300 V
C612A	1 μa	100 MA	100 V

E/M

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MEASUREMENTS
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Interstate Electronics is delivering the most reliable* ENERGY STORAGE ELECTROLYTIC CAPACITORS *in the industry

SANGAMO

TYPE DCM ELECTROLYTIC CAPACITORS are especially designed for use as energy storage components in DC circuitry where peak power requirements exceed the maximum output of the associated power supply. They operate under high temperature conditions, minimize ripple voltage and add stability and long life to low voltage power supplies.

That's why many computer manufacturers use the SANGAMO DCM. That's why you gain by turning to SANGAMO for your capacitor needs.

Complete data on Type DCM Capacitors is detailed in SANGAMO'S Engineering Catalog 2231. Write us for your copy.

"Definitely Dependable"

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Specialists in Military & Industrial Electronic Components

THE BATTERY THAT'S USED IN GUIDED MISSILES NOW RELEASED AS GOVERNMENT SURPLUS

For Photography, Aircraft, Models, Searchlights, Radios, etc.

$1.95 ea. Postpaid

Sintered-plate Nickel-Cadmium alkaline storage batteries designed for "Nike" Missile and now surplus due to design change. A lifetime battery with no known limit of service (over 5000 recharges on test without loss of capacity). Other features: Virtually indestructible, compact & lightweight, withstands heavy shock and vibration. Flat voltage curve during discharge, retains charge year or more, high discharge rate up to 50 amps. for this cell, no corrosive fumes to harm clothing or equipment, spill-proof construction, discharge in any position, indefinite storage without deterioration, operates in temperatures —60° F to +200° F. Each cell is approx. 4 ampere hour capacity, nominal voltage per cell is 1.2 volts. (A 6 V. battery requires 5 cells.) Cell size 6" H. × 2" W. × ½" T. Wt. 6 oz. ea. Uses Potassium-Hydroxide (30%) electrolyte. Negligible loss during lifetime service. Add only distilled water once a year. A fraction of Government cost.

Used Test Cells .... $1.95 ea. Postpaid
Brand New Cells .... 2.95 ea. Postpaid
6 Ampere Hour Cells size 4½" H. × 2⅜" W. × ¾" T.
Used test cells .... $2.95 New Cells .... $3.95 ea.
All cells guaranteed to your satisfaction or money refunded (less Postage).

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CREI

IRE People

(Continued from page 48A)

Sage Laboratories, Inc. has announced the appointment of **Jurgen Elkan** (S'50–A'51–M'55) as Microwave Engineering Specialist. In this capacity he will be responsible for the development of new lines of microwave components, with initial effort directed to the field of band-pass filters. Before joining Sage he was affiliated with Hermes Electronics Co., where he directed the communications planning for Weather Observing and Forecasting System 433L and had contributed to the design of HOT LINE, ACE HIGH and other UHF telecommunication systems. From 1951 to 1955, he was a staff member of M.I.T.'s Lincoln Laboratory and Research Laboratory for Electronics.


J. ELKAN

Mr. Elkan holds the B.S.E.E. and M.S.E.E. from M.I.T., Cambridge, Mass. He is a Registered Professional Engineer in Massachusetts.

❖

Appointment of **Richard K. Eskeland** (S'60–M'62) as district sales manager for the complete line of Du Mont Electronic Instruments at Allen B. Du Mont Laboratories, Division of Fairchild Camera and Instrument Corporation was announced recently. He will have responsibility for sales of Du Mont oscilloscopes, pulse generators, and oscilloscope record cameras in all of Long Island, N. Y. He comes to Du Mont from Lavoie Laboratories where he served as a regional sales manager. Prior to that he was with Ameresco, Inc. as an export sales engineer. He received his engineering education at Fairleigh Dickinson University.

❖

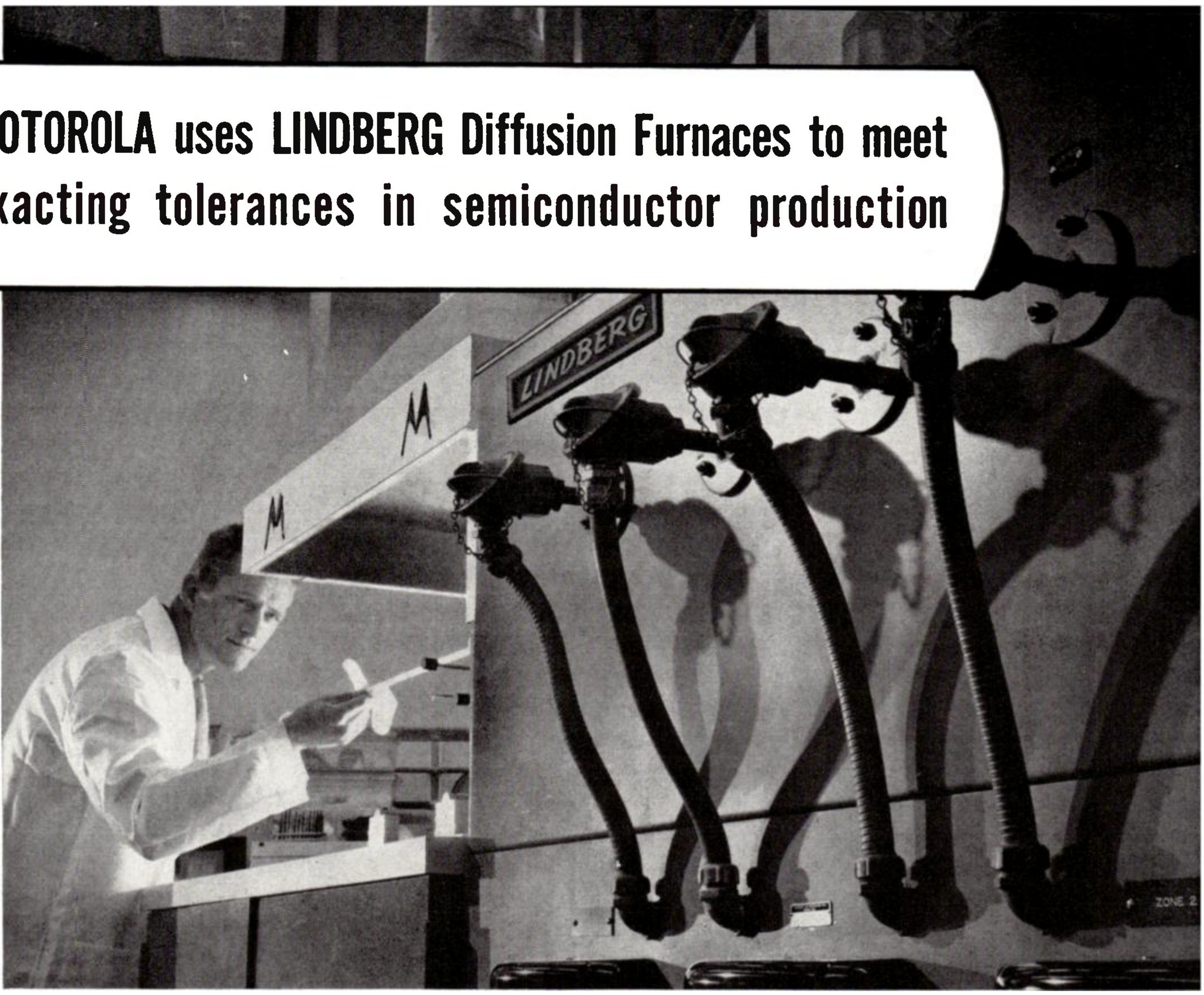
Thomas Evans (M'57) has been appointed Senior Engineer at White Electromagnetics, Inc. it was announced recently. He had been associated with the Jansky and Bailey Corporation since 1956.

He completed his undergraduate and graduate study in engineering at the University of Maryland. White Electromagnetics, Inc. (WEI) deals principally in RFI prediction and control, space communications systems and design, weapons sys-


T. EVANS

(Continued on page 54A)

MOTOROLA uses LINDBERG Diffusion Furnaces to meet exacting tolerances in semiconductor production



QUOTE
from
Mr. Huffman

Mr. Tom Huffman, Project Leader for Diffusion Laboratories, Motorola Semiconductor Products Division, Phoenix, Arizona.

"The precise engineering of Motorola silicon and germanium diffused mesa transistors requires control of diffusion depths in the order of a few hundred angstroms. With our Lindberg Type RDT8-236 Gaseous Solid State Diffusion Furnace, we are able to obtain this close control of transistor diffusion depths."

More than 25 Lindberg Diffusion Furnaces of the type illustrated are used by Motorola Semiconductor Products Division at Phoenix, Arizona. These furnaces are from Lindberg's complete line of equipment for production of semiconductors, solid state and other electronic devices.

Processes covered by Lindberg equipment include gaseous and solid diffusion, germanium reduction, crystal growing, zone refining, bonding, and special processes with controlled atmospheres or vacuum. This Lindberg equipment is precision-designed for precision results and dependable production performance. For complete information write for our new Semiconductor Equipment Catalog or get in touch with your local Field Representative (see your classified phone book). Semiconductor Equipment Division, Lindberg Engineering Company, 2482 West Hubbard Street, Chicago 12, Illinois.

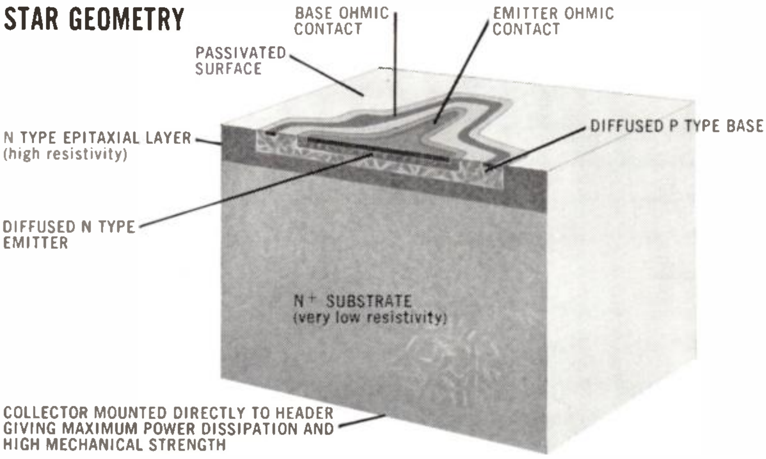
Los Angeles plant: 11937 S. Regentview Avenue, Downey, California. In Canada: Birlefco-Lindberg, Ltd., 15 Pelham Ave., Toronto 9, Ont. Also, Lindberg plants in Argentina, Australia, England, France, Italy, Japan, Africa, Spain, Switzerland and West Germany.

The type RDT8-236 Furnace used at Motorola is equipped with 2½" O.D. quartz work tube inside Inconel radiator and provides 16" at plus or minus 1.0°C. Special alloy rod overbend heating elements, 2 zones of control (saturable core reactor).

LINDBERG
heat for industry

THE MOTOROLA SILICON

HOW TO PRODUCE A "UNIVERSAL" TRANSISTOR WITH OPTIMUM PERFORMANCE AND MAXIMUM RANGE OF USAGE



FIRST - Create a new geometry to provide *increased* edge periphery of the emitter-base for desirable current rating with decreased junction area for good high frequency characteristics.

SECOND - Combine the new geometry with Motorola's superior epitaxial process to provide higher frequency response, lower total control charge, high breakdown voltages, and at the same time, lower saturation voltage and collector capacitance.

THIRD - Protect the new epitaxial device against junction contamination during manufacture, and against change with time, and ensure more uniform gain at low and high currents by passivating the surface through the planar process.

WHAT THE STAR PLANAR CAN MEAN TO DESIGNERS

To one company it meant being able to reduce the number of different devices required in their equipment from 7 types to 1 type. To another it meant effective high-current switching 4 times faster than the type previously used.

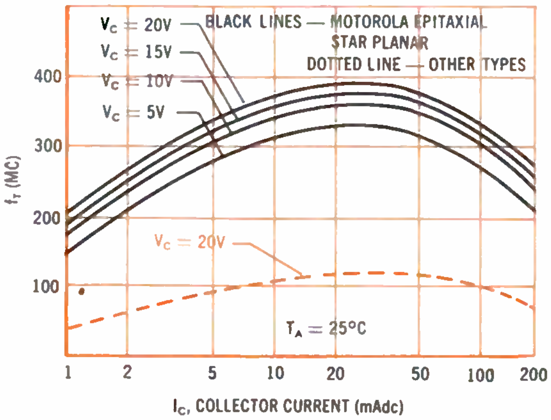
HERE ARE THE NEW MOTOROLA SILICON EPITAXIAL STAR PLANAR TYPES

The introduction of Motorola's new Epitaxial "Star" Planar family marks a dramatic advance in the design and manufacture of silicon transistors. The new devices are available now in TO-5 and TO-18 packages. Only Motorola offers these star planar types.

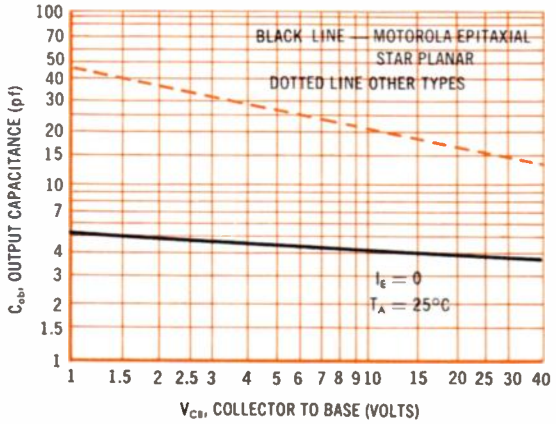
	TO-5 Package			TO-18 Package		
	MM486	MM487	MM488	MM511	MM512	MM513
h_{FE} @ $I_C = 150$ mA	20-60	40-120	100-300	20-60	40-120	100-300
$V_{CE(sat)}$ volts (max) $I_E = 500$ mA, $I_B = 50$ mA	—	1.6	1.6	—	1.6	1.6
C_{ob} $I_E = 0$, $V_{CB} = 10$ V	8 picofarads (maximum) — All Types					
f_T $I_C = 20$ mA, $V_{CE} = 20$ V	250 mc (minimum) — All Types					
Switching Time (total) non-saturated	14 nsec (typical) — All Types					
Switching Time (saturated) t_{on}	60 nsec (typical) — All Types					
t_{off}	45 nsec (typical) — All Types					

EPITAXIAL STAR PLANAR

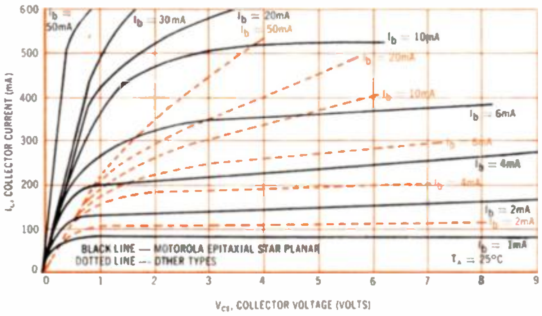
COMPARE THESE PERFORMANCE FEATURES



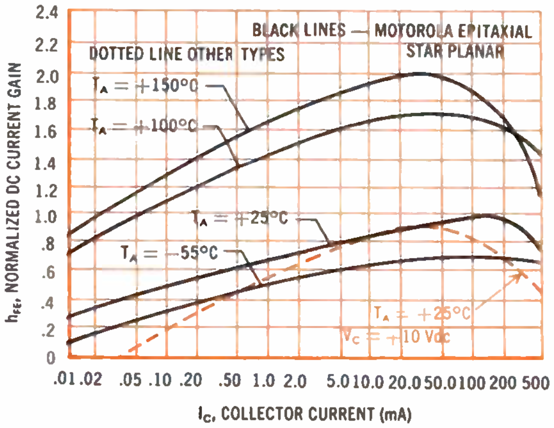
400% HIGHER GAIN-BANDWIDTH PRODUCT The small emitter area (star) and Motorola's advanced diffusion techniques combine to improve frequency performance.



ONE FIFTH THE OUTPUT CAPACITANCE The reduced emitter area of Star geometry, plus the high-resistivity epitaxial layer combine to substantially lower collector capacitance.



ONE SEVENTH THE SATURATION VOLTAGE With the low substrate resistance of the epitaxial process, collector saturation voltage is greatly reduced.



EXTENDED BETA RANGE — FROM 10 μA TO .5 AMPS Passivated to stabilize surface characteristics and eliminate recombination effects, this new transistor provides more uniform gain at low and high current.

The Motorola Silicon Epitaxial Star Planar types are immediately available from your Motorola Semiconductor Distributor. So, if yours is an application requiring high-speed, high-current switching for line driving applications . . . or if you need low I_{CBO}, low-noise front end silicon amplifiers . . . or you want the low output capacitance at low voltage which makes it possible to choose either a low or high capacitance tank circuit in a tuned RF Power Amplifier, the "Universal" Epitaxial Star Planar is your answer.

Compare the Motorola Epitaxial Star Planar with the type you are presently using in your application. Prove to yourself that this "universal" transistor gives you the improved performance you want.

For more information on this remarkable new Motorola Epitaxial "Star" Planar family, contact your local Motorola Distributor or Distributor, or call or write: Motorola Semiconductor Products, Inc., Technical Information Department, 5005 East McDowell Road, Phoenix 8, Arizona.

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4545 West Augusta Blvd., Chicago 51, Ill. .... SPaulding 2-6500

IRE People

(Continued from page 50A)

tem evaluation, system evaluation by computer simulation, and other phases of systems analysis.

❖

Appointment of **Alfred M. Faries** (M'59) as Systems Division representative in the Seattle office of Electro-Mechanical Research, Inc. was recently announced. His duties will include field liaison relative to the Dyna-Soar manned space-glider program. EMR is designing and manufacturing the test-instrumentation subsystem for Dyna-Soar; the Boeing Company is the prime contractor for this USAF project which will result in controlled orbital flight by an astronaut.

A. M. Faries

After receiving a BSEE degree from Stanford University in 1955, he became a research assistant in Stanford's Radio Propagation Laboratory. He then spent three years as a development engineer at Lenkurt Electric Company, San Carlos, Calif. His last position prior to joining EMR was as a senior research engineer with the Lockheed Missiles and Space Division, Sunnyvale, Calif., where he coordinated the integration of components and systems for the Polaris missile program.

Mr. Faries is a member of the Special Working Group on Digital Telemetry of the U. S. Navy Special Projects Office.

❖

Dr. Raoul J. Freeman (M'59) has been appointed President and Chief Executive Officer of Researnalysis Company, Santa Monica, Calif. He assumes responsibility for guiding the activities of the international scientific consulting firm. Researnalysis specializes in R&D Planning, Operations Research, Value Analysis, Systems Analysis, and PERT Systems. Via its branch office in Amsterdam, Holland, it provides service to the Common Market countries. He was formerly Project Director with CEIR Inc., where he was concerned with military and commercial operations research and large scale electronic computer applications.

Dr. Freeman received his Ph.D. degree in 1957 in Operations Research at Massachusetts Institute of Technology, Cambridge, Mass. and since then has published over twenty papers in the fields of R&D planning, operations research, and mathematics. He is a member of the Operations Research Society of America, and the Institute of Management Sciences (Chairman of the Southern California Chapter). He has also given seminars in the Graduate School-U.C.L.A., and has lectured extensively in Europe and the United States.

❖

Interstate Electronics Corporation, Anaheim, Calif., a wholly-owned subsidiary of Interstate Engineering Corporation, has announced the appointment of **A. V. Gangnes** (SM'58) to Vice President and General Manager. The Anaheim, California firm is prime contractor for test instrumentation in the U. S. Navy Fleet Ballistic Missile Weapon System (POLARIS) program.

A. V. Gangnes

He is a graduate of the University of Washington where he received a degree in Electrical Engineering. He has been active in the electronics systems development field for fourteen years. He is responsible for coordination of the various departments.

❖

Dr. Constantine D. J. Generales (SM'59) was the recent recipient of an honorary certificate from the Greek government for original research in space science and space medicine. The award was presented through the Hellenic Astronautical Society. He is an assistant professor of space medicine and coordinator of the Space Medicine Program at New York Medical College, Metropolitan New York Medical Center.

(Continued on page 57A)

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

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

COMPARE PERFORMANCE, PRICE PER CHANNEL

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

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

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

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

(Specifications subject to change without notice)

SPECIFICATIONS

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

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

Bandwidths (Max)

Speed	*FM*	*Direct*
3¾″/sec	0-625 cps	50-6,250 cps
7½″/sec	0-1,250 cps	50-12,500 cps
15″/sec	0-2,500 cps	50-25,000 cps
30″/sec	0-5,000 cps	100-50,000 cps

(100% modulation on FM = ± 40% carrier deviation)

Linearity Maximum error due to nonlinearity: 0.2%.

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

Signal-to-Noise Ratio (Min)

Direct: 40 db at all speeds.

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

Readout, as well as input monitoring during magnetic recording, may be provided by this compatible 17-inch, 8-channel Viso-Scope or other Sanborn monitoring instruments, or by direct writing systems.

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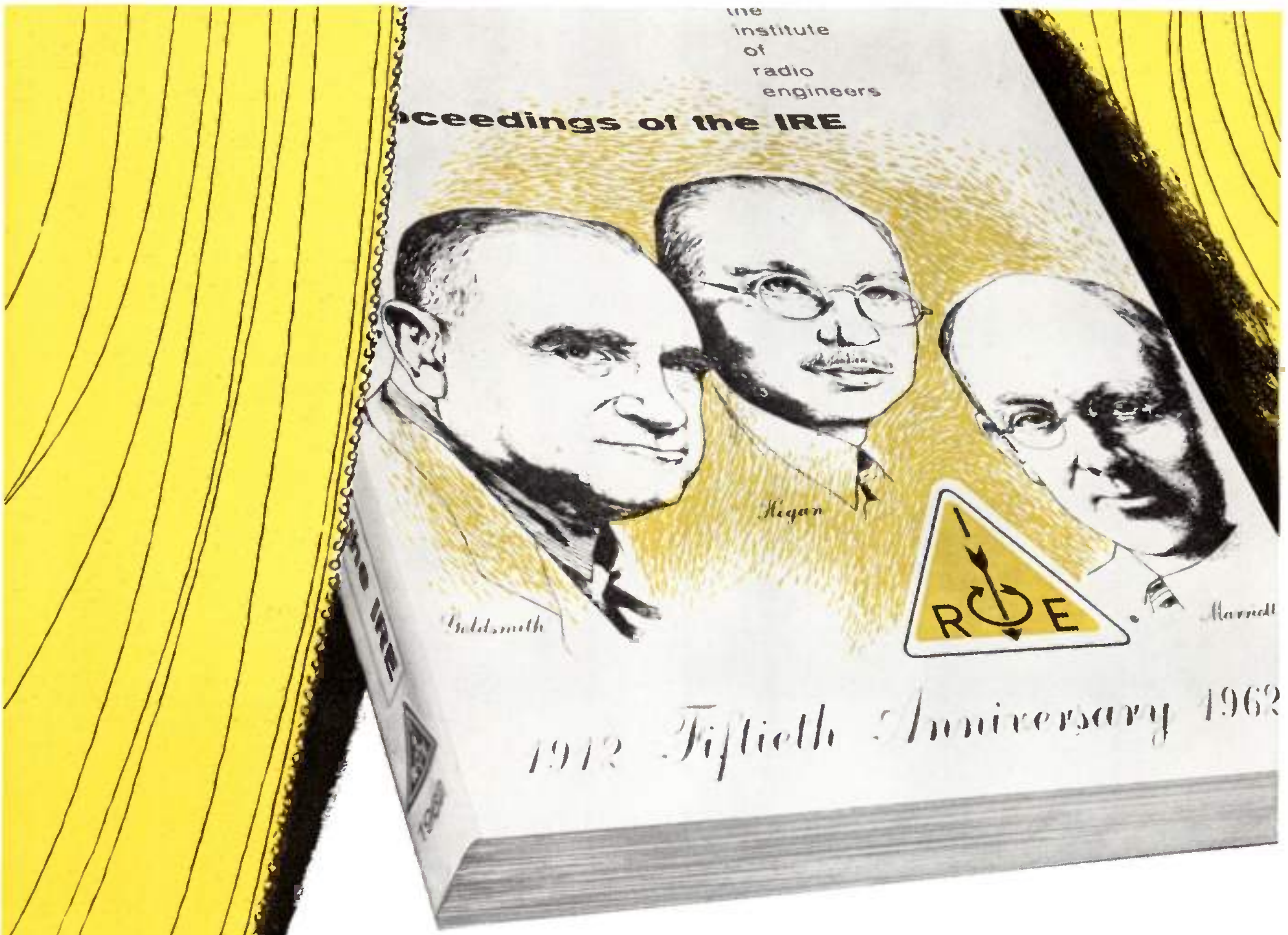

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GOLDEN ANNIVERSARY

ISSUE OF **Proceedings**

In honor of the IRE's 50th anniversary, the May 1962 issue of *Proceedings of the IRE* will survey the progress of electronics from 1912 to 1962. It will forecast the future of electronics for the next 50 years. And this special issue of *Proceedings* will have more than 600 pages of additional editorial text, discussing developments in

- Nuclear Science
- Circuit Theory
- Broadcasting
- Education
- Audio
- Instrumentation
- Electron Devices
- Components Parts
- Information Theory
- Electronic Computers
- Automatic Control
- Vehicular Communications
- Communications Systems
- Engineering Management
- Industrial Electronics
- Bio-Medical Electronics
- Ultrasonics Engineering
- Military Electronics
- Antennas and Propagation
- Broadcast and Television Receivers
- Product Engineering and Production
- Reliability and Quality Control
- Radio Frequency Interference
- Space Electronics & Telemetry
- Microwave Theory and Techniques
- Human Factors in Electronics
- Aerospace & Navigational Electronics

Electronics experts from all over the world contribute to make this anniversary issue concise yet all-inclusive. 100,000 copies will be printed. Only members of the IRE, in good standing in May 1962, and subscribers will receive a copy. We regret that unlike special issues of the past, no copies can be made available to non-members, or for single copy sales. As a member of the IRE you will find the May anniversary issue of *Proceedings* the master reference work you will treasure always.

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IRE People

(*Continued from page 54A*)

He graduated from Harvard College and received his medical education at the Universities of Athens, Heidelberg, Paris, Zurich and Berlin. He holds the degrees of Doctor of Medicine and Doctor of Philosophy, the latter received at the University of Berlin. He has served as a flight surgeon in the Air Force and has been chief of surgical service at various hospitals. He has also served on the Repatriation Committee for the International Red Cross. In 1960 he served with the Space Science group on its visits to Leningrad and Moscow after the International Aeronautical Congress held in Stockholm.

Dr. Generales is a member of the American Astronautical Society, American Rocket Society, British Interplanetary Society, and others.

❖

R. Karl Honaman (A'23–SM'44) has been elected Chairman of the Board and Chief Executive Officer of Floating Floors, Inc., New York, N. Y. This company provides environmental facilities for data processing systems and installations of a similar nature.

❖

Algert G. Grimaila (S'48–A'51–M'56) has been appointed Chief Engineer for Space Technology at Western Center, Motorola Military Electronics Division, Scottsdale, Ariz. He was formerly Chief Engineer of the Communications and Navigation Laboratory.

A. G. Grimaila

The newly formed Space Technology Group will include members of the Advanced Techniques Section and will report to him. His responsibility is to further expand Motorola technology in the space electronics field, direct technical studies and proposals on space programs, and supply engineering support to Marketing. He has been with Motorola for eleven years. The first seven years were spent in performing research and development of Military Electronics equipment in the Chicago laboratories of Motorola. In 1957 he joined the Western Center of Motorola Military Electronics Division as Assistant Chief Engineer of the Defense Systems Laboratory. Appointed Chief Engineer of the Communications and Navigation Laboratory in 1959, he provided technical direction for the extensive missile and space communications, guidance, and control projects of that facility.

Mr. Grimaila received the B.S. degree

(*Continued on page 58A*)

IRE People

(*Continued from page 57A*)

in Electrical Engineering from the University of Connecticut in 1950 and his M.S. in Electrical Engineering from Illinois Institute of Technology in 1955. Since then he has pursued further graduate studies at Illinois Institute of Technology, University of Arizona, and Arizona State University. He is a member of Tau Beta Pi, Eta Kappa Nu, Institute of Navigation, and Association of the U. S. Army.

❖

G. R. Hetrich

George R. Hetrich (M'57–SM'61) has joined the Research Division of Electronic Communications, Inc. in Timonium, Md. as a Senior Engineer. He will have responsibilities in the area of communication systems. He comes to this new position after having been associated with Westinghouse Electric Corporation for seven years. Most of his career was spent at the Air Arm Division in Baltimore, Md., working in the Advanced Development Section of the Engineering Department. His activities there were related to all phases of coherent radar development.

Mr. Hetrich holds the B.S.E.E. degree and a B.S. degree in engineering physics from Lehigh University and an M.S.E.E. degree from the University of Pittsburgh. He is a Member of the American Institute of Electrical Engineers.

❖

R. E. Hiatt

Ralph E. Hiatt (M'47–SM'58) has been appointed Head of the Radiation Laboratory of the Department of Electrical Engineering of The University of Michigan. He had been Associate Head since July, 1958 and was responsible for the supervision of the laboratory's anechoic chamber and experimental facility.

He received the M.A. degree in physics from Indiana University, Bloomington, Ind. in 1939 and took post-graduate work in mathematics at Boston University, Boston, Mass. Prior to joining the Radiation Laboratory he was Chief of the Antenna Laboratory at the Air Force Cambridge Research Center. During World War II, he was Chief of the Ipswich Antenna Field Station of the Radiation Laboratory of Massachusetts Institute of Technology. He supervised and participated in experimental and theoretical research in the fields of electromagnetic radiation and scattering, ground-based antennas, airborne antennas, and waveguide components. He joined The University of Michigan staff in 1958. His research at the University has included experimental and theoretical investigations of the scattering properties of basic geometrical shapes, both rough and smooth, coated and uncoated, and of the properties of microwave absorbing materials.

Mr. Hiatt is listed in *American Men of Science* and is a member of the American Institute of Physics and Sigma Xi. He was President of the Air Force Cambridge Research Center Branch of the Research Society of America in 1957–1958.

❖

The appointment of **Milton Jennis** (M'60) to the position of Projects Director of Malaker Laboratories, Inc., was announced recently. He attended Stevens Institute, was manager of the manufacturing engineering department for the W. L. Maxson Corp., and was also associtiated with Sperry Gyroscope, Mergantahler Linotype Co. and Universal Camera. He has been in the electronic and electro-mechanical field for more than twenty years.

Mr. Jennis is a member of the Army Ordnance Association and the American Society of Metals.

❖

L. I. Kent

Leonard I. Kent (S'49–A'51–M'55–SM'58) has been appointed President of the Consolidated Microwave Corporation, the newly formed affiliate of Decitron Electronics Corporation, Brooklyn, N. Y. He is also Vice President and Director of Engineering of Decitron. CMC will participate in the design, production, research and development of specialized microwave equipment and instrumentation. He is a graduate of the Cooper Union School of Engineering, New York, N. Y. where he received the B.E.E. degree. He has also been awarded the M.E.E. degree and is completing requirements for the D.E.E. degree at the Polytechnic Institute of Brooklyn.

He was Vice President of the Engineering Division of Microwave Semiconductor and Instruments, Inc. Prior to this he was Director of Engineering at Narda Microwave Corporation. He joined Narda as Chief Engineer in 1957. He was previously associated with FXR, PRD, Airborne Instruments Laboratory and the Espey Manufacturing Company. Among his accomplishments have been contributions in the field of directional couplers, ridged waveguides, millimeter-wave techniques, bolometric and solid state devices. Several papers have been written by him on these subjects.

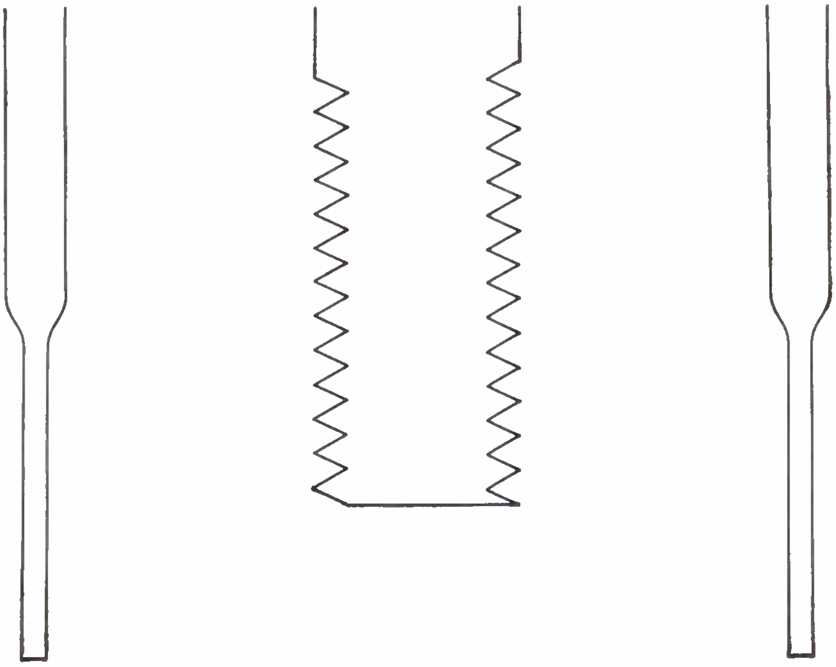
Mr. Kent is a member of the Committee on Waveguides and Fittings of the EIA, and a member of the Transducer Committee of the ISA.

❖

(*Continued on page 60A*)

TAKE A SECOND LOOK

IT'S THE 2N174—PART OF DELCO RADIO'S POWER TRANSISTOR FAMILY WHICH HAS PROVED ITS STUFF FOR YEARS IN HUNDREDS OF MILITARY AND INDUSTRIAL APPLICATIONS: MISSILES, COMMUNICATIONS, DATA PROCESSING, AND ULTRASONICS, TO NAME A FEW. THIS MULTI-PURPOSE PNP GERMANIUM POWER TRANSISTOR HAS THE HIGH PERFORMANCE AND VERSATILITY TO MEET OR EXCEED THE MOST RIGID ELECTRICAL AND ENVIRONMENTAL REQUIREMENTS. ℄ DESIGNED FOR GENERAL USE WITH 28-VOLT POWER SUPPLIES, THE 2N174 MAY ALSO BE USED WITH 12 VOLTS WHERE HIGHER RELIABILITY IS DESIRED. MAXIMUM EMITTER CURRENT—15 AMPERES, MAXIMUM COLLECTOR DIODE RATING—80 VOLTS, THERMAL RESISTANCE—BELOW .6°C/W AND MAXIMUM POWER DISSIPATION—50 WATTS AT 71°C, MOUNTING BASE TEMPERATURE. THE 2N174'S LOW SATURATION RESISTANCE PROVIDES HIGH EFFICIENCY IN SWITCHING OPERATIONS. ℄ LIKE ALL DELCO TRANSISTORS, EVERY 2N174 MUST PASS AT LEAST A DOZEN ELECTRICAL AND ENVIRONMENTAL TESTS—BEFORE AND AFTER AGING—BEFORE IT LEAVES DELCO RADIO'S LABORATORIES. THIS 200 PERCENT TESTING, COMBINED WITH FIVE YEARS OF REFINEMENTS IN MASS PRODUCTION, MEANS CONSISTENT UNIFORMITY IN THE PRODUCT . . . AT A LOW PRICE. ℄ THE 2N174 IS JUST ONE OF MANY DEPENDABLE TRANSISTORS PRODUCED BY DELCO RADIO TO SUPPLY ALL YOUR TRANSISTOR NEEDS. FOR MORE DETAILS OR APPLICATIONS ASSISTANCE ON THE 2N174 OR OTHER DELCO TRANSISTORS, CONTACT YOUR NEAREST DELCO RADIO SALES OFFICE.



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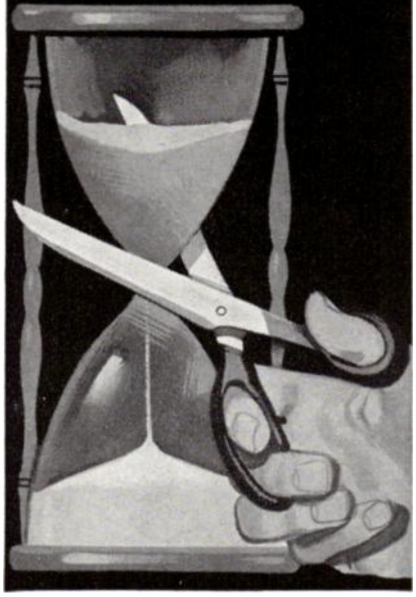

IRE People

(Continued from page 58A)

Dr. Andrew O. Jensen (S'50–M'57) will head the newly established Energy Conversion Department in the Advanced Power Systems Division of Electro-Optical Systems, Inc., Pasadena, Calif., it was recently announced. This group will be responsible for applied research, development, test, and production activities in the field of static energy conversion. He will direct the Department's activity in areas of thermionics, thermoelectrics, solar cells and research, leading to their integration into complete power packages for a variety of space vehicle and auxiliary ground power unit applications.

He joined Electro-Optical Systems in September, 1961. Prior to this time, he was employed as a research specialist at General Electric Company's Power Tube Laboratory in Schenectady, N. Y. During his five years at GE, he was actively engaged in numerous vacuum and vapor thermionic diode development programs for space and ground applications, as well as research on super power hydrogen thyratrons and other general gas discharge problems. Numerous technical articles and papers covering the above work have gained him an international reputation.

Dr. Jensen received his B.S., M.S. and Ph.D. degrees in electrical engineering from the University of Minnesota, Minneapolis, Minn.

He is a member of Eta Kappa Nu, and Gamma Alpha.

❖

General Microwave Corporation, Farmingdale, N. Y., announces the appointment of **Charles Lamensdorf** (S'55–M'60) as Project Engineer. This work will involve him in the design and development of microwave components, test equipment and systems equipments. He received the B.S.E.E. degree in 1959 from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. and is presently pursuing the M.S.E.E. degree at the same Institute.

C. Lamensdorf

Prior to joining General Microwave Corporation, he was employed at the Airborne Instruments Laboratory as a Design Engineer, from 1959 to 1961. In this capacity, he was involved with the design and development of front ends of microwave receivers in the UHF and microwave region. He became specialized in designing low power triode and klystron oscillators. From 1955 to 1959, he was employed by the Microwave Research Institute as an Engineering Assistant and was involved in the design and development of microwave ferrite components. During the same period of time, he was engaged as a consultant to Microwave Consultants, a group organized by several staff members of the Microwave Research Institute.

❖

Dr. Sumner N. Levine (SM'58) has been appointed Professor of Engineering and Chairman of the Materials Sciences Department of the College of Engineering at the State University of New York, Long Island, N. Y. He received a B.S. degree in physics and chemistry from Brown University in 1946 and a Ph.D. in physical chemistry from the University of Wisconsin in 1949. Prior to assuming his present position, he directed the advanced materials and electronics components research at RCA Surface Communications from 1958 to 1961. He served on the faculties of the University of Chicago from 1949 to 1950 and Columbia University from 1950 to 1954, and headed advanced materials research at the American Machine and Foundry Company from 1956 to 1958. He has also been active in medical and biological research and organized and directed the Research Laboratories of the U. S. Veterans Administration, East Orange, N. J. from 1954 to 1956.

S. N. Levine

Dr. Levine is a member of the Ameri-

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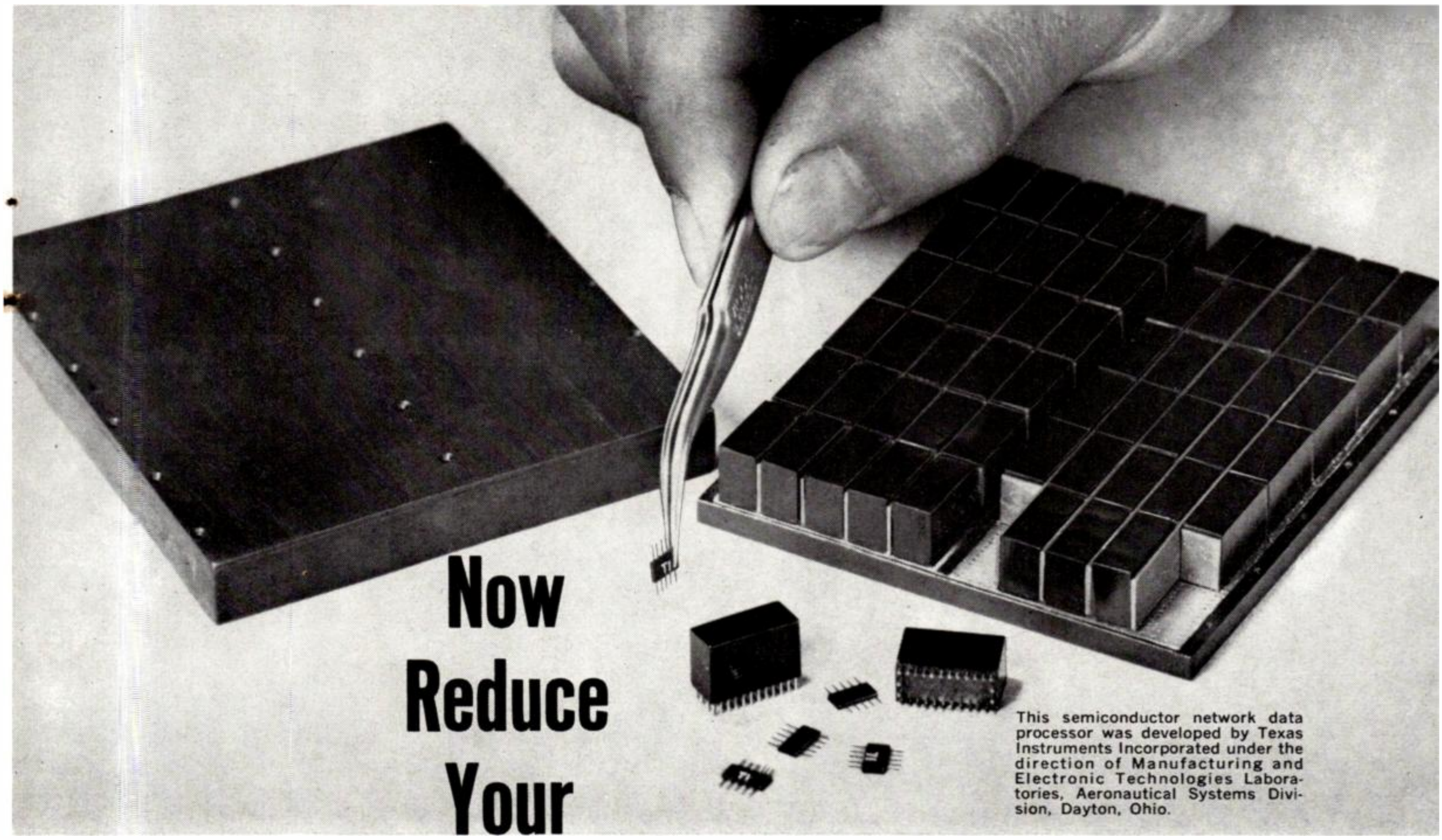
This semiconductor network data processor was developed by Texas Instruments Incorporated under the direction of Manufacturing and Electronic Technologies Laboratories, Aeronautical Systems Division, Dayton, Ohio.

Now Reduce Your Microelectronic Design Time

with Series 51 SOLID CIRCUIT semiconductor networks*

HERE'S HOW:

- designed to fulfill logic functions of complete equipment assemblies—compatible with most of today's logic circuitry.
- low power drain minimizes thermal problems and reduces power supply requirements.
- advanced manufacturing techniques including diffused planar structures, deposited leads, oxide protection, and hermetically sealed package—offer you the potential for improved circuit reliability.
- today's ultimate in microelectronics—with the highest function/size ratio for your digital circuits or equipments.
- provide reduced microelectronic cost through TI's standard silicon wafer design.
- meet military requirements:

Power Drain	2-4 mw @ 3 volts
Fan-Out (TI SN 510, 512, 514, 515)	4
Fan-Out (TI SN 511, 513)	20
Propagation Delay	75 to 450 nsec
Power Supply	3 to 6 volts
Temperature Range	—55° to +125°C

UNIT	TI SN 510	TI SN 511	TI SN 512	TI SN 513	TI SN 514	TI SN 515
FUNCTION	Flip Flop, Counter	Flip Flop with emitter follower output	NOR/NAND Gate (6 input)	NOR/NAND Gate (6 input) with emitter follower output	Two NOR/NAND Gates (3 inputs each)	Exclusive OR
	Clock pulse is internally capacitive-coupled					

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(*Continued from page 60A*)

can Physical Society, the American Chemical Society, the American Society for Metals, Sigma Xi and the New York Academy of Sciences.

❖

The appointment of **Dr. R. E. Malm** (SM'56) as Manager of the Research and Development Department of the Apparatus Division of Texas Instruments Inc., Dallas, Tex. was announced recently. The Research and Development Department provides product-oriented study and experimentation leading to new business in the defense electronics field for the Apparatus Division.


R. E. Malm

Before joining TI, he was a department manager in the Advanced Communications Systems Laboratory of Sylvania Electric Products. At Sylvania, he was responsible for research in physical and mathematical aspects of communication as well as synthesis and analysis of advanced antijam and low-detectibility communication systems. Prior to 1957 he was a member of the technical staff at Bell Telephone Laboratories for three years, where he worked in the areas of military system studies, electromagnetic propagation, and information theory. He was an associate physicist at Argonne National Laboratory from 1950 to 1954, where he did research in nuclear physics concerned with energy levels in light nuclei.

Dr. Malm holds a Ph.D. degree in physics from Massachusetts Institute of Technology, Cambridge, Mass., and a B.S.E.E. degree from the University of Minnesota, Minneapolis, Minn. He is a graduate of the Sylvania Management Institute, and is a member of American Physical Society, American Association for Advancement of Science, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

❖

The appointment of **A. Ray McCord** (M'57) as Manager of the Surveillance Department of Texas Instruments Inc., Dallas, Tex. was announced recently. He will investigate systems applications of thin film circuits, semiconductor networks and magnetic memories. His efforts will be directed to broadening the scope of TI's contractual work and commercial applications in these areas.


A. R. McCord

He joined Texas Instruments in 1951 as an engineer. He worked in the microwave design group and did development work on a tail-warning radar project and a transistorized interphone amplifier. He has also served as a project engineer on two reconnaissance radar systems, as section leader of a radar reconnaissance branch, and as production manager of the Apparatus Division's manufacturing department. He holds a B.S. degree in chemistry from Centenary College, Shreveport, La., and has taken graduate courses in electrical engineering at Southern Methodist University, Dallas, Tex.

◆

Aerospace computer developments and their application to military weapons systems is the responsibility of **S. L. McDonough** (S'49–A'50–M'55) manager of the new weapons applications department at the Univac Military department, St. Paul, Minn. Former group manager of digital computer engineering at McDonnell Aircraft Corp., St. Louis, Mo., he has the job of "uniting St. Paul Univac's advanced computer techniques (*i.e.*, thin-film memories) and microtronic circuitry to effect very small, lightweight, highly reliable computers for aerospace applications." During his 18 months at McDonnell Aircraft, he pioneered the design of light weight digital computers for guidance and control of missiles and space vehicles. As a principal engineer in the electronics department of Cornell Aeronautical Laboratories, Buffalo, N. Y. from 1949 to 1960, he was credited with design of the INTACC (Intercept Tracking and Control Computer) used by the Navy for radio-controlled vectoring of aircraft.


S. L. McDonough

Mr. McDonough holds the B.S.E.E. degree and the M.S.E.E. degree from the University of Iowa. He is a member of Tau Beta Pi, Eta Kappa Nu and Sigma Xi.

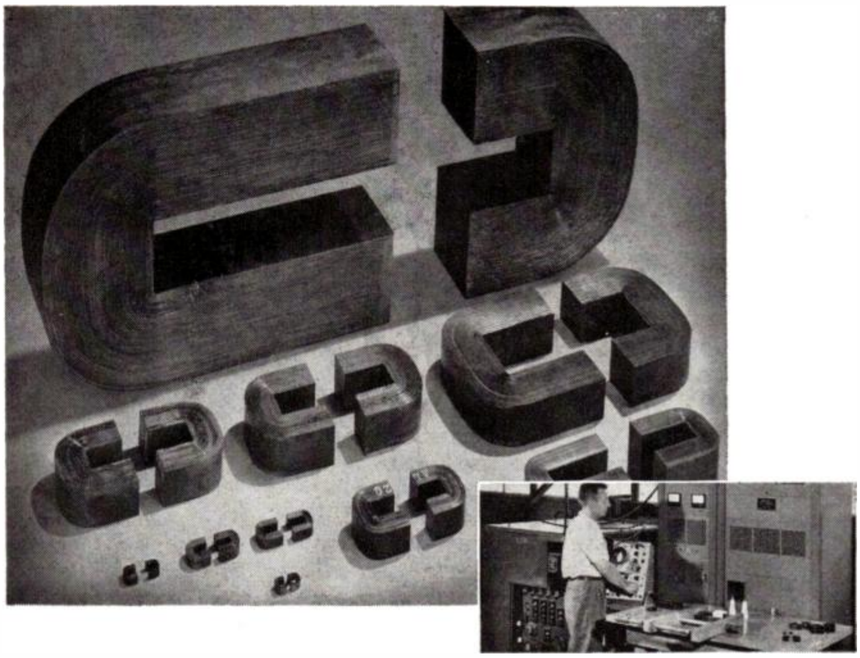
◆

Richard J. Mohr (M'58) has been appointed Microwave Department Head of the Narda Microwave Corporation, Mineola, N. Y. He will be in charge of the engineering research and development effort to expand Narda's line of microwave instruments and components.


R. J. Mohr

He has an extensive background in microwave engineering, having worked for more than nine years for the Naval Material Laboratory, Brooklyn, N. Y., Airborne Instruments Laboratory, FXR, Inc. and recently as Microwave Section Head at the Microwave Dynamics Corp., Plain-

(*Continued on page 66A*)



Pioneering New Oil Wells in the Laboratory...

500,000 ALLEN-BRADLEY HOT MOLDED RESISTORS HELP MAP STRATA TO FIND "PRODUCERS"

This network of over 1,000 A-B hot molded resistors is one of hundreds of similar grids developed by the Schlumberger Well Surveying Corp. for studying ground strata to locate producing zones.

The unusually large number of resistors in use per unit makes reliability of paramount importance. Therefore, Allen-Bradley resistors—with their history of *complete freedom from catastrophic failure*—were a logical selection for this unusually critical project. The *exclusive* A-B hot molding process makes possible the amazing uniformity for which Allen-Bradley resistors are famous. To eliminate the probability of resistor failure in your equipment, Allen-Bradley resistors can be your only choice.

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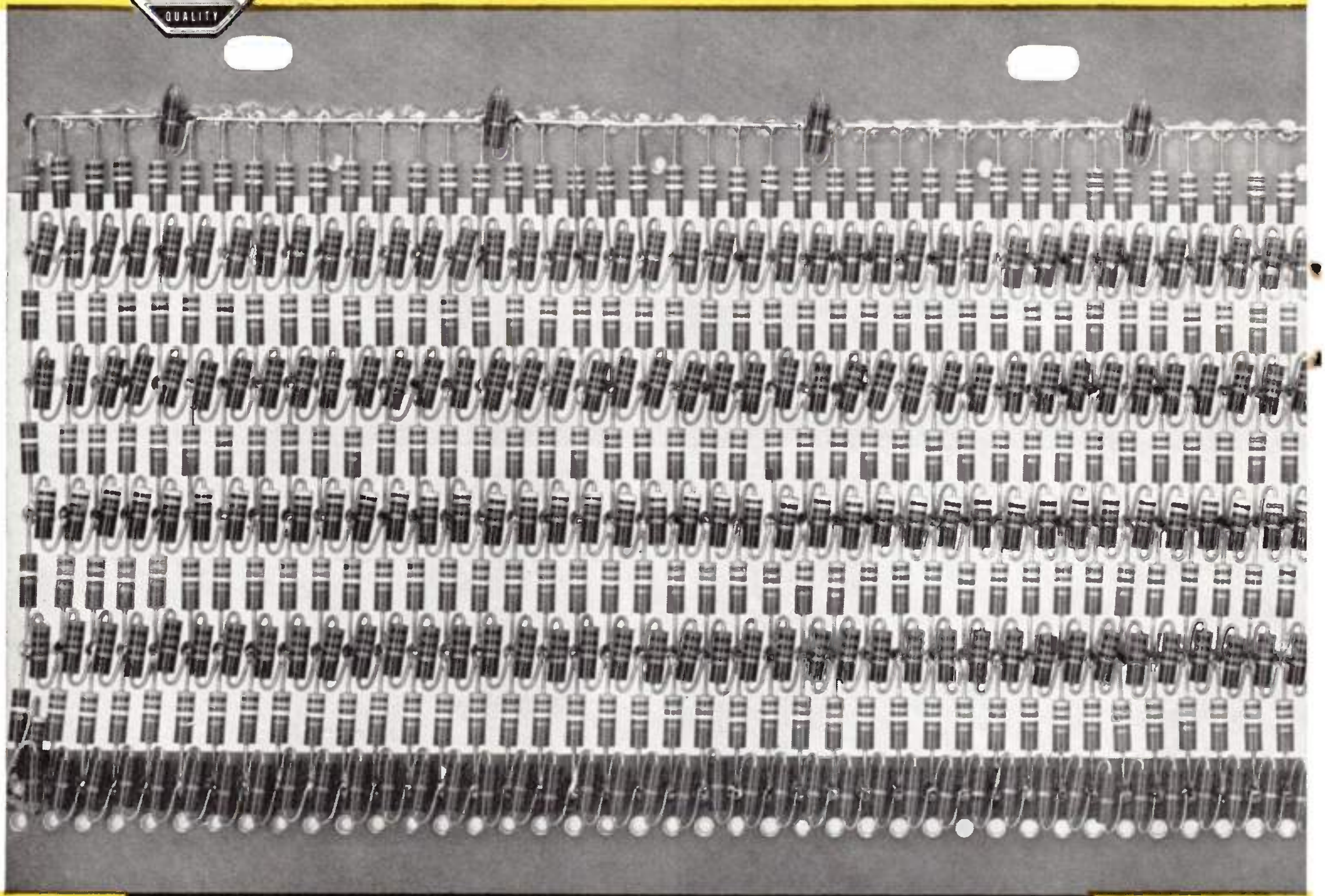

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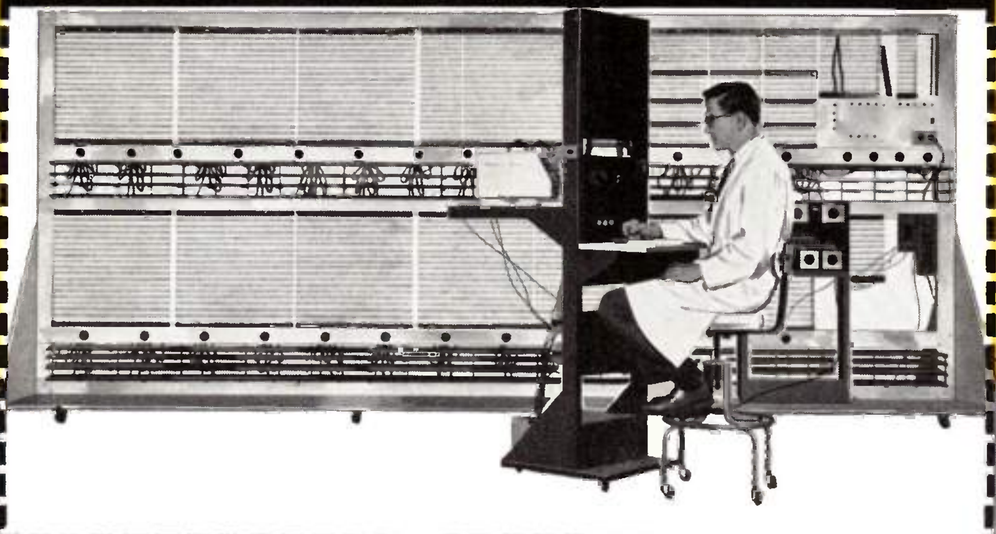
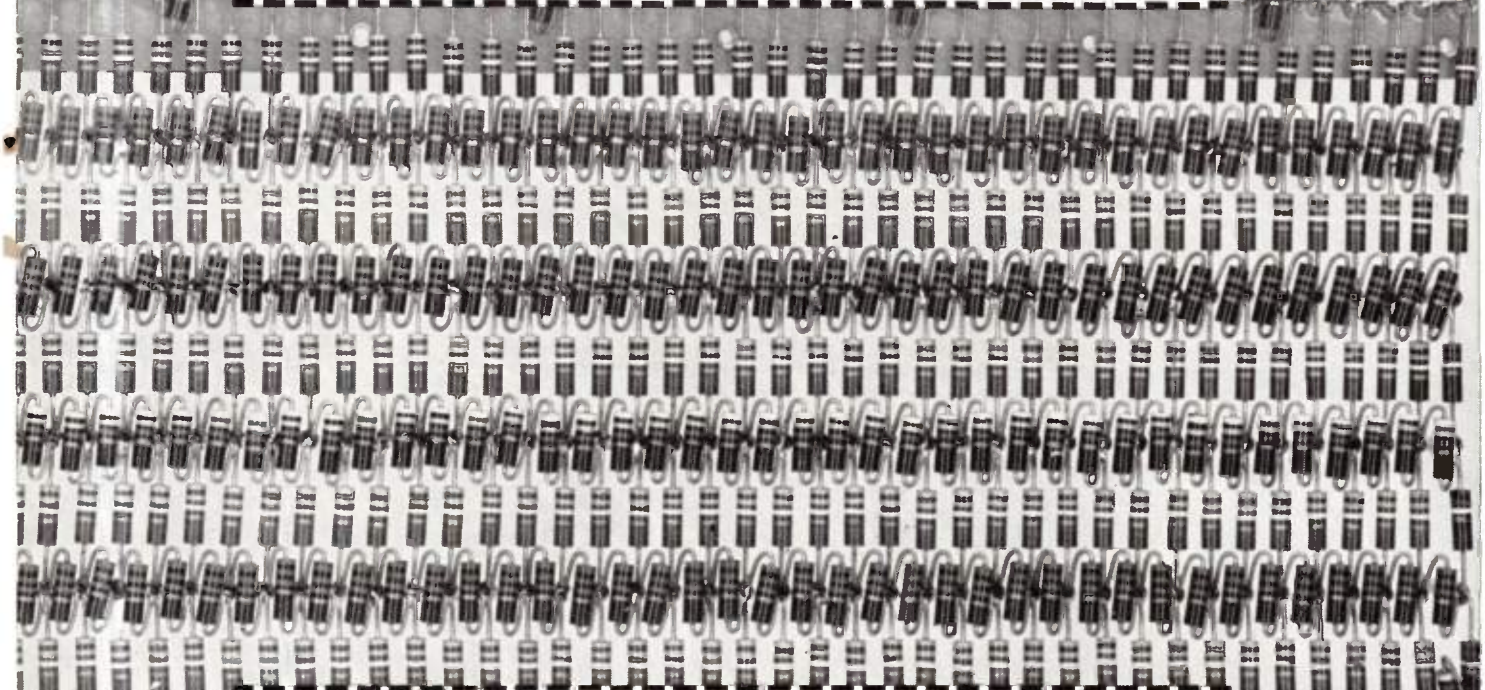
QUALITY **ELECTRONIC COMPONENTS**

This complex resistor network at Schlumberger's Research Center contains in its basic assembly some 150,000 of the 500,000 Allen-Bradley hot molded resistors which they have assembled into grids simulating earth formations. By inserting interchangeable grids into the network in various combinations, it is possible to simulate the borehole and formation parameters which affect resistivity measurements. Duplication of formation characteristics permits a more precise examination and interpretation of the different resistivity logs used in locating potential gas and oil producing zones.

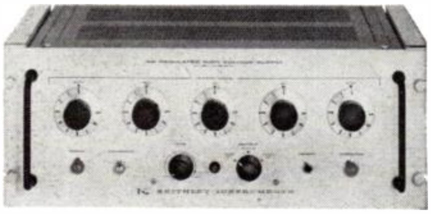




ALLEN-BRADLEY HOT MOLDED RESISTORS are available in all standard EIA and MIL-R-11 resistance values and tolerances. Shown actual size from left to right:

Type TR 1/10 watt (MIL Type RC 06), Type CB 1/4 watt (MIL Type RC 07), Type EB 1/2 watt (MIL Type RC 20), Type GB 1 watt (MIL Type RC 32), Type HB 2 watts (MIL Type RC 42).

A-B also makes a quality line of hermetically sealed precision resistors. Using metal grid construction, they are noninductive. Ratings are 1/4, 1/2 and 1 watt at 125° C with tolerances of 0.1, 0.25, 0.5 and 1.0%; and TC ±25 PPM.



IRE People

(Continued from page 63A)

view, N. Y. He obtained the B.S.E.E. degree in 1952 from the College of the City of New York and M.S.E.E. degree 1957 from the Polytechnic Institute of Brooklyn.

❖

P. Nooteboom

Appointment of **Peter Nooteboom** (M'61) as sales engineer in the Los Angeles District office of Cain & Co., nationwide instrumentation marketing firm, was announced recently. Most recently associated with the Electrada Corporation as sales engineer, he has also held key sales, engineering and administrative posts with Wright Engineering Company, Stratham Instruments Inc., and Hughes Aircraft.

A mathematics graduate of UCLA, Nooteboom served as a pilot in the U. S. Coast Guard during World War II. He is a member of ISA and the Institute of Environmental Sciences.

❖

J. W. Moyer

Dr. James W. Moyer (SM'56) has joined Servomechanisms/Inc. as Director of Research of the Company's Research and Development Center in Santa Barbara, Calif. As Director of the Research Group, he will coordinate his activity with the Center's development group.

Before joining SM/I, he served as Director of Research at the Sperry Rand Research Center, Sudbury, Mass., where he organized an initial program for basic and applied research.

Earlier he was associated with the General Electric Company for 12 years, where his work included studies on liquid metals properties, thermionics, hydrogen isotope separation and analysis, radiation effects in semiconductors and the first direct conversion of fission energy to electricity at the Knolls Atomic Power Laboratory. He served as physicist and consulting engineer at the G. E. Microwave Laboratory in Palo Alto, Calif., later joining G. E.'s Technical Military Planning Operation (TEMPO) as staff physicist and manager of the physical sciences in the environment operation. At TEMPO, he was concerned with analysis of trends in basic research and applications to the broad military spectrum and the emphasis on interdisciplinary activities between all the sciences. From 1941 to 1946, he worked with the Rochester Ordnance District and on the Manhattan Project at Berkeley and Oak Ridge. He acted as consultant to the National Bureau of Standards from 1956 to 1959, where he helped organize the Free Radicals Research Project.

Dr. Moyer received the A.B. degree in physics from Cornell University in 1938. He spent three years in physics at Rutgers University, from 1938 to 1941. He received the Ph.D. degree in physics from the University of Rochester in 1948. His professional affiliations include membership in the American Physical Society, American Association for the Advancement of Science, American Rocket Society, and Sigma Xi Philosophical Society of Washington.

❖

The appointment of **Dr. Henry M. O'Bryan** (M'51–SM'54) as scientific assistant to the vice president of engineering and research of The Bendix Corporation was announced recently. He will coordinate Bendix engineering and research activities with scientists, engineers and government agencies in the Washington, D. C. area.

For 10 years prior to joining the Bendix staff he was with the General Telephone and Electronics Corporation and its subsidiary, Sylvania Electric Products. He also has been on the staffs of the New Jersey Zinc Company, Reeves Sound, and Baird Associates. In recent years he was a professor of industrial management at Brooklyn Polytechnic Institute. He also has taught physics at Johns Hopkins University, Massachusetts Institute of Technology, Harvard University, and Georgetown University. He was educated as a physicist at Clark, Northwestern, and Johns Hopkins universities.

Dr. O'Bryan is a licensed professional engineer in the state of New York, a member of the American Institute of Electrical Engineers, the Optical Society of America, and The Institute of Management Sciences, and a fellow of the American Physical Society.

❖

L. J. Perenic

Lawrence J. Perenic (SM '52) has joined S-F-D laboratories as senior project engineer, sub-systems. He will be engaged primarily in the development of microwave sub-systems. He was formerly associated with ACF Electronics Division of ACF Industries, Inc., Paramus, N. J., as director of the microwave equipment laboratory. His work there included the development of solid-state tracking beacons as well as related command and telemetry sub-systems for the Discoverer, Mercury and Minuteman missile programs. He received his degree in Electrical Engineering from Brooklyn Polytechnic Institute in 1943, where he later taught in the evening school.

❖

Aaron Rosenfeld (S'49–A'51–M'57) has been appointed President of Micro-Power, Inc., 20-21 Steinway Street, Long Island City, N. Y., manufacturer of a comprehensive line of catalog power supply units designed specifically for microwave tubes. He will direct the engineering and manufacture of Micro-Power's products. He was formerly Chief Engineer for Kepco, Inc., where he was responsible for the formulation and design of their line of power regulators.

A. ROSENFELD

Mr. Rosenfeld received the B.S. degree in engineering from the City College of New York.

❖

Walter E. Rudd (M'61), formerly with Electro-Instruments, Inc., has been named Northwest district sales manager of Dunn Engineering Corporation. He will make his headquarters in Sunnyvale, Calif., and will direct sales in Northern California, Oregon and Washington.

W. E. RUDD

Prior to serving as district sales manager for Electro-Instruments, he was a research engineer with the Lockheed Missile and Space Division and the Autonetics Division of the North American Aviation, Inc. He was in the far East from 1950 to 1954 as an electronics technician in the Navy. Mr. Rudd was educated in Fullerton Junior College, Santa Ana College and Howard Payne College.

❖

Dr. Morton B. Prince (SM '55), former corporate vice president of Hoffman Electronics, has joined the staff of Electro-Optical Systems, Inc., as Manager of Microelectronics. One of the country's leading authorities in solar cells and solid state electronics, he will be responsible for the development of a broad range of microelectronic programs at Electro-Optical Systems. These programs will be both of a research and development nature as well as directed toward eventual commercial utilization and manufacture. In addition to his primary functions in the microelectronics area, he will also be active in the company's programs in advanced space power systems using photovoltaic solar cells.

Prior to joining Electro-Optical Systems, he was Manager of Hoffman's Semiconductor Division, where he was responsible for over-all administrative, technical, marketing, manufacturing, and financial operation of that group. In this position, he directed the Division's activities toward strong research and de-

(*Continued on page 68A*)

IRE People

(Continued from page 67A)

velopment efforts which resulted in the development of three new product lines.

He was also a member of the technical staff of the Bell Telephone Laboratories, where he performed extensive work in the properties of semiconductor materials and semiconductor devices. He was responsible for the development of the solar energy converter and an extremely efficient silicon power rectifier.

Dr. Prince received the Ph.D. degree in physics in 1951 from the Massachusetts Institute of Technology. He is a member of the American Physical Society, Sigma Xi, the Association for Applied Solar Energy, and the Electrochemical Society.

❖

W. P. Schneider

The appointment of **William P. Schneider** (M'55) as manager of engineering for Electrodynamic Instrument Corporation, Houston, Tex. was announced recently. He will direct the firm's engineering research and design programs in the fields of geophysical instrumentation, electronic components and subsystems, and oceanographic instrumentation. He was formerly a project development engineer at Schlumberger Well Surveying Corporation.

He holds both a B.S. degree in mathematics and a B.S. degree in electrical engineering from the University of Houston and a B.S. degree in electrical engineering from Massachusetts Institute of Technology. He was named the outstanding student in the school of engineering while attending the University of Houston, and was awarded a graduate fellowship at MIT for "high scholastic attainment and promise of professional competence." He taught at the University of Houston after graduating from MIT, then returned to MIT to join its research staff, where he worked on Project Lincoln.

Mr. Schneider is a member of Phi Kappa Phi, Sigma Xi, and the Engineers Council of Houston.

❖

Dr. Jacob Shekel (A'52–M'55–SM'57) for the last four years head of the electronics division of the Israel Ministry of Defense, has joined the development staff of Spencer-Kennedy Laboratories, Inc., Boston, Mass. Dr. Shekel was with SKL during his doctoral studies at M.I.T. and rejoins the organization as Director of Instrument Development. He is a former Chairman (1960) of the Israel section of the IRE.

❖

The appointment of **Donald O. Schwennesen** (A'43–M'55) as Vice President in charge of the Pacific Division of The Arnold Engineering Company was announced recently. In this newly created position Mr. Schwennesen will make his headquarters at Arnold's manufacturing plant at Fullerton, Calif. He was formerly a Vice President of Magnetic Metals Company, Camden, N. J.


D. O. Schwennesen

He is a graduate of Illinois Institute of Technology, Chicago, formerly Armour Institute of Technology, with a Degree in Electrical Engineering, and has done post-graduate study at Armour Institute of Technology and at Northwestern University. From 1935 until 1955, he was employed by Chicago Standard Transformer Company, working his way up from Laboratory Engineer to Vice President of Engineering and Sales. Mr. Schwennesen is a member of the American Institute of Electrical Engineers.

⁂

The advancement of **William P. Shvidrik** (S'57–M'61) to technical director of Hayden Advertising Inc., Montclair, N. J. was announced recently. He joined the agency a year ago as product information adviser and writer in the fields of electronics, nondestructive testing, industrial process controls, metallurgy, microwave equipment and antenna systems. His new post will involve technical copy supervision as well as product research for the agency's industrial and specialty consumer accounts. He came to Hayden from ITT Federal Laboratories, Clifton, N. J. As an electrical engineer in the reliability group there, he evaluated components developed for ITT's airborne communications equipment, including high-frequency power tubes, cavities, motors and controls used in DMET and TACAN programs, the advanced navigational aids in use today by both civilian and military aircraft.


W. P. Shvidrik

This assignment followed four years in the U. S. Navy, where he served as aviation electronics technician, stationed in Frankfort, Germany, and at Port Lyautey, in French Morocco. Earlier, he attended the Navy's "Annapolis of the Air," the Aviation Electronics School at Memphis, Tenn. He has also served as electronic technican for Gawler-Knoop Co., electronic sales engineers of Roseland, N. J. He is a graduate of Newark College of Engineering and Bound Brook High School, where he was a member of its Honor Society.

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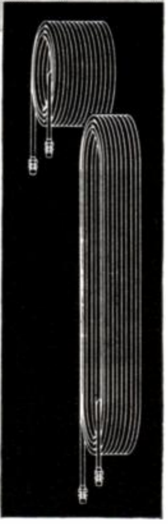


IRE People

(Continued from page 69A)

The General Microwave Corporation announces the appointment of **Leonard O. Sweet** (S'51–A'53–M'58) to the post of Chief Microwave Engineer. In this capacity, he is directly responsible for all aspects of microwave engineering effort, encompassing both external and company-sponsored programs.

L. O. Sweet

He received the degree of Bachelor of Electrical Engineering in 1951 from the College of the City of New York and the degree of Master of Electrical Engineering from the Polytechnic Institute of Brooklyn in 1952. He is currently continuing his studies towards a doctoral degree in Electro-Physics at the Polytechnic Institute of Brooklyn. From 1951 to 1953 and 1955 to 1956, he was employed at the Microwave Research Institute of the Polytechnic Institute of Brooklyn as a Research Associate. His work dealt primarily with the evaluation of existing methods of microwave power measurement and the development of absolute calorimetric standards for the measurement of such power. Prior to joining General Microwave Corporation, he was employed at PRD Electronics, Inc. from 1956 to 1961, where he headed the Microwave Section responsible for all microwave special order work and product development for catalog sales. From 1953 to 1955, he served as a First Lieutenant in the United States Air Force at Eglin Air Force Base, Florida. He conducted flight tests of high speed bombing and navigation computers and coordinated efforts of the numerous support personnel involved in such a project.

Mr. Sweet is a member of Sigma Xi.

⁂

The Antenna and Microwave Department of TRG Inc., Somerville, Mass., announces the formation of TRG-West, a new division to be located in the Stanford area of California This division will specialize in applied electromagnetic theory, antennas, microwave components and quantum electronics. **Dr. Robert L. Tanner** (S'51–A'52–SM'56) has been appointed Vice President of the division, and **Dr. Edward M. T. Jones** (S'45–A'51–M'55–SM'56) will be the Director of Research.

Dr. Tanner was formerly Head of the Electromagnetic Sciences Group at the Electromagnetics Laboratory of Stanford Research Institute where his interests embraced a number of areas in the field of antennas and related systems problems. Dr. Jones also came from the Stanford Research Institute where he was Head of the Microwave Group of the Electromagnetics Laboratory. His fields of specialty include both low and high power microwave filters, microwave antennas, strip-line components, directional couplers and parametric amplifiers.

Both are members of Sigma Xi and the Scientific Research Society of America.

⁂

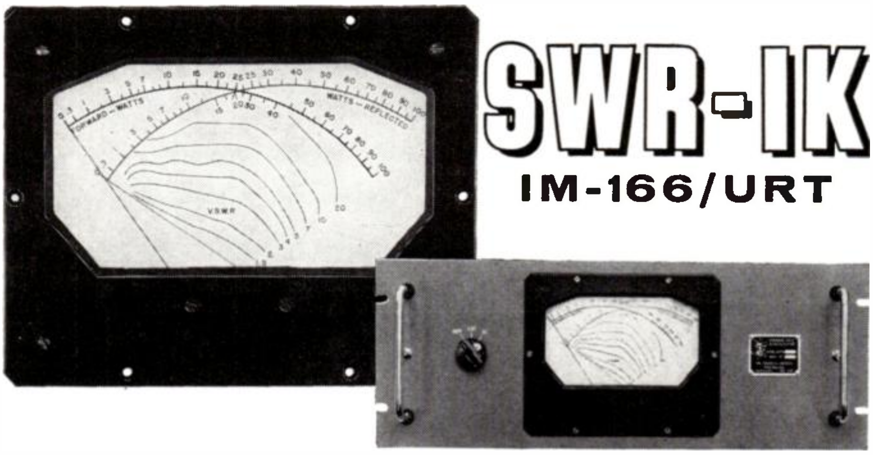
The appointment of **James R. Weiner** (S'41–A'47–SM'54) as associated manager of the guidance and controls division of Hughes Aircraft Company's aerospace group was announced recently. He was formerly vice president of engineering for Philco Corporation's government and industrial group in Philadelphia, where he was responsible for the coordination of the engineering and development work of the five divisions within the group.

J. R. Weiner

Mr. Weiner received his bachelor of science degree in engineering physics from the University of Illinois and his master of electrical engineering degree from Polytechnic Institute of Brooklyn, and did further graduate work at Massachusetts Institute of Technology and Harvard University. He is a member of numerous technical societies and has served as a consultant for the Department of Defense.

⁂

(Continued on page 72A)



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. . . for single-shot or recurrent or triggered main-sweep presentations . . . for either conventional or triggered jitter-free delayed-sweep presentations.

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. . . from a company that has specialized in manufacturing ONLY laboratory oscilloscopes and associated instrumentation.

. . . from a company that has emphasized quality in design for quality in performance for 15 years.

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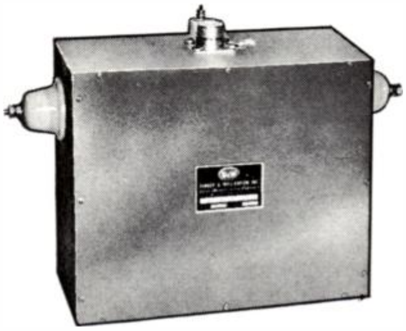

IRE People

(Continued from page 70A)

Dr. S. Dean Wanlass (S'44–A'53–M'58), one of the nation's top electronics scientists, has been elected president of Packard Bell Computer Corporation, it was announced recently. In addition to his new post, he will continue to serve as group vice president—Defense and Industrial Group of the parent Packard Bell Electronics Corporation.

One of the other new officers named by the board was **Kenneth R. Jackson** (M'46–SM'54) vice president and director of administration.

⁂

William J. Weisz, (S'48–A'49–M'53–SM'60) Manager of Mobile and Portable Communications Products in the Communications Division, has been elected Vice President, Communications Products of Motorola, Inc., Communications Division. A 1948 graduate of Massachusetts Institute of Technology in Electrical Engineering, he joined Motorola that year and was appointed Manager of the Subminiature Equipment Engineering Section in 1952. He became Product Manager for Portable Communications products in 1954, Chief Engineer for two-way mobile and portable products in 1956, and Manager of that department in 1958.

Mr. Weisz is a member, American Institute of Electrical Engineers and has represented Motorola on various committees of the Electronics Industry Association.

⁂

Industrial Engineering Notes

(Continued from page 38A)

empirical data in order to provide scientists with primary information applicable to immediate problems. Target time for information retrieval would be on the order of minutes for simple questions, ranging upward to a few hours for more complex searches. Background, "fill-in" information chores would be left to existing libraries. A report describing the proposed system, "Considerations and Recommendations for Developing a Materials Information Processing Capability," by Belfour Engineering Co. for Wright Air Development Division is available from OTS, U. S. Department of Commerce, Washington 25, D. C. Price 50 cents. Order AD 258 643. . . . **A new OTS Selective Bibliography listing U. S. government and military research reports, translations, and other technical documents on data processing and programming has been published by the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C. The bibliography, which may be purchased from OTS for 10 cents is: SB-474 "Data Processing and Programming."**

Section Meetings

AKRON

"Machines that Think & People that Don't," J. E. Steele, Wright-Patterson AFB; 9/19/61.

ALAMOGORDO-HOLLOMAN

"The 1000 Foot Spherical Reflector Antenna at Arecibo, Puerto Rico," B. G. Hagaman, Deco Electronics, Joint with American Rocket Society; 10/17/61.

ALBUQUERQUE-LOS ALAMOS

"On Computers," E. A. Voorhees, Los Alamos Scientific Lab.; "History of Los Alamos," Peggy P. Church, Author; Tour of "Stretch" Computer Facilities; 10/20/61.

ANCHORAGE

Talk by Ted Young, Chairman, on his trip to IRE Headquarters; 10/2/61.

ATLANTA

"Electronics in Medicine," H. L. Williams, Emory University; 10/27/61.

BAY OF QUINTE

"Radio Telescopes," R. M. Chisholm, Queen's University, Kingston, Ont.; 10/18/61.

BEAUMONT-PORT ARTHUR

"RF Interference," E. Coburn, U.S. Government; "Dynamic Resistance Test of Spot Welds," L. Cherry, Lamar College; 10/10/61.

"Modern Laboratory Oscilloscopes," R. D. Patterson, Tektronix, Inc.; 11/7/61.

BENELUX

Trip to Radioastronomy Station at Humain, Belgium, R. J. Coutrez, Observatoire Royal de Belgique; 10/27/61.

"Realisabilite des filtres sans inductance mutuelle," J. Meinguet, University of Louvain; "Basic Network Elements for the Synthesis of Non-Linear Systems," S. Duinker, Philips Research Lab.; "Contributions to Crystal Filter Design," M. A. Fettweis, Bell Telephone; "Circuits Electriques a Parametres Complexes," M. O. Beaufays, University of Bruxelles; 10/28/61.

BINGHAMTON

"Optical Masers," P. P. Kisliuk, Bell Telephone Labs.; 9/11/61.

"Communication Satellites," P. H. Sawitz, Institute for Defense Analyses; 10/16/61.

BUFFALO-NIAGARA

Tour of new Niagara Power Project; 9/23/61.

"Field Effect Transistors," L. J. Sevin, Texas Instruments; 10/18/61.

CHICAGO

"Electromagnetic Compatibility Analysis Center," S. Cohn, Director, Technical Operations, Annapolis; 11/10/61.

CINCINNATI

"A Reverberation Scheme For Baldwin Electronic Organ," A. Bissonette, Baldwin Piano Co.; 9/19/61.

"The Laser and Quantum Electronics," L. Seidenstein, Avco Corp.; 10/17/61.

COLOMBIA

"Measurement of Radiation," R. Yusti, Instito Colombiano Asuntos Nucleares-Bogota; 7/14/61.

"Tunnel Diode," W. Westphal, Universidad Javeriana; 11/9/61.

(Continued on page 74A)

CANNON SUBMINIATURE GOLDEN-D MARK II

Golden-D Mark II Subminiature, the crimp snap-in version of the famous "D" line, is available for immediate delivery! ■ MATES WITH ORIGINAL D ■ MONOBLOC INSULATORS ■ SNAP-IN REMOVABLE CRIMP CONTACTS DESIGNED TO MS 3190 ■ CADMIUM PLATED, GOLD IRIDITE FINISH SHELLS ■ FIVE SIZES WITH 9, 15, 25, 37, and 50 CONTACTS ■ PROBE-PROOF CLOSED ENTRY SOCKET CONTACTS ■ LOW ASSEMBLY COST FOR THE USER. Wherever maximum reliability is needed in a subminiature multicontact plug—for both military and industrial applications—specify the Golden-D Mark II...*it is engineered to deliver superior performance by the world's most experienced manufacturer of electrical connectors.* Write for D-8 Catalog.

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Section Meetings

(Continued from page 72A)

COLUMBUS

"Micrologic Elements," R. Schulz and R. Major, Fairchild Semiconductor Corp.; 10/19/61.

DALLAS

"Some Color Slide and Color Television Experiments Using the Land Technique," W. L. Hughes, Oklahoma State University; 10/31/61.

DENVER

A report on the design and test of extra-high voltage lines at high altitudes and the final construction of the first 240,000 volt line in the Rocky Mountain area, Mr. Robertson and Mr. Farnsworth, Public Service Co. of Colorado; 10/13/61.

ELMIRA-CORNING

"Getters," P. D. Porta, SAES, Italy; 10/9/61.

"Thin Film Circuitry," D. McLean, Bell Telephone Labs.; 11/13/61.

EVANSVILLE-OWENSBORO

Tour of Dewey & Almy Chemical Plant; 11/8/61.

FLORIDA WEST COAST

"Low-noise Microwave Parametric Amplifiers," G. P. Rodrigue, Sperry Microwave Electronics Co.; 10/18/61.

"Graduate Technical Education in the Golden Triangle—What Lies Ahead," panel discussion moderated by J. S. Allen, Panel members—M. Blee, Board of Control, L. E. Grinter, Univ. of Florida, J. H. Leutwiler, Sperry Microwave Elect. Co., S. D. Sanson, N. K. Winston-Sanson Florida Corp., J. Weil, University of Florida; 11/8/61.

FORT HUACHUCA

"Area Transforms," W. Brown, University of Michigan; 10/23/61.

FORT WORTH

"Thin Film Circuits Ready for Production," W. O. Faith, Varo, Inc.; 10/10/61.

GAINESVILLE

"Parametric Amplifiers," W. H. Louisell, Bell Telephone Labs.; 11/8/61.

HAMILTON

"Semiconductors—State of the Art," L. J. Sevin, Texas Instruments; 10/18/61.

"Electronic Engineers Stake in Canada," G. P. Adamson, Cdn. Westinghouse Co.; 11/6/61.

HAWAII

"Reliability in Systems," A. Graps, Jr., Quality Evaluation Lab. USN; 8/9/61.

"Compatible Stereo-Multiplex," D. H. Dashiell, Bureau of Ships, USN; 9/13/61.

"Airglow Zodiacal Light Studies on Haleakala," W. R. Steiger, University of Hawaii; 10/11/61.

"Recent Developments in Aerial Surveillance, Air Radar, Drone System, Air Infra-Red and Air Photo," J. A. Godbey & D. Coen, Schefield Barracks, U.S. Army; 11/8/61.

HOUSTON

"Electronic Seismometers," L. Erath, Erath, Inc.; 9/20/61.

"Quantized Flux in Superconductors," H. E. Rorschach, Rice University; 10/17/61.

HUNTSVILLE

"Pert," H. A. Matheny, Brown Engineering Co.; 10/26/61.

INDIANAPOLIS

Tour of Teleonic Industries, Inc. Plant facility, L. Abbott, Pres.; 10/26/61.

ISRAEL

"Stabilized Satellite Equipped with Telescope Television Camera and Computer for Ultraviolet Astronomical Observation," E. Arazi, M.I.T. & Smithsonian Astrophysical Observation; 8/24/61.

ITHACA

"Radio Astronomy at the Paris Observatory," M. H. Cohen, Cornell University; 11/9/61.

KANSAS CITY

"Tunnel Diode Applications," E. Gottlieb, General Electric; 10/12/61.

KITCHENER-WATERLOO

"Exploring Space with Radio Waves," P. M. Millman, National Research Council; 10/16/61.

LONG ISLAND

"Fundamental Characteristics of Hydrofoil Craft," J. K. Roper, Grumman Aircraft Engrg. Corp., Movie: "The Americas' Cup Races"; 9/12/61.

"Introduction to Molecular Electronic Engineering," A. von Hippel, Mass. Institute of Technology; 10/19/61.

"Analysis and Synthesis of Molecular Systems," T. D. Callinan, IBM; 10/26/61.

"Molecular Electronics—A Key to Reliability," H. V. Noble, Wright Air Devel. Div.; 11/2/61.

LOS ANGELES

"Can Our Electronic Engineering Resources Meet the Future Challenge?" J. M. Bridges, Dept. of Defense, Joint meeting with Los Angeles Chapter PGMIL; 10/20/61.

MILWAUKEE

"Biomedical Electronics," S. Larks, Marquette University; 10/17/61.

MOBILE

"Computer Programming For Engineers," J. Jones, IBM Corp; 6/30/61.

Film—Air Force Missiles' Mission; 9/29/61.

MONTREAL

Tour of Laval Institute; 9/20/61.

NEW YORK

"Communications and Trailing Aspects of Project Mercury," R. M. Goetchius; 11/1/61.

NORTHERN ALBERTA

"Some New Electronic Tubes, Semiconductors and Components and Educational Aids," D. S. Simkins, Philips Electronics Industries Limited; 10/24/61.

NORTHERN NEW JERSEY

"How to Use Computers," W. Fazar, U. S. Dept. of Navy; 10/11/61.

OKLAHOMA CITY

"Revolution in Technology," L. Berkner, President of IRE; 10/23/61.

"Analog Transformations," C. E. Harp, University of Oklahoma; 11/13/61.

OTTAWA

"The New Ottawa Labs. of the Northern Electric Co.," R. E. Tanner, Northern Electric Co. after dinner talk, J. C. R. Punchard, Fellow IRE; 11/9/61.

(Continued on page 78A)

29 Fields of Special Interest-

The 29 Professional Groups are listed below, together with a brief definition of each, the name of

Aerospace and Navigational Electronics

Annual fee: $3.

The application of electronics to operation and traffic control of aircraft and to navigation of all craft.

Mr. George M. Kirkpatrick, Chairman, General Electric Co., Syracuse, N.Y.

40 Transactions, *6, & *9, Vol 1, No. 3; Vol. 2, No. 1-3; Vol 4, No. 1, 2, 3; Vol, 5, No. 2, 3, 4; Vol. 6, No. 1, 2, 4; Vol. 7, No. 1, 2, 3, 4; Vol. 8, No. 1, 2.

Antennas and Propagation

Annual fee: $6.

Technical advances in antennas and wave propagation theory and the utilization of techniques or products of this field.

Dr. Harry Fine, Chairman, Applied Propagation Branch FCC, Washington, D.C.

38 Transactions, *Vol. AP-2, No. 2; AP-4, No. 4; AP-5, No. 1-4; AP-6, No. 1, 2, 3, 4; AP-7, No. 1, 2, 3, 4; AP-8, No. 1, 2, 3, 4, 5, 6; AP-9, No. 1, 2, 3, 4, 5.

Audio

Annual fee: $2.

Technology of communication at audio frequencies and of the audio portion of radio frequency systems, including acoustic terminations, recording and reproduction.

Prof. Cyril M. Harris, Chairman, Columbia University, 632 W. 125 St., New York 27, N.Y.

60 Transactions, *Vol. AU-1, No. 6; *Vol. AU-2, No. 4; Vol. AU-3, No. 1, 3, 5; Vol. AU-4, No. 1, 5-6; Vol. AU-5, No. 1, 2, 3, 4, 5, 6; AU-6, No. 1, 2, 3, 4, 5, 6; AU-7, No. 1, 2, 3, 4, 5, 6; AU-8, No. 1, 2, 3, 4, 5, 6; AU-9, No. 1, 2, 3, 4.

Automatic Control

Annual fee: $3.

The theory and application of automatic control techniques including feedback control systems.

Mr. John M. Salzer, Chairman, 909 Berkeley St., Santa Monica, Calif.

14 Transactions, PGAC-3-4-5-6, AC-4, No. 1, 2, 3; AC-5, No. 1, 2, 4; AC-6, No. 1, 2, 3.

Bio-Medical Electronics

Annual fee: $3.

The use of electronic theory and techniques in problems of medicine and biology.

Mr. George N. Webb, Chairman, Dept. of Med. Biophysical Div., Johns Hopkins Hospital, Baltimore 5, Md.

21 Transactions, 8, 9, 11, 12; ME-6, No. 1, 2, 3, 4; ME-7, No. 2, 3, 4; BME-8, No. 1, 2, 3.

Broadcast & Television Receivers

Annual fee: $4.

The design and manufacture of broadcast and television receivers and components and activities related thereto.

Mr. Robert R. Thalner, Chairman, Sylvania Home Electronics, Batavia, N.Y.

29 Transactions, *7, 8; BTR-1, No. 1-3, BTR-2, No. 1-2-3; BTR-3, No. 1-2; BTR-4, No. 2, 3-4; BTR-5, No. 1, 2; BTR-6, No. 1, 2, 3; BTR-7, No. 1, 2.

Broadcasting

Annual fee: $2.

Broadcast transmission systems engineering, including the design and utilization of broadcast equipment.

Mr. Raymond F. Guy, Chairman, 264 Franklin St., Haworth, N.J.

20 Transactions, No. 10, 11, 12, 13, 14; BC-6, No. 1, 2, 3,; BC-7, No. 1, 2, 3

Circuit Theory

Annual fee: $4.

Design and theory of operation of circuits for use in radio and electronic equipment.

Dr. James H. Mulligan, Jr., Chairman, College of Eng., New York University, New York 53, N.Y.

31 Transactions, CT-4, No. 3-4; CT-5, No. 1, 2, 3, 4; CT-6, No. 1, 2, 3, 4; CT-7, No. 2, 3, 4; CT-8, No. 1, 2.

Communications Systems

Annual fee: $2.

Radio and wire telephone, telegraph and facsimile in marine, aeronautical, radio-relay, coaxial cable and fixed station services.

Mr. Ralph L. Marks, Chairman, Rome Air Dev. Center, Griffiss AFB, N.Y.

22 Transactions, CS-5, No. 2, 3; CS-6, No. 1, 2; CS-7, No. 1, 3, 4; CS-8, No. 1, 2, 3, 4; CS-9, No. 1, 2.

Component Parts

Annual fee: $3.

The characteristics, limitation, applications, development, performance and reliability of component parts.

Mr. Floyd E. Wenger, Chairman, Headquarters ARDC, Andrews AFB, Washington 25, D.C.

24 Transactions, CP-4, No. 1, 2, 3, 4; CP-5, No. 1, 2, 3, 4; CP-6, No. 1, 2, 3, 4; CP-7, No. 1, 2, 3, 4; CP-8, No. 1, 2.

Education

Annual fee: $3.

To foster improved relations between the electronic and affiliated industries and schools, colleges, and universities.

Mr. George E. Moore, Chairman, Westinghouse Electric Corp., East Pittsburgh, Pa.

14 Transactions, Vol. E-1, No. 3, 4; E-2, No. 1, 2, 3, 4; E-3, No. 1, 2, 3, 4; E-4, No. 1, 2.

Electron Devices

Annual fee: $3.

Electron devices, including particularly electron tubes and solid state devices.

Mr. Willis A. Adcock, Chairman, Texas Instruments Co., Dallas 9, Tex.

34 Transactions, *Vol. ED-1, No. 3, 4; ED-3, No. 2, 4; ED-4, No. 2, 3, 4; ED-5, No. 2, 3, 4; ED-6, No. 1, 3; ED-7, No. 2, 3, 4; ED-8, No. 1, 2, 3, 4.

Electronic Computers

Annual fee: $4.

Design and operation of electronic computers.

Dr. A. A. Cohen, Chairman, Remington-Rand Univac, St. Paul 16, Minn.

39 Transactions, EC-6, No. 2, 3; EC-7, No. 1, 2, 3, 4; EC-8, No. 1, 2, 3, 4; EC-9, No. 1, 2, 3, 4; EC-10, No. 1, 2, 3.

Engineering Management

Annual fee: $3.

Engineering management and administration as applied to technical, industrial and educational activities in the field of electronics.

Mr. T. W. Jarmie, Chairman, Engineered Electronics, 1441 East Chestnut Ave., Santa Ana, Calif.

22 Transactions, EM-3, No. 2; EM-4, No. 1, 3, 4; EM-5, No. 1-4; EM-6, No. 1, 2, 3; EM-7, No. 1, 2, 3, 4; EM-8, No. 1, 2.

Engineering Writing and Speech

Annual fee: $3.

The promotion, study, development, and improvement of the techniques of preparation, organization, processing, editing, and delivery of any form of information in the electronic-engineering and related fields by and to individuals and groups by means of direct or derived methods of communication.

John M. Kinn, Jr., Chairman, IBM Journal, 545 Madison Ave., New York, N.Y.

9 Transactions, Vol. EWS-1, No. 2; EWS-2, No. 1, 2, 3; EWS-3, No. 1; EWS-4, No. 1, 2.

THE INSTITUTE OF RADIO

IRE's 29 Professional Groups

he group chairman, and publications to date.

* Indicates publications still available

Geoscience Electronics

Annual fee: None

Research and development in electronic instrumentation for geophysics and geochemistry, especially gravity measurements, seismic measurements, magnetics, well-logging, space exploration, meteorology, oceanography, etc.

Human Factors in Electronics

Annual fee: $2.

Development and application of human factors and knowledge germane to the design of electronic equipment.

Mr. Robert R. Riesz, Chairman, Bell Tel. Labs, Murray Hill, N.J.

4 Transactions, HFE-1, No. 1, 2; HFE-2, No. 1, 2.

Industrial Electronics

Annual fee: $3.

Electronics pertaining to control, treatment and measurement, specifically, in industrial processes.

Mr. J. E. Eiselein, Chairman, RCA Victor Div., Camden, N. J.

16 Transactions, *PGIE 1, 3, 5, 6, 7, 8, 9, 10, 11; IE-7, No. 1, 2, 3; IE-8, No. 1, 2.

Information Theory

Annual fee: $4.

The theoretical and experimental aspects of information transmission, processing and utilization.

Dr. George L. Turin, Chairman, Dept. of E.E., Univ. of California, Berkeley 4, Calif.

27 Transactions, PGIT-4, IT-1, No. 3! IT-2, No. 3; IT-3, No. 1, 2, 3, 4; IT-4, No. 1, 2, 3, 4; IT-5, No. 1, 2, 3, 4; IT-6, No. 1, 3, 4, 5; IT-7, No. 1, 2, 3.

Instrumentation

Annual fee: $3.

Measurements and instrumentation utilizing electronic techniques.

Mr. Harvey W. Lance, Chairman, Natl. Bureau of Standards, Boulder, Colo.

20 Transactions, PGI-4, Vol. 1-6, No. 2, 3, 4; Vol. 1-7, No. 1, 2; Vol. 1-8, No. 1, 2; Vol. 1-9, No. 1, 2, 3; Vol. 1-10, No. 1.

Microwave Theory and Techniques

Annual fee: $3.

Microwave theory, microwave circuitry and techniques, microwave measurements and the generation and amplification of microwaves.

Mr. Tore N. Anderson, Chairman, EMT Corp., Syosett, Long Island, N.Y.

37 Transactions, MTT-4, No. 3; MTT-5, No. 3, 4; MTT-6, No. 1, 2, 3, 4; MTT-7, No. 2, 3, 4; MTT-8, No. 1, 2, 3, 4, 5, 6; MTT 9, No. 1, 2, 3, 4.

Military Electronics

Annual fee: $2.

The electronics sciences, systems, activities and services germane to the requirements of the military. Aids other Professional Groups in liaison with the military.

Mr. Willie L. Doxey, Chairman, EC Dept., USASRDL, Fort Monmouth, N.J.

13 Transactions, MIL-1, No. 1; MIL-2, No. 1; MIL-3, No. 2, 3, 4; MIL-4, No. 2-3, 4; MIL-5, No. 1, 2, 3.

Nuclear Science

Annual fee: $3.

Application of electronic techniques and devices to the nuclear field.

Mr. Louis Costrell, Chairman, N.B.S., Washington 25, D.C.

21 Transactions, NS-1, No. 1; NS-4, No. 2; NS-5, No. 1, 2, 3; NS-6, No. 1, 2, 3, 4; NS-7, No. 1, 2-3, 4; NS-8, No. 1, 2, 3.

Product Engineering & Production

Annual fee: $2.

New advances and materials applications for the improvement of production techniques, including automation techniques.

Mr. Alfred R. Gray, Chairman, Astronics of Florida, Inc., Orlando, Fla.

8 Transactions, No. 2-3, 4, 5, 6, PEP-5, No. 1, 2.

Radio Frequency Interference

Annual fee: $2.

Origin, effect, control and measurement of radio frequency interference.

Mr. Harold E. Dinger, Chairman, Naval Research Lab., Washington 25, D.C.

3 Transactions, RFI-1, No. 1, RFI-2, No. 1, RFI-3 No. 1.

Reliability and Quality Control

Annual fee: $3.

Techniques of determining and controlling the quality of electronic parts and equipment during their manufacture.

Mr. L. J. Paddison, Chairman, Sandia Corp., Sandia Base, Albuquerque, N.M.

21 Transactions, *3, 5, 10, 11, 12, 13, 14, 15, RQC-9, No. 1, 2, 3; RQC-10, No. 1, 2.

Space Electronics and Telemetry

Annual fee: $3.

The control of devices and the measurement and recording of data from a remote point by radio.

Mr. Kenneth V. Uglow, Chairman, Electro-Mechanical Research Inc., Sarasota, Fla.

19 Transactions, TRC-1, No. 2-3; TRC-2, No. 1; TRC-3, No. 2, 3; TRC-4, No. 1; SET-5, No. 1, 2, 3, 4; SET-6, No. 1, 2, 3-4; SET-7, No. 1, 2, 3.

Ultrasonics Engineering

Annual fee: $2.

Ultrasonic measurements and communications, including underwater sound, ultrasonic delay lines, and various chemical and industrial ultrasonic devices.

Dr. Vincent Salmon, Chairman, Stanford Research Inst., Menlo Park, Calif.

10 Transactions, PGUE, 5, 6, 7; UE-7, No. 1, 2; UE-8, No. 1.

Vehicular Communications

Annual fee: $3.

Communications problems in the field of land and mobile radio services, such as public safety, public utilities, railroads, commercial and transportation, etc.

Mr. Richard P. Gifford, Chairman, General Electric Co., Lynchburg, Va.

18 Transactions, 5, 8, 9, 10, 11, 12, 13; Vol. VC-9, No. 1, 2, 3; Vol.-10, No. 1, 2

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Section Meetings

(Continued from page 75A)

PHOENIX

Election of Officers; 6/23/61.

"Field Effect Transistor," R. M. Warner, Motorola Semiconductor; 9/15/61.

QUEBEC

"High Speed Analog Simulation of Antenna Arrays," J. Gilbert, Canadian Armament R & D Establishment; 10/10/61.

REGINA

Report on National IRE Convention, H. Calder, Saskatchewan Power Corp., "Developments in Electronic Tubes and Semiconductors," D. S. Simkins, Philips Electronics, Ltd.; 10/19/61.

SACRAMENTO

"Australian Communication Systems," R. C. Gant, Royal Australian Air Force; 10/21/61.

SALT LAKE

"MINUTEMAN Solid Propellant Rocket Engines," C. W. Kennedy, Thiokol Chemical Corp.; 10/13/61.

SAN ANTONIO-AUSTIN

Conference on Telluric and Geomagnetic Field Variations; joint with URSI; 10/23-24-25/61.

SAN DIEGO

"Design Problems of Very Large Antennas," K. H. Lippett, Telecomputing Corp.; 11/1/61.

SHREVEPORT

"Satellite Communications," J. M. Doyle, AT&T; 10/3/61.

SOUTH BEND-MISHAWAKA

Tour of Drewrys Plant, D. Howell and W. Trenerry, Drewrys Limited; 9/28/61.

SOUTHERN ALBERTA

"Some New Electronic Tubes, Semiconductors and Components and Educational Aids," D. S. Simkins, Philips Electronics, Ind., Ltd.; 10/20/61.

SYRACUSE

"Magnetohydrodynamics," D. H. Marquis, General Electric Co.; 10/16/61.

TOLEDO

"The Selection and proper use of the Oscilloscope," C. W. Spencer, Tektronix, Inc.; 10/24/61.

TORONTO

"Infrared Radiation," H. Gush, University of Toronto; 11/13/61.

TUCSON

"Army Spectrum Signatures," E. Rivera, Bell Aerosystems Co.; 9/20/61.

TULSA

"The Possibility of Simplifying Color TV Transmission," W. L. Hughes, School of Electrical Engineering; 10/12/61.

"Radioactive Fallout—Its Chemical Composition, Propagation, Safeguards, etc.," J. West, Oklahoma State University; 11/9/61.

VANCOUVER

"Space Communications," Dr. Berkner, President, IRE; 10/12/61.

VIRGINIA

"Analog Computer Applications," L. J. Lane, General Electric Co.; 10/6/61.

"F.M. Stereophonic Broadcasting," C. G. Eilers, Zenith Radio Corp.; 10/20/61.

"Communications for Mercury Capsule," L. Leopold, NASA; 10/27/61.

WESTERN MASSACHUSETTS

"Stereo Hi-Fi and Stereo F-M," D. R. VonRecklinghausen, H. H. Scott, Inc.; 9/26/61.

Proposed Merger AIEE and IRE, moderator —R. H. Shupe; 10/30/61.

SUBSECTIONS

CATSKILL

"The Role of Electronics in Highways and Autos of the Future," G. W. Gray, RCA Labs.; 10/25/61.

EASTERN NORTH CAROLINA

"Direct Distance Dialing," C. A. Jolly, Southern Bell Tel. & Tel. Co.; 10/13/61.

LANCASTER

"Air Surveillance and Required Communications for Air Traffic Control," D. W. Tisdale, Federal Aviation Agency; 10/18/61.

MID-HUDSON

"Thin Film Memories," F. L. Post, IBM Components Div.; 10/17/61.

MONMOUTH

"Micron Size Field Emission Devices," L. N. Heynick, U. S. Army; 10/18/61

PIKES PEAK

"Hearing and Hearing Experiments," J. Victoreen, Victoreen Labs.; 10/18/61.

VICTORIA

"Fun and Frolic with Fearsome Filters," R. Weir, Defence Research Board; 9/27/61.

"Advances in Space Science," G. Odgers, Dominion Astrophysical Observatory; 10/25/61.

WESTERN NORTH CAROLINA

"Survival In The Atomic Age," K. Williams, City of Charlotte and Mecklenburg Co.; 10/27/61.

Membership

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

Transfer to Senior Member

Ball, L. F., Oadby, Leicester, England
Benjamin, C. E., Pittsburgh, Pa.
Chimera, A. J., Buffalo, N. Y.
Cohn, M., Timonium, Md.
Egger, R. J., Detroit, Mich.
Lambert, J. S., Albuquerque, N. M.
Levine, S., Westport, Conn.
Radsliff, J. L., Topeka, Kans.
Sites, F. J., Eglin AFB, Fla.

Admission to Senior Member

Alfred, D. L., Chatham, N. J.
Bigbie, C. R., Thousand Oaks, Calif.
Block, T. S., Oceanside, L. I., N. Y.
Bowers, K. D., Murray Hill, N. J.
Clapper, G. L., Vestal, N. Y.
Deapen, J. A., Jr., Cleburne, Tex.
Debacker, T. A., Bethesda, Md.
Dougherty, J. J., Silver Spring, Md.

Endresen, K., Lillestrom, Norway
Fischer, F. C., Westminster, Calif.
Foote, E. W., Santa Monica, Calif.
Green, J. H., Jr., Richardson, Tex.
Huber, F., Valley Stream, L. I., N. Y.
Kawiecki, C. J., Chicago, Ill.
LaVan, J. T., Philadelphia, Pa.
Lienesch, J. P., Foxboro, Mass.
Nankervis, B. J., Lake Jackson, Tex.
Pierstorff, B. C., Sudbury, Mass.
Stemmer, C. P., Little Neck, L. I., N. Y.
Warchol, E. J., Portland, Ore.
Wilson, L. E., Laurel, Md.
Wist, A. O., Great Neck, L. I., N. Y.

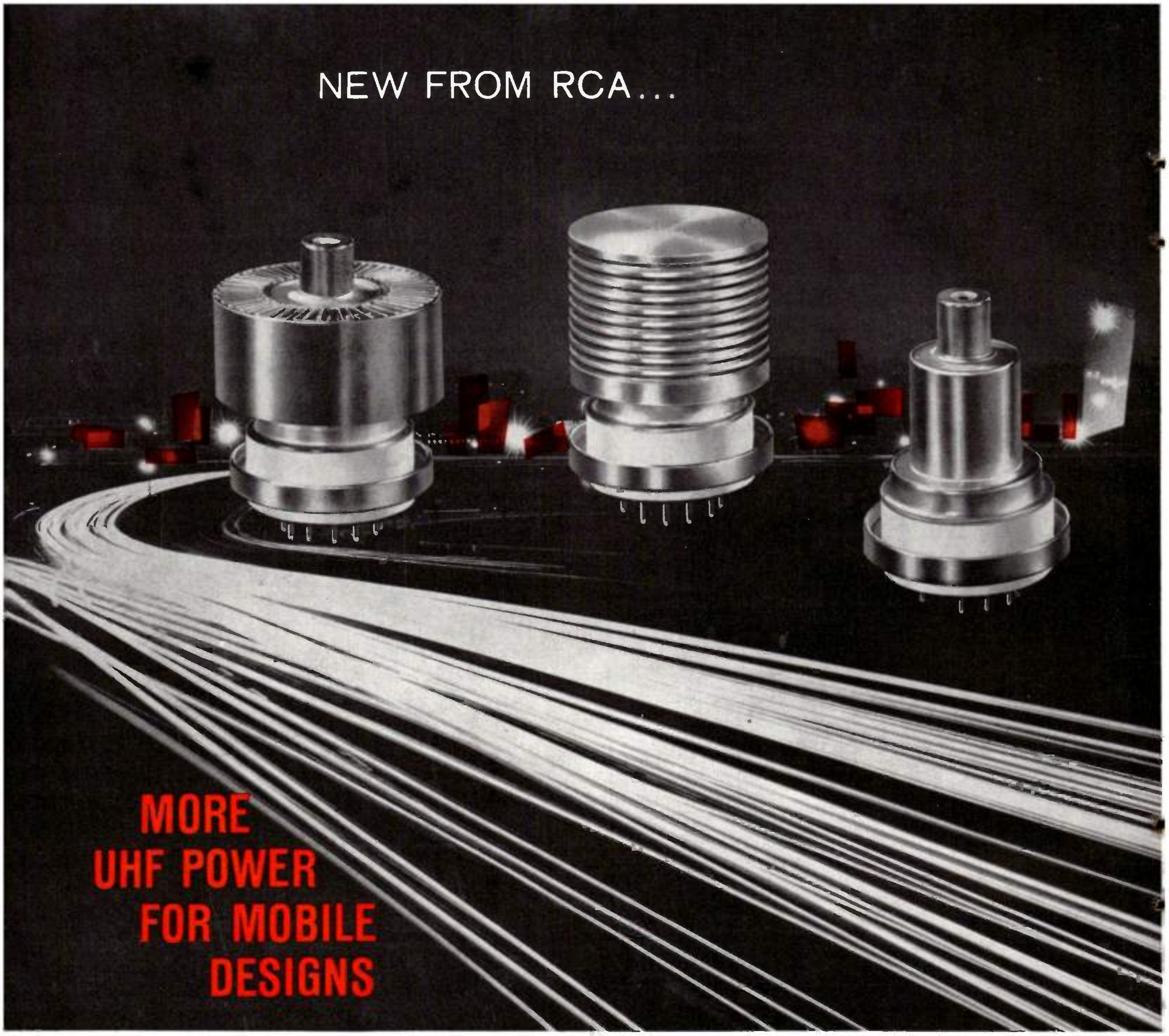
Transfer to Member

Barton, J. R., Chelmsford, Mass.
Beckett, F. J., Sandgate, Brisbane, Australia
Bodden, A. L., Jr., Houston, Tex.
Czerniak, J. L., Muskegon, Mich.
Davis, E. D., McLean, Va.
Dixon, A. E., Kingston, Jamaica, W.I.
Dostal, C. A., Syracuse, N. Y.
Fabrin, K. W., Aarhus, Denmark
Foster, R. M., Ridgefield Park, N. J.
Gosselin, L., Quebec, Que., Canada
Harman, G. F., Jr., Media, Pa.
Hayes, R., Vancouver, B.C., Canada
Killam, K. F., Palo Alto, Calif.
Liuzza, V., Bridgeton, Mo.
Nawata, T. T., Calgary, Alta., Canada
Roberts, H. D., Lutherville, Md.
Selvaggio, V. F., Kailua, Oahu, Hawaii
Smith, J. M., Vancouver, B. C., Canada
Terry, D. M., La Guardia Airport, L. I., N. Y.
Walter, E. R., Lafayette, Ind.

Admission to Member

Abell, M. A., Framingham, Mass.
Adams, S. E., Jr., Atlanta, Ga.
Alger, R. G., Maynard, Mass.
Alstatt, J. D., Merrian, Kans.
Anderson, J. P., Flourtown, Pa.
Anderson, R. J., Ludlowville, N. Y.
Anderson, S. F., Poughkeepsie, N. Y.
Anderson, T C., Sunnyvale, Calif.
Andrews, F. E., Huntsville, Ala.
Aritake, S., Flushing, L. I., N. Y.
Arnold, S. L., Mountain View, Calif.
Babbitt, A. E., Jr., Rockville, Md.
Bankey, G. R., Midwest City, Okla.
Barker, J. H., Minneapolis, Minn.
Baskind, D., Montreal, Que., Canada
Bazemore, T. R., Jr., Burlington, N. C.
Beal, J. W., Jr., Sedgwick, Kans.
Beaumont, J. R., Montreal, Que., Canada
Bemis, R. G., Albuquerque, N. M.
Berghammer, J., Munich, Germany
Biechler, C. S., Acton, Mass.
Blevins, L. L., Liberty, Mo.
Brandon, J. D., Winston-Salem, N. C.
Brown, C., Wappingers Falls, N. Y.
Butterweck, H-J., Aachen, Germany
Buxbaum, C. L., Plainview, L. I., N. Y.
Bybee, J. E., Fairfield, Ohio
Byrd, R. E., Tucson, Ariz.
Cambria, R. C., New York, N. Y.
Campbell, B. J., Sydney, N.S.W, Australia
Carpenter, J. B., Durham, N. C.
Carreras, L. M., Washington, D.C.
Carter, J. L., Jr., Montclair, N. J.
Cassinelli, L. J., Vicente Lopez (B), Argentina
Casciari, R. J., Corning, N. Y.
Cavoto, F. V., Philadelphia, Pa.
Ciemielewski, A. F., Fort Wayne, Ind.
Clark, D. F., Arlington, Va.
Connaughton, W., Jr., Mattapan, Mass.
Conti, M., Milanc, Italy
Costenbader, H. L., Wright-Patterson AFB, Ohio
Cravino, L., Haedo (B), Republic Argentina
Crumrine, B. E., Jr., Media, Pa.
Cusdin, A. G., King City, Ont., Canada
Davey, R. J., Nairobi, Kenya, East Africa
DeVilaiss, A. J., Pasadena, Calif.
Deyo, J. A., North Andover, Mass.
Dieterich, R. A., Paramus, N. J.

(Continued on page 82A)



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Now! Power never before available for compact mobile communications is yours— —with an extraordinary family of RCA ceramic-metal tubes. These high-perveance tetrodes make possible mobile designs capable of achieving new heights in power and compactness.

Designed by RCA expressly for mobile communications, these unique tubes provide a choice of cooling methods. The RCA-8072 is conduction cooled. Specify it in designs where space is at a premium. The RCA-8121 and RCA-8122, air-cooled radiator versions, offer higher power than the 8072 in a tube of comparable size. The three types feature a 13.5-volt heater and are designed to give dependable performance with battery operation.

All three members of this new RCA tube family should become important mobile communications types. Furthermore, other types having different characteristics can be custom designed to fill specific needs. Direct inquiries regarding application of these new types and other RCA electron tubes to your nearest RCA Field Representative or order directly from your RCA Industrial Tube Distributor.

Technical bulletins on these three new tubes are available from: Commercial Engineering, Section A-35-Q, RCA Electron Tube Division, Harrison, N. J.

			Typical CW Operation		
Type	Cooling	Maximum Plate Dissipation Watts	Plate Volts	Freq. Mc.	Useful Power Output Watts
8072	Conduction	100*	700	50 175 470	110 105 85
8121	Forced-Air	150	1500	up to 500	235
8122	Forced-Air	400	2000	up to 500	300

*May be higher depending on heat-sink design.

RCA Electron Tube Division Field Offices

INDUSTRIAL PRODUCT SALES...NEWARK 2, N. J., 744 Broad St., HU 5-3900 • CHICAGO 54, ILLINOIS, Suite 1154, Merchandise Mart Plaza, WH 4-2900 • LOS ANGELES 22, CALIF., 6801 E. Washington Blvd., RA 3-8361 • BURLINGAME, CALIF., 1838 El Camino Real, OX 7-1620. GOVERNMENT SALES... HARRISON, N. J., 415 South Fifth St., HU 5-3900 • DAYTON 2, OHIO, 224 N. Wilkinson St., BA 6-2366 • WASHINGTON 7, D.C., 1725 "K" St., N.W., FE 7-8500

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Vol. 50 *No. 1*

Proceedings of the IRE

Poles and Zeros

Amalgamation. In October, 1961, President Berkner informed all Section Chairmen, by letter, that discussions had been taking place relative to a possible merger of AIEE and IRE. He also sent to Section Chairmen a copy of a joint resolution that had been approved by the Board of Directors of each organization. This joint resolution set forth the background and reasons for the appointment of a committee, named in the resolution, to undertake the studies necessary and appropriate to determine the feasibility, practicability, and form of a possible consolidation. President Berkner's letter and the joint resolution were both published in the PROCEEDINGS in December, 1961, to inform the entire membership of IRE.

The eight man committee, named in the resolution, began its work immediately. Nine two man committees were appointed to study and report on detailed aspects of the consolidation problem. In order to keep the membership fully informed of the progress of this work, President Berkner issued a second letter to Section Chairmen on November 28 outlining the steps taken by the eight man committee to date. The text of the letter appears in full on page 4 of this issue.

President Berkner stated in his initial letter to Section Chairmen, "Both AIEE and IRE negotiators have approached the problem with a sense of statesmanship that would look to creation of a new institute having the best characteristics of both AIEE and IRE . . . ." If we, as engineers, cannot design a single institute to best meet the needs of our profession, who can?

International Activities. The establishment of the Ad Hoc Committee on IRE International Activities Outside of Existing Regions, by the IRE Board of Directors, was reported in "Poles and Zeros" last June. Past President R. L. McFarlan, Chairman, reported for the committee to the Board of Directors at its meeting in October. As a first step in its study the Committee visited Europe during the past summer. It met and carried on discussions with representatives of the Institution of Electrical Engineers (in London), the Societe Francaise des Electroniciens et des Radioelectriciens, the Danish Engineering Society, the Norwegian Society of Engineers, the Swedish Engineering Society, and the German Engineering Society. The Committee reported that the initial discussions were most fruitful. It anticipates continuing liaison with the European community. Two specific results of the visit are the establishment of a section in France, noted in "Poles and Zeros" last month, and an agreement activated between IRE and IEE for the formation of an Advisory Committee in the United Kingdom. This Advisory Committee, when appointed, can act on behalf of the Board of Directors in the United Kingdom, and will report directly to the IRE Executive Committee.

Number Twenty-Nine. Upon recommendation of the Professional Groups Committee, the Executive Committee, at its meeting on November 14, 1961, approved a petition for the formation of a new Professional Group, the twenty-ninth, to be known as the IRE Professional Group on Geoscience Electronics. The group is designed to serve the growing needs of those interested in the research and development of electronic instrumentation for geophysics and geochemistry. Instrumentation for gravity measurements, seismic measurements, magnetics, well-logging, meteorology, and oceanography are typical of the interests of the founding members of the new Professional Group. The petitioners, as one would expect, were from IRE Sections in the southwest, but interest in other parts of the United States and the world should promote this new group into the same healthy pattern of growth that has been typical of its twenty-eight predecessors.

Antiques. In common parlance one would not expect the word antique to have any bearing on fifty year old IRE. On the other hand, the value of an object of yesteryear is comprised not only of its age but also of its rarity. This is simply a preface to calling attention to the value of past issues of IRE publications. There are occasions when circumstances force the dispossession of these treasured items. Complete sets of the PROCEEDINGS have a market value; early complete volumes of the Group TRANSACTIONS are rare. Some ten years ago, when the Professional Group system was inaugurated and when TRANSACTIONS publication started, relatively few copies were printed, and the supply was soon exhausted. Today, as new libraries are stocked, there is a demand for such back issues. So, when you are cleaning out the attic or the radio shack, spare those back issues. They may be of dollar value to you, or they may enable you to make a valuable gift to the technical library of your choice.

100,000—Plus. IRE statistics are published monthly in the PROCEEDINGS. Steady growth in membership toward the oft predicted 100,000, in 1962, is evident. As society membership grows many of its components reach astounding proportions. For example, the size of the PROCEEDINGS print order—it exceeded 100,000, for the first time, in December, 1961.

Reference Directory. The National Science Foundation announced the publication of a new desk reference entitled "Specialized Science Information Services in the United States." This 530 page document, describing 427 scientific information centers, is available for $1.75 from the Superintendent of Documents, Washington 25, D. C.

—F.H.Jr.

P. E. Haggerty

President, 1962–1963

Patrick E. Haggerty (A'45–SM'53–F'58) was born March 17, 1914, in Harvey, N. D. He received the B.S.E.E. degree in 1936 from Marquette University, Milwaukee, Wis. From 1949 to 1951 he did part-time graduate work at Southern Methodist University, Dallas, Tex. In 1959 he received an honorary LL.D. degree from St. Mary's University, San Antonio, Tex., and in 1960 an honorary LL.D. degree from Marquette University.

From 1935 to 1942 he was employed by the Badger Carton Company in Milwaukee. While still a student he worked at Badger as a part-time student engineer. Following graduation he became Production Manager of the company, then Assistant General Manager, with responsibility for all engineering, manufacturing, and administrative functions, except sales.

During World War II he was a Lieutenant, U. S. Naval Reserve, in the Bureau of Aeronautics, Department of the Navy, Washington, D. C., serving, toward the end of the war, as Head of the Electronics Production Branch of the Electronics Components Group, which was responsible for all procurement and production of Naval airborne electronic equipment. In 1945 he joined Geophysical Service Inc., as General Manager of the Laboratory and Manufacturing Division. This organization is now known as Texas Instruments Inc. In 1950 Mr. Haggerty became Executive Vice President and Director of Texas Instruments, and in 1958 he was elected President. He also serves as Director or Officer in major wholly-owned subsidiaries.

Mr. Haggerty is a member of the IRE Executive Committee and Board of Directors, a Life Member and former Vice Chairman of the Board of Trustees of the National Security Industrial Association, and a member of the Society for Exploration Geophysicists. He is also a member of Sigma Phi Delta, Tau Beta Pi, Alpha Sigma Nu, and an Honorary Member of Sigma Iota Epsilon.

Scanning the Issue

An Open Letter from the President—II (Berkner, p. 4)—Last month the PROCEEDINGS published a letter from IRE President Lloyd V. Berkner to Section Chairmen describing the actions taken last October by the Boards of Directors of IRE and AIEE to consider a consolidation of the two societies into a single new society. Because of the importance of the action taken by the two Boards and the high degree of interest in the studies under way, President Berkner recently sent a second letter to IRE Section Chairmen apprising them of the progress being made and arranged for its publication in this issue.

Penetration of the Ionosphere by Very-Low-Frequency Radio Signals—Interim Results of the LOFTI I Experiment (Leiphart, *et al.*, p. 6)—At 10:45 P.M. on February 21, 1961, a Thor-Able-Star rocket left its launching pad at Cap Canaveral with an unusual double payload consisting of a TRANSIT navigational satellite with another, smaller satellite mounted on top of it. The smaller satellite was the first of a new family of satellites called LOFTI. During its 562 revolutions around the earth, LOFTI received and recorded VLF signals originating from the Navy transmitting stations in Panama and at Jim Creek, Wash. By the end of its 14 million mile trip, the satellite had telemetered a vast quantity of data to 10 receiving stations in North and South America and to one in Australia. This paper presents the preliminary results of an important satellite experiment which will shed a great deal of light on VLF propagation in the ionosphere.

Decreasing the Threshold in FM by Frequency Feedback (Enloe, p. 18)—With the advent of space communications there has been widespread interest in receiving systems which have extreme sensitivity. This, in turn, has focused attention on modulation techniques which trade bandwidth for improved signal-to-noise ratios. In the course of developing a highly sensitive receiving station for the Project Echo communication satellite experiments, the designers turned their attention to FM. Signal-to-noise ratios can be considerably improved by using large index frequency modulation, but unfortunately this improvement applies only when the carrier-to-noise ratio is relatively high. It was then recalled that in 1939 Chaffee had proposed an almost-forgotten frequency feedback circuit which greatly compressed the frequency deviation of the large index wave so that it could be passed through a much narrower band-pass IF filter before demodulation. The effect of this compression was to substantially reduce the minimum carrier-to-noise ratio (*i.e.*, the threshold) needed to yield a signal-to-noise improvement. By using this frequency feedback circuit, together with a maser amplifier and a special low-noise horn antenna, the Project Echo designers succeeded in building the world's most sensitive communications receiver. There has remained, however, considerable misunderstanding as to how and where the above mentioned threshold sets in. This paper performs a double service by clearing up the false conclusions of previous papers and by making it possible for the first time for a designer to predict accurately the manner in which this important technique will extend the threshold.

Theory and Design of Wide-Band Parametric Converters (Kuh, p. 31)—The author treats the parametric amplifier and converter from a circuit theory point of view which leads him to a general study of maximum gain-bandwidth relationships and to developing practical procedures, which he illustrates, for designing coupling networks that approach the optimum gain-bandwidth. The approach developed here is more general than any that have been published heretofore. The results therefore can be applied by many workers in many fields.

A New Precision Low-Level Bolometer Bridge (Reisener and Birx, p. 39)—Using radiometer principles, a new type of bridge has been developed for measuring extremely low-level RF voltages and powers. The instrument works by comparing and matching the heating effect of the RF pulse under measurement to that of a controllable dc pulse of known value on a bolometer element. The bridge was found to have a sensitivity of 3×10^{-10} watts. Although designed primarily for RF measurement, the bridge may eventually be used to measure any kind of electromagnet radiation when suitable detectors and modulators are developed. As the best bridge detection and calibration method available today, this development will interest many engineers working with signal generators, receivers and radiometers.

***P-N* Junction Charge-Storage Diodes** (Moll, *et al.*, p. 43)—This paper presents the design theory for a class of semiconductor diodes which are deliberately designed to conduct for a brief time in the reverse direction after the forward bias has been turned off. This feature distinguishes these diodes from those intended for switching and rectification applications, in which the ideal recovery time is zero, and makes them of particular interest for circuit applications involving fast pulse generation, wave shaping and harmonic generation. The diodes described here are also noteworthy in that following the reverse conduction period the current decays very rapidly to zero. The discussion adds substantially to the theoretical understanding of reverse transient phenomena and calls attention to the circuit uses of the "storage phase" and the "decay phase" of the reverse recovery period of diodes.

Average Decay Laws for VLF Fields (Wait, p. 53)—By making a number of reasonable approximations, the author concisely derives a simple formula for relating the average field strength of VLF waves to distance and frequency for intermediate distances of from 400 to 4000 kilometers. This formula gives a simple physical picture of a number of broad characteristics of VLF fields that are of interest in a number of fields, including wave propagation, communications and navigation.

Theory, Design and Performance of Maximum-Efficiency Variable-Reactance Frequency Multipliers (Utsunomiya and Yuan, p. 57)—The high efficiency obtainable with nonlinear-reactance multipliers makes attractive the possibility of using them as multipliers of considerably high order. The amount of fundamental power that can theoretically be converted to any specified harmonic has been fully worked out in previous papers. However, in practice there are many factors which contribute to lowering the efficiency of a real multiplier. The analysis presented here takes these factors into account so that a designer, once he knows the characteristics of a nonlinear element and the circuit losses, can predict the maximum obtainable efficiency. The results also show the important role played by intermediate harmonic currents.

The Study of Large-Signal High Frequency Effects in Junction Transistors Using Analog Techniques (Beale and Beer, p. 66)—An ingenious analog of a junction transistor has been devised in which various transistor parameters are represented by circuits made up of linear and nonlinear elements. The analog makes it possible to determine the magnitude of several effects that arise at large signal levels and high frequencies, and from these observations to obtain general expressions governing the operation of transistors.

Parametric Behavior of an Ideal Two-Frequency Varactor (Montgomery, p. 78)—This brief paper fills in the previously incomplete theory of operation of the simplest type of varactor network, *i.e.*, one in which power is exchanged at only two frequencies. In particular, the theory identifies all the signal-to-pump frequency ratios at which operation is possible and makes clear those ratios at which operation is impossible.

Open Letter from the President—II

The December, 1961, issue of the PROCEEDINGS *carried the text of a letter dated October 20, 1961, from Lloyd V. Berkner, President of the IRE, to IRE Section Chairmen concerning the possible consolidation of the IRE and AIEE. In order that the membership may be fully informed, published below is a second letter from President Berkner regarding the progress being made.—The Editor*

November 28, 1961

Dear Mr. Section Chairman:

On October 20, 1961, I advised you of the important action of the Boards of IRE and AIEE toward consolidation of the two societies into a single Institute. The action toward consolidation is so important that I am writing to advise you of the progress being made.

Following the actions of the two Boards, the legal counsel of IRE advised me that the AIEE resolution was substantially identical to that of the IRE Board and I therefore declared the IRE resolution in force. Immediately thereafter the eight men appointed by the resolutions to the Joint Committee on Consolidation were notified. These members include Messrs. Warren H. Chase, Clarence Linder, Elgin Robertson, Sr. and Dr. B. R. Teare, Jr. for the AIEE, and Dr. P. E. Haggerty, Dr. Ronald McFarlan, Mr. Haraden Pratt and me for the IRE.

I designated President-Elect P. E. Haggerty as principal representative on the Committee to act for IRE.

Following these notifications, two-man study committees were appointed by Dr. Haggerty to study and develop guidelines on prime aspects of a consolidation plan, each member being notified by letter of his appointment. These committees were organized as follows:

Assets, Resources & Finances

Stuart Bailey (IRE) Jansky & Bailey, Incorporated	J. Harold Moore, for H. I. Romnes (AIEE) Western Electric Company

Editorial Policy & Publications

Ferdinand Hamburger, Jr. (IRE) Johns Hopkins University	Morris D. Hooven (AIEE) Public Service Electric & Gas Company

Extra Society Relationships

Ronald L. McFarlan (IRE) Consulting Engineer	T. M. Linville (AIEE) General Electric Company

International Activities

Ronald L. McFarlan (IRE) Consulting Engineer	To be appointed (AIEE)

Merging of Professional Groups and Technical Committees

Ernst Weber (IRE) Polytechnic Institute of Brooklyn	Hendley Blackmon (AIEE) Westinghouse Electric Corporation

Relations to Standardization

Ernst Weber (IRE) Polytechnic Institute of Brooklyn	C. G. Veinott (AIEE) Reliance Electric & Engineering Company

Merging of Sections: Finances, Geographical Boundaries, Sectional Publications

G. K. Teal (IRE) Texas Instruments, Incorporated	Walter Scholz (AIEE) Philadelphia Electric Company

Personnel

Donald G. Fink (IRE) Philco Corporation	C. F. Savage (AIEE) General Electric Company

Student Branches

Dean J. D. Ryder (IRE) Michigan State University	R. T. Weil, Jr. (AIEE) Manhattan College

Appointments are also being made for IRE and AIEE to two-man study committees on Honors and Awards and Regional Structure.

Under the direction of President-Elect Haggerty, a work book was prepared and provided to each member of the eight-man joint committee, which contained complete information on the principal subjects to be considered in effectuating consolidation. The principal features of this work book are contained in the following outline:

I Organization and Schedule to Prepare Draft
II Organizational Charts
III Comparative Alignment of Constitutions
IV. Comparative Alignment of Bylaws (and Pertinent Constitutional Articles)
  A. Membership
    1. Privileges
    2. Classes
    3. Admissions, Transfers, etc.
    4. Terminations
  B. Membership Fees
    1. Entrance
    2. Transfer and Annual Dues
    3. Publication
  C. Organization—Membership
    1. Districts/Regions
    2. Sections
    3. Professional Groups
    4. Student Branches
    5. Overseas Chapters
  D. Organization—Management
    1. Nominations, Elections, Balloting
    2. Board of Directors
    3. Executive Committee
    4. Officers
    5. Standing Committees—General
      Administrative
      Technical
  E. Meetings
  F. Miscellaneous
  G. Publications
V. Statistical Data
VI. Comparative Time Table
VII. Two-Man Study Committee Recommendations.

By November 17, the initial reviews of most of the two-man study committees were completed and their reports distributed to all members of the eight-man joint committee for inclusion in the work books.

On the evening of November 16, and throughout the day of November 17, a four-man subcommittee met, composed of Mr. Chase and Dr. Teare for the AIEE and Mr. Pratt and Dr. Haggerty for the IRE. They were joined by Mr. Elgin Robertson, Sr. of the AIEE. The reports of the two-man study committees were reviewed in detail. Based on these reports and the considerations and judgments of the joint subcommittee, Mr. Haraden Pratt was asked to prepare an initial constitution and bylaws draft for a combined society of AIEE and IRE. It will be his goal to have the initial constitution and bylaws draft in the hands of the eight-man joint committee for review by December 5, and he will additionally and simultaneously provide a copy to Mr. John Tobin, Attorney for the IRE, and Mr. Simon Presant, Attorney for the AIEE.

The full eight-man committee will meet in New York on December 13, 1961, for first consideration of the draft of the constitution and bylaws. At that time the replies of the Sections to my letter of October 20 will be given full consideration.

The plan of action of the committee is divided into two parts:

1. To produce a constitution and bylaws for the new society, drawing upon the best elements of the experience of each society, which would best serve the technical and professional requirements of the profession.
2. Following completion of the action of paragraph #1, the committee will make such modifications in the proposed constitution and bylaws as are necessary to effect a merger. This will involve inserting temporary clauses to be effective for a year or more in order to complete the consolidation.

Throughout these negotiations, I have been impressed by the statesmanlike approach of all members of the committee. The framework of the negotiations remains within the outline of my letter of October 20, 1961. There is no feeling that one society is trying to take over the other, but rather a genuine spirit of jointly creating a single new and stronger Institute out of the already strong roots of the two existing Institutes, thereby eliminating undesirable competition and combining resources to better serve and advance our rapidly evolving profession. In this, the central Headquarters of the two Institutes are giving their full cooperation. Mr. Chase, President of AIEE, and I are in constant communication and we continually exchange correspondence relating to the proposed consolidation.

I will continue to write frequently so that you can be fully and authoritatively informed, and prepared to answer any questions that may be raised in your Section. It would be most helpful if you could keep your membership fully informed so that incorrect rumors do not multiply. Likewise, I would be happy to receive further comments from your Section at any time.

Sincerely yours,

L. V. Berkner

L. V. BERKNER
President

Penetration of the Ionosphere by Very-Low-Frequency Radio Signals—Interim Results of the LOFTI I Experiment*

J. P. LEIPHART†, MEMBER, IRE, R. W. ZEEK†, SENIOR MEMBER, IRE, L. S. BEARCE†, ASSOCIATE, IRE, AND E. TOTH†, FELLOW, IRE

Summary—**LOFTI I is the first of a series of Navy satellite experiments designed for determination of the degree and extent of very-low-frequency (VLF) radio wave penetration of the ionosphere. This investigation has provided unique data on intensity, time delay, echoes, and other characteristics of VLF signals received in the ionosphere from transmitters on the Earth's surface.**

Theoretical studies carried on concurrently with planning and implementation of the experiment have indicated that VLF radio energy from transmitters operating below the lower atmosphere-ionosphere interface should appear in the ionosphere in useful amount. The LOFTI I telemetry records show relatively strong signals in the ionosphere from 18-kc transmitters on the terrestrial surface. The signals were attenuated less at night and appeared as far away as Australia, 10,000 miles from the transmitter.

Statistical treatment of the data so far reduced shows that the attenuation of magnetic field intensity of the 18-kc time pulses from Naval Radio Station NBA, as measured near extreme line of sight distances to the north of the station, was less than 13 db 50 per cent of the time at night and less than 38 db 50 per cent of the time during the daylight hours. The data studied so far shows very little effect of altitude on signal intensity, an observation which agrees with the theoretical treatment based on a model ionosphere.

The observed time delays of the 18-kc signals range from 10 to 200 msec, indicating that VLF propagation velocity in the ionosphere is much lower than in free space. The occurrence of whistler-path propagation is indicated by sequential echoes of the NBA time pulses which appear in the telemetry records as much as 4/3 seconds after the directly-received pulses.

Much data remains to be reduced from LOFTI I. Further experiments are planned to provide data on electric dipole impedance in the ionosphere (lost to the LOFTI I experiment because of a launch vehicle malfunction), on the effect of operation at higher latitudes and over a wide range of signal frequency, and on the characteristics of the signals produced in the lower atmosphere by VLF transmitters radiating in the ionosphere.

INTRODUCTION

LOFTI (LOw Frequency Trans-Ionospheric) I is the first of a series of Navy satellite experiments designed to explore low-frequency radio wave propagation in the ionosphere. The LOFTI studies were proposed early in 1958 by the Naval Research Laboratory as part of a comprehensive satellite communication research program.

The LOFTI I experiment was planned as an initial study of the magnitude and other characteristics of VLF radio signals emitted at the earth's surface as observed in the ionized regions of the atmosphere. Sufficient definitive data was expected to guide the planning of further LOFTI experiments, which would add to the fund of scientific information and lead to improved communication systems.

GENERAL NATURE OF THE IONOSPHERE

Earth's ionosphere impedes the passage of low-frequency radio waves from the terrestrial surface into outer space. At the higher frequencies, its effectiveness as a barrier decreases to a degree such that it becomes essentially negligible above, roughly, 100 Mc, so far as transmission loss is concerned.[1] The ionization of the atmosphere, which is caused largely by solar radiation, varies above any particular point on the earth's surface with time of day, season of the year, solar storms, the sunspot cycle, etc. The observed influence of the ionosphere on terrestrial radio wave propagation has led to its classification into strata, the D region and E, F_1 and F_2 layers which, when all are present, appear in ascending order above the Earth from about 50 to about 400 km (roughly, 30 to 250 statute miles) altitude.

Somewhat more than half of the Earth's rotating atmosphere always faces the sun and is subjected to the intense and variable solar radiation while the rest is in shadow. The depth of atmosphere through which the sun's rays must pass becomes greater toward the periphery of the illuminated hemisphere. A major change of atmospheric ionization occurs in the transition from sunlight to shadow and vice versa, resulting in significant diurnal variations in the performance of terrestrial radio systems operating below approximately 100 Mc. Fig. 1 illustrates one effect of the daily cycle.[2] It shows the change in phase delay of a VLF signal originating about 3200 km (2000 miles) due south of a receiver, as observed over a 24-hour period. The 18-kc transmitter at Naval Radio Station NBA, Canal Zone (which was the main source of VLF signals in the LOFTI I experiment) radiated the signals observed in the Fig. 1 measurements. The measuring receiver was located at the

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* Received by the IRE, April 10, 1961; revised manuscript received, October 13, 1961.

† Radio Div., U. S. Naval Research Lab., Washington, D. C.

[1] K. G. Budden, "Radio Waves in the Ionosphere" Cambridge University Press, London, Eng.; 1961.

[2] From A. R. Chi and H. F. Hastings of the Radio Division, USNRL.

Naval Research Laboratory in Washington, D. C. The transmitter and receiver sites passed through the sunlight to shadow zone and back almost simultaneously during the period in which these measurements were made (near the autumnal equinox). The increase in phase delay observed after sunset indicates lengthening of the signal path between the stations. Greater path length would be one result of the disappearance of the *D* region with the onset of darkness and the consequent increase in height of the earth-ionosphere space.

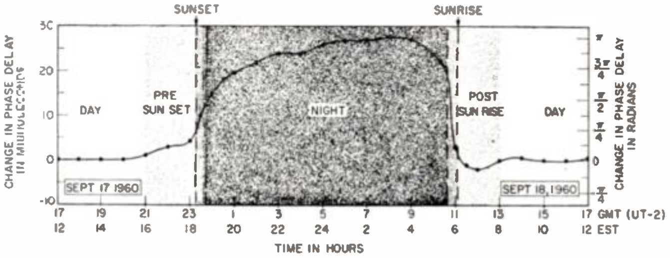

Fig. 1—Effect of diurnal change of solar illumination on the propagation time of an 18-kc radio signal. North-south path; distance between transmitter and receiver approximately 3200 km; both stations passing almost simultaneously from daylight into darkness and from darkness into daylight. Note rapid reduction of phase delay immediately preceding sunrise, indicating swift reconstitution of *D* region after a less rapid disappearance with the onset of darkness. (From Chi and Hastings, NRL.)

The data available indicates that the density of the atmosphere even at the lower boundary of the *D* region is very low, about 1×10^{-8} of the sea level value, and drops rapidly with increasing height. At 400 km (250 miles) altitude, atmospheric density apparently is near hard vacuum proportions, about 1×10^{-11} of the sea level atmosphere. The average rate of electron collision with other particles in the ionosphere (the collision frequency) decreases correspondingly, although it is also affected by other factors such as temperature and ion concentration.

Fig. 2 indicates degree of ionization in terms of free-electron concentration as a function of altitude above the earth's surface.[3,4] It also shows the general nature of the variation of collision frequency with altitude.[3]

Propagation Loss in the Earth-Ionosphere Circuit

The propagation loss between a VLF transmitter on the ground and a receiver within line-of-sight distance in the ionosphere involves three elements. One of these is the spreading loss in the lower atmosphere to the point of signal entry into the ionosphere. Additional circuit attenuation is then introduced as the result of splitting of the radio wave as it impinges on the boundary region or interface between the ionosphere and the lower atmosphere. Part of the signal energy is reflected back toward the Earth's surface and the remainder is propagated upward in a magneto-ionic mode toward outer space.

The further attenuation of the signal caused by absorption of energy as the radio wave traverses the ionized medium depends on wave frequency, inclination of the wave normal relative to the Earth's surface, the electron density, and the collision frequency. The last two factors are the ones most altitude dependent. Magneto-ionic theory indicates that VLF radio waves will propagate within the ionosphere with least absorption when the angle of inclination θ of the wave normal relative to the geomagnetic field is zero or 180°. Since the VLF wave front after entry into the ionosphere is not planar, some reduction in field intensity due to spreading as the wave moves upward may be expected.

The computed rate of attenuation of 18-kc signal level due to absorption within the model ionosphere of Fig. 2 (for the condition $\theta = 0$) is shown in Fig. 3 as a function of altitude above the terrestrial surface.

The attenuation rate is approximately

$$\alpha = 1.8 \times 10^{-8} N^{1/2} \nu, \tag{1}$$

in which α is attenuation in db/km, N is the number of free electrons per cubic cm and ν is the collision frequency in number of collisions per second. This expression is derived from a statement defining the complex refractive index of a magneto-ionic medium (Appendix I). The coefficient (1.8×10^{-8}) of (1) is valid only when the electron density is greater than 10^3 electrons per cubic cm, the collision frequency is less than 10^6 collisions per second, and the gyro frequency is approximately 1000 kc. These conditions hold generally for the LOFTI I experiment. The equation in Appendix I was used in calculating Fig. 3 to arrive at more exact values than provided by (1).

Fig. 4 (derived from Fig. 3) shows the cumulative least possible attenuation of an 18-kc signal traversing the ionosphere radially outward from Earth within the range of altitude covered by LOFTI I. This loss does not include whatever attenuation of signal level may result from spreading of the signal energy within the ionosphere and also does not include the interface or entrance loss previously mentioned. It appears that most of the absorption loss attributable to the ionosphere is accumulated in passing through the *D* region between about 60 and 130 km altitude (37 and 80 miles) and that there is little further attenuation with further increase in altitude. The calculated total attenuation caused by absorption in the model ionosphere in its midday state is less than 30 db (about 27 db) under the conditions here considered. The corresponding least possible loss in passing through the model night ionosphere is about 2 db.

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[3] J. A. Ratcliffe, "Physics of the Upper Atmosphere," Academic Press, New York, N. Y., and London, Eng.; 1960. See especially p. 108: summary graph of data for summer noon, middle latitudes, and sunspot minimum.

[4] Stanford Research Institute model of electron density in the night ionosphere.

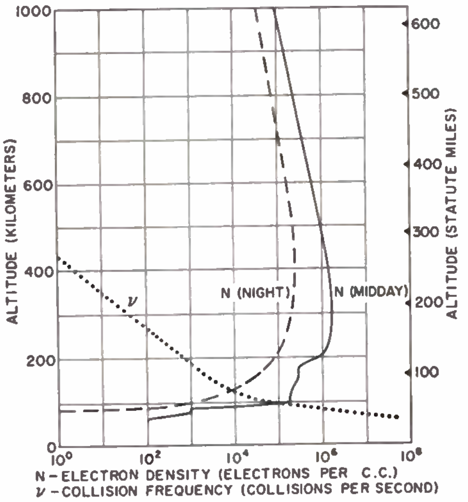

Fig. 2—Estimated concentration of electrons in the night and day atmosphere and electron collision frequency vs altitude. (After SRI and Ratcliffe.)

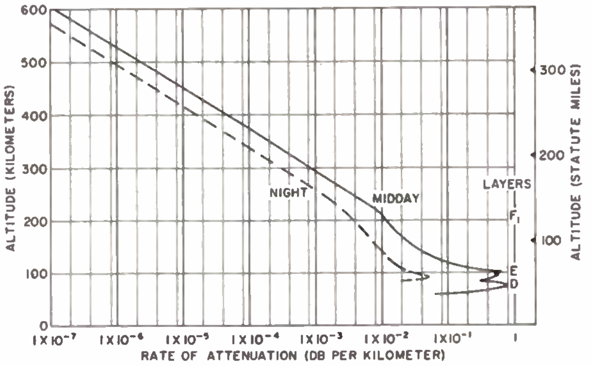

Fig. 3—Computed rate of attenuation of an 18-kc radio wave in the model ionosphere of Fig. 2 vs altitude.

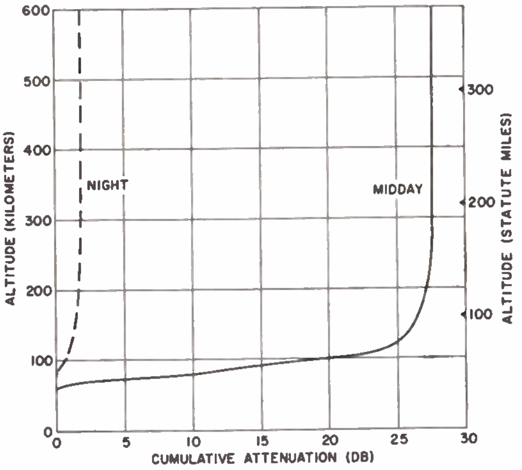

Fig. 4—Minimum possible transmission loss (least cumulative attenuation) of an 18-kc radio wave passing vertically upward through the model ionosphere of Fig. 2 vs altitude. Geomagnetic field aligned with direction of propagation. (Part of the signal energy that reaches the lower boundary of the ionosphere from the lower atmosphere does not penetrate the interface. The consequent reduction in signal level (the interface loss) and any spreading loss of the ionosphere are not included in the attenuation shown above.

SYSTEM CONCEPT OF THE EXPERIMENT

The LOFTI I experiment required three radio links, as shown in Fig. 5. Navy shore transmitting installations were the sources of VLF signal. Originally, it was planned that three mutually orthogonal magnetic dipoles (loop antennas) would be mounted on the satellite, each connected to an individual VLF receiver. In addition to measurement of VLF signal level, comparison of the signal output of the three loops in conjunction with satellite aspect data would provide indication of direction of wave arrival and other information. Another VLF receiver in the satellite, employing an electric dipole antenna (two whips) and a test oscillator periodically connected to the antenna, would indicate the effect of the ionosphere on electric antenna impedance and tuning. A similar test oscillator would be provided for the loop receivers.

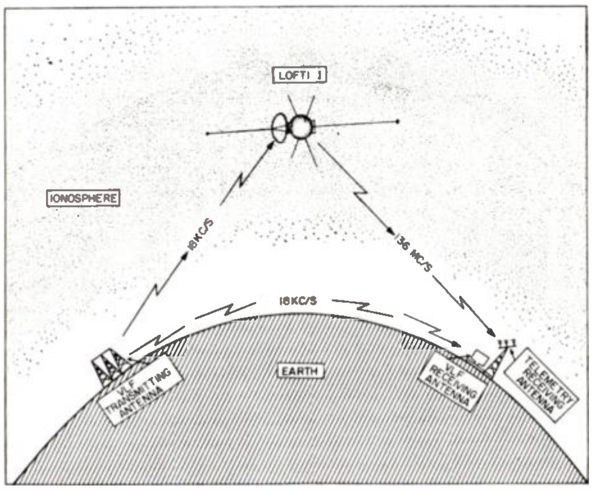

Fig. 5—System concept of the LOFTI I experiment showing the three basic radio links.

The various VLF signal outputs would each modulate the frequency of an assigned subcarrier of the satellite's VHF amplitude-modulated telemetry transmitter. The subcarriers of the VHF telemetry transmissions and terrestrial signals in the range of 0.3 to 30 kc received at the ground stations were to be recorded on magnetic tape together with other related data. The records would subsequently be reduced and analyzed to derive attenuation, amplitude distribution, group delay, phase delay, Doppler shift, noise in the ionosphere and other desired items of information.

IMPLEMENTATION OF THE EXPERIMENT

In the fall of 1959, it was found that additional payload could be accommodated in a forthcoming satellite launch (Navy's TRANSIT III B). Simultaneous launch of LOFTI I with TRANSIT would make it possible to proceed with the VLF study with least delay and expense. It was intended that the two satellites separate and go their individual ways when orbit had been attained. The inclination of the orbit (28°) and its constant height (about 1000 km or 600 statute miles) had been chosen for the TRANSIT experiment. Allowable payload considerations governed LOFTI satellite size and weight, and limited its VLF antenna complement to a single plane loop and a two-element whip. Two identical VLF receivers were coupled to the two antennas through suitable networks. Both receivers were tuned to 18 kc, the frequency of Naval Radio Station NBA's VLF transmissions.[5]

Station NBA is located in Panama (Canal Zone) at close to 9° north latitude, well inside the ±28 degree limits of the intended orbit. This station transmits highly precise time pulses on a very accurately controlled constant frequency carrier.[6] Radiated power is about 30 kw. The carrier frequency, nominally 18 kc, is very slightly offset to compensate for the difference between Universal Time (UT-2) and Atomic Time (A-1). These signals would be easily identifiable in the telemetry records and would in addition serve as a highly accurate time and frequency reference. Also, the periods between pulses would provide sequential intervals for determination of VLF noise level in the ionosphere.

ORBIT ATTAINED

The two-satellite combination was launched at 10:45 P.M. Eastern Standard Time (EST) on February 21, 1961 from Cape Canaveral, Florida, on a Thor/Able-Star rocket vehicle. Unfortunately, the second stage of the rocket did not separate from the satellite assembly. This malfunction interrupted the programmed sequence of events and prevented satellite separation. The three components (rocket and satellites) remained locked together, moving as a unit along an orbit with the intended apogee but a perigee so low that the life of the assembly turned out to be only about 36 days and 9 hours. LOFTI I made its last transmission on March 30, 1961.

The new composite satellite, designated 1961 ETA, had the following computed major orbital elements four days after launch:[7]

apogee—960 km (597 statute miles),
perigee—166 km (103 statute miles),
period for one revolution, perigee to perigee—95.9 minutes,
progression of perigee—+10.67 degrees per day.

By accident, LOFTI I had achieved an eccentric orbit which closely approached the ideal originally envisioned for this VLF experiment. It had begun a short career of plunging up and down through the upper reaches of the ionosphere 15 times each 24 hours, passing through the *F* layers and approaching the *E* layer about as closely as a satellite could without burning up immediately because of atmospheric friction.

LIMITATIONS ON THE EXPERIMENT

Other departures from the intended conditions of the experiment were not so desirable as the eccentric orbit or as acceptable as the shorter life expectancy of the satellite. It had been planned that the two 15-foot whip antennas wound on spools inside LOFTI I would be unreeled some time after satellite separation to function as a 30-foot electric dipole. LOFTI should then have been rotating about an axis parallel to the plane of its loop and perpendicular to a line drawn from the center of the loop to the center of the satellite's shell. Extension of the whips would have slowed this rotation to between 15 and 30 rpm. The resulting continuous slow spin (about 120°/sec) would have swept the loop and whip antennas through a maximum output position relative to the signal wave front twice each revolution of the satellite. Satellite attitude information would have been provided by the variations in earth and solar aspect sensor output.

The unseparated trio (rocket and satellites) did achieve a very slow tumbling and rotating motion (roughly, 1/20 rpm) which, however, was not a satisfactory substitute for the intended rotation. As a result, the aspect sensors scanned much too slowly to be effective. LOFTI's whip antennas could not be extended because of interlocks in the interrupted launch sequence and the physical presence of TRANSIT.

Another difficulty resulting from the failure to separate was that TRANSIT electronics interfered with VLF signal reception by LOFTI. The radio and electrical systems of both satellites had been designed on the premise of wide physical separation of the two systems in orbit. The desired signal data must consequently be separated not only from background noise but also from TRANSIT interference.

Thus, although all of LOFTI's instrumentation was operative and capable of performing as intended, only the loop receiving system could function at full inherent capability so far as VLF operation was concerned, and even it was misoriented and partially jammed by TRANSIT interference much of the time. NBA signals did appear in the output of the whip receiver during launch and also during night passes when the satellite came close to the transmitter, even though the whips remained within LOFTI's shell unextended. However, it is possible that the VLF energy was coupled into the whip receiver from the loop receiving system in some manner.

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[5] An experiment of similar configuration was proposed by Storey in 1959 for determination of local electron density by measurement of wave admittance in the ionosphere (L. R. O. Storey, "A method for measuring local electron density from an artificial satellite," *Res. NBS*, vol. 63D, pp. 325–340; November–December, 1959.)

[6] See U. S. Naval Observatory Time Service Notice No. 8; November 18, 1959. Time and frequency control is provided by the Observatory; control instrumentation was provided by the U. S. Naval Research Lab. in 1959.

[7] Orbital data from NASA; more complete tabulation in Appendix II.

Electric antenna detuning could not be observed because of inability to extend the whips. No significant change in magnitude of loop impedance appears in the data so far examined.

PORTIONS OF THE IONOSPHERE OBSERVED IN THE EXPERIMENT

During its 562 revolutions around Earth, LOFTI I and its companions traveled more than 14 million statute miles at an average velocity of approximately 4.7 miles per second. In its many traversals of the atmosphere, LOFTI moved within a belt roughly 4000 miles wide, 500 miles deep, 100 miles off the Earth's surface and centered above the equator.

Ten ground stations located in North and South America (indicated by small circles in Fig. 6) were instrumented to receive the 136.17 Mc telemetry signals radiated by LOFTI. The VLF transmitting stations which were the sources of 18-kc signals in the experiment appear in Fig. 6 as small stars. Three of the receiving stations were on an approximately east-west line and eight were approximately in line north and south along the meridian of station NBA. An eleventh station located in south-central Australia received the telemetry signals in that region.

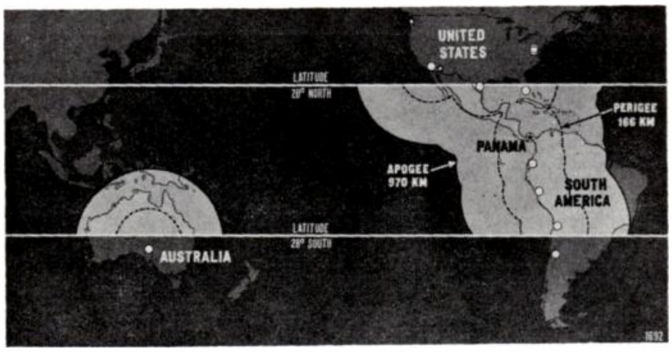

Fig. 6—Terrestrial areas over which VLF reception by the satellite (at highest apogee and lowest perigee altitude) could be observed by telemetry. Location of telemetry receiving stations is indicated by circles; sites of VLF transmitters involved in the experiment are indicated by stars.

The eleven stations could observe VLF signal reception by the satellite when it was over the territory shown lightly shaded in Fig. 6. These two regions were entirely covered by telemetry at highest apogee of the satellite but the area of coverage became smaller as the apogee became progressively lower.[8] Coverage at perigee, which changed very little during the life of the satellite, was limited to the smaller areas outlined by the dotted lines.

From one to five of the ground stations could receive the LOFTI I telemetry signals as the satellite passed over the North and South American regions. Reception by these stations overlapped on some passes, providing occasional redundancy of data. Six of the telemetry stations were each equipped with a 10-foot square loop antenna and VLF receiver for direct reception of 18-kc signals. This facility made possible comparison of the telemetered VLF signals with VLF signals which had arrived at these stations via the usual terrestrial paths.

SIGNALS RECEIVED FROM NBA

As the satellite came within their range during its first revolution around the Earth, the North American field stations began reporting good telemetered VLF signals. The South American station tapes subsequently received also showed good VLF signal reception and the tapes received from Australia showed that NBA signals were reaching the satellite more than 10,000 miles away.

Figs. 7 and 8 are twelve-second samples of LOFTI data as recorded at the NRL satellite research station at Stump Neck, Maryland. In Fig. 7, the time is 35 minutes after midnight EST on February 23, 1961. The satellite was then in its 16th revolution around the earth, 550 km (340 statute miles) high and north-northeast of the transmitter at a ground distance of nearly 1600 km (1000 miles).

The upper trace (telemetry subchannel IRIG[9] band 4), shows a section of the keying envelope of the NBA signal as it appeared in the detector output[10] of LOFTI's loop receiver. The detector output of the whip receiver and that of the loop receiver time-shared IRIG band 4. The parts of the trace labeled "marker" are intervals in which the subcarrier frequency was shifted in the satellite to either the upper or the lower limit of band 4 to indicate that loop receiver or whip receiver output, respectively, would follow. Since the whips were not extended, the detector output of the whip receiver contained little or no desired signal information. For this reason, only a few seconds of the whip receiver record are shown.

The envelope of eight of NBA's time ticks (one pulse per second, each 0.3 second long) follows after the loop marker interval in the upper trace of Fig. 7. The sharp downward spikes appearing in the marker interval and at the tops of the time pulse envelopes are indicative of disturbances in telemetry reception rather than field strength variation of the NBA signal at the satellite. Satellite VLF receiver output saturation level is exceeded part of the time in this sequence, indicating that the magnetic field of the NBA signal at the satellite was equivalent in intensity to the magnetic field associated

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[8] See Appendix II.

[9] Inter-Range Instrumentation Group Standards were used in assignment of subcarrier frequency bands (see Appendix III).

[10] An envelope detector channel branched out from the first intermediate frequency (IF) amplifier system of each VLF receiver. Its input (first IF) bandwidth was 28 cps at 3 db down and its output (RC filter) cutoff was about 30 cps.

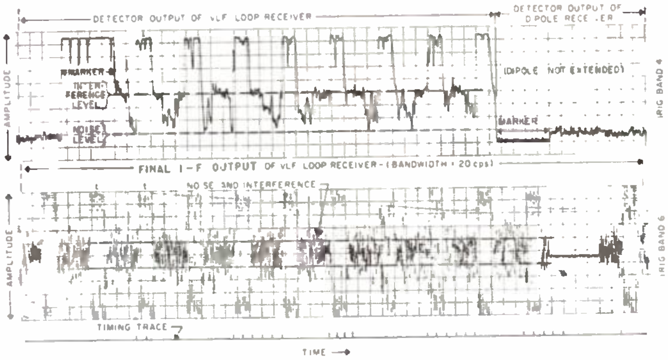

Fig 7—Typical signal output of LOFTI I loop receiver (detector and final IF amplifier). Orbit revolution 16, February 23, 1961, 12:35:38 to 12:35:50 A.M. EST. Satellite over Caribbean area; height, approximately 550 km; ground range from VLF transmitter (NBA, Panama), approximately 1600 km, north-northeast.

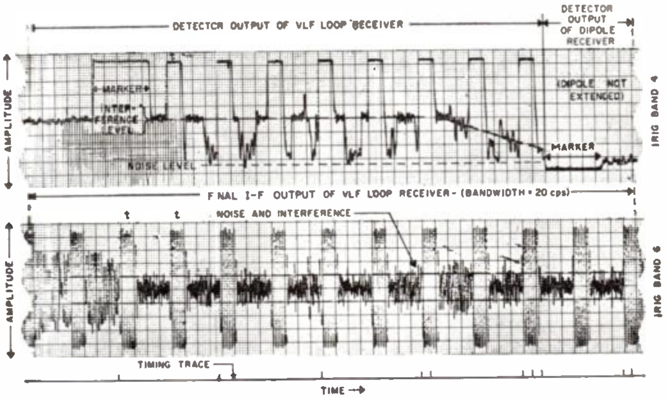

Fig. 8—Signal output of LOFTI I loop receiver under better telemetry conditions. Orbit revolution 29, February 23, 1961, 9:10:33 to 9:10:45 P.M. EST. Satellite over Gulf of Mexico; height, approximately 200 km; ground range from VLF transmitter (NBA, Panama), approximately 2500 km, north-northeast.

with an electric field intensity of at least 100 μv/m[11] in free space.

Interference generated by TRANSIT's electronic system is evident in the intervals between the time ticks. While obscuring the desired signal only partially in this case, it is evident that TRANSIT interference could be expected to make determination of signal intensity from the telemetry records difficult when the VLF signals are weak.

The lower trace in Fig. 7 shows the second (final) IF amplifier output of LOFTI's loop receiver (25 cps center frequency, 20 cps bandwidth at 3 db down). The receivers were designed to compress an input signal range of 40 db into an output signal range of 20 db. Corrected for this compression, the NBA signal to ambient noise ratio (in a 20-cps band and excluding TRANSIT interference) is 46 db for the eight-second period shown.

Fig. 8 is similar to Fig. 7 but was recorded during orbit revolution number 29, 10 minutes after 9 P.M. EST on February 23, 1961. The satellite is lower, at about 190 km (120 miles) altitude, and farther from the transmitter (about 1500 miles north-northeast). The signal and interference waveforms are more clearly defined due, most likely, to better telemetry reception. However, the corrected signal-to-noise ratio (SNR) is the same (46 db). The signal level in Fig. 7 is somewhat greater, which likely is indicative of different ionospheric conditions. It is also possible that the satellite loop was oriented for more nearly optimum response in that case.

SIGNALS RECEIVED FROM NPG

During the middle of March 1961, Naval Radio Station NPG (Jim Creek, near Seattle, Washington) transmitted signals at 18 kc for LOFTI I. The radiated power of the transmitter was 200 to 250 kw. A special keyer produced the keying waveform shown in the lower trace of Fig. 9. This is a 45-second sample of the signal as it appeared March 14, 1961, about 37 minutes after midnight EST, in the detector output of the ground VLF receiver at Stump Neck.

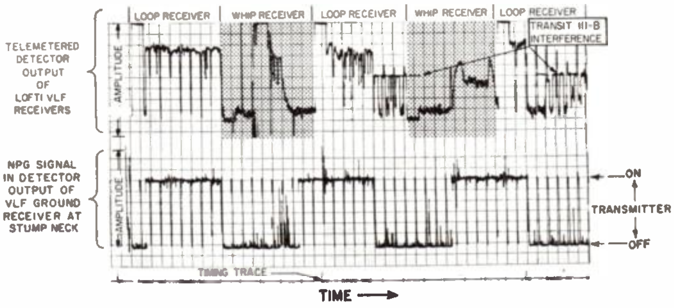

Fig. 9—Reception from transmitter to north of satellite; LOFTI I loop and whip receiver detector output compared to detector output of Stump Neck VLF ground receiver. Orbit revolution 313, March 14, 1961, 12:37:20 to 12:38:05 A.M. EST. Satellite over Florida; height, approximately 767 km; ground range from transmitter (NPG, State of Washington), approximately 4000 km, east-southeast.

The keyer switched NPG's carrier wave alternately on and off for 7½ seconds each interval. During the "on" time, it interrupted the carrier briefly at 1-second intervals, resulting in the downward spikes seen one second apart in the graph. Similarly, the keyer switched the carrier on briefly seven times in each "transmitter-off" period. This keying pattern could be easily identified in the presence of noise and interference.

The upper trace shows the signal as it appeared in the detector output of the LOFTI I receivers. The satellite was moving in its 313th orbital revolution and at that time (12:37 P.M. EST) was approximately 3850 km (2400 miles) east-southeast of NPG (ground range) and about 767 km (475 miles) high. The keying spikes seen in the output of the ground receiver also appear (at a slightly later time) in the output of the satellite's loop receiver. Except for TRANSIT interference at the satellite, the SNR at the satellite and the ground installation do not differ greatly. TRANSIT interference is very evident in two of the three loop sections of the record; however, NPG's signal is clearly evident through the interference.

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[11] Electric field intensity at the loop center at which the loop receiving system of the satellite reached output saturation in calibration measurements made on the ground.

SIGNALS RECEIVED FROM NBA OVER AUSTRALIA

Fig. 10 is a 23-second section of telemetry record from the telemetry station in Australia. It shows the final IF amplifier output of LOFTI's loop receiver during orbit number 130 (March 2, 1961) as the satellite passed over Australia at a height of 725 km (450 miles). LOFTI's ground range to the transmitter at Panama was then close to 16,800 km (10,400 miles). The time was close to 3:30 A.M. Australian standard time (night) at the satellite (about 2:30 P.M. EST (day) at NBA). The NBA keying waveform has been sketched in below the record as an aid to recognition of the one-second time pulses. Although the mean SNR at the loop is estimated as approximately 10 db in this sequence, large cycle-to-cycle variations are apparent in the IF output wave, indicating that rapid fluctuations of signal and noise field intensity have been encountered as the satellite raced along its path.

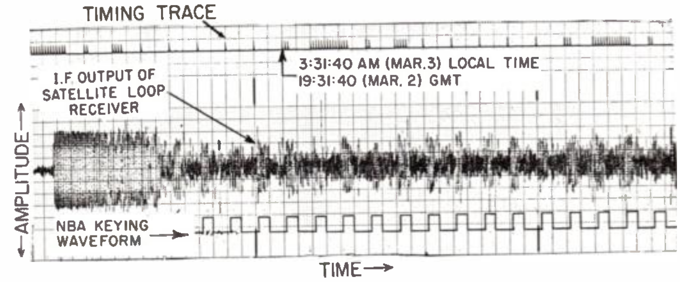

Fig. 10—Reception 10,400 miles from VLF transmitter; output of LOFTI I loop receiver (final IF amplifier). Orbit revolution 130, March 2, 1961, 3:31:32 to 3:31:54 A.M. Australian Standard Time. Satellite over Australian area; height, approximately 725 km; ground range from transmitter (NBA, Panama), approximately 16,775 km, west by southwest.

VLF SIGNAL ECHOES IN THE IONOSPHERE

Fig. 11 shows the final IF amplifier output of the satellite loop receiver as recorded shortly before 7 A.M., EST (about dawn) on March 22, 1961, during orbit number 434. The satellite was approximately 385 km (240 miles) high and at about 2100 km (1300 miles) ground range north of NBA. The envelope of the NBA time pulses as seen in the detector output of the telemetry station's VLF ground receiver appears directly below the satellite IF output trace. A strong echo (marked E) can be seen in the output of the satellite receiver about two thirds of a second after the onset of the first time pulse T_1. A similar echo appears following each time pulse T_2 through T_9 except for the period between E_2 and E_5, which is obscured by noise. A smaller echo, marked $2E_5$, may be the result of a second cycle of reflection of the time pulse (T_5) transmitted about one and a third seconds earlier.

The echoes are probably the result of whistler-path signal propagation. Attenuation in this mode is not very great, as is indicated by the small differences in amplitude of the echoes as compared to the directly received pulses. The most likely path for the VLF wave in this mode is southward along the contour of the geomagnetic field, then back north in the same fashion after reflection and change of polarization at the upper face of the ionosphere south of the geomagnetic equator. However, traversal of the full length of a geomagnetic line at free-space propagation velocity, even from as far north of NBA as 40° geometric latitude, would account for only a small fraction of the time delay of the echoes. Most of the delay must be the result of slower propagation velocity of the VLF wave, an effect to be expected in the ionized medium.

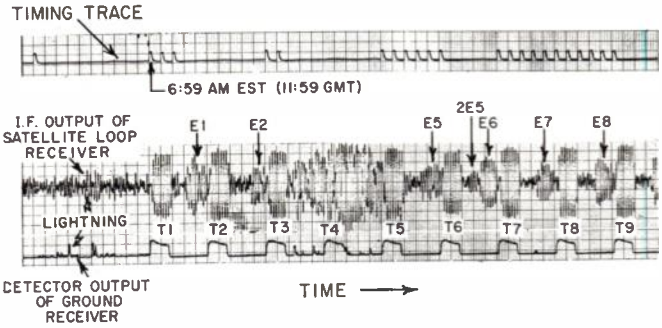

Fig. 11—VLF echoes in the ionosphere. Signal output of LOFTI I loop receiver (final IF amplifier). Orbit revolution 434, March 22, 1961, 6:58:58 to 6:59:08 A.M. EST. Satellite off east coast of Florida; height, approximately 385 km; ground range from transmitter (NBA, Panama), approximately 2100 km north. T_1 to T_8 are time pulses of NBA as received by the satellite, E_1 to E_8 are primary echoes of these pulses and 2E5 is probably a secondary echo.

SFERICS IN THE IONOSPHERE

The random short-duration pulses preceding the first time pulse of the ground receiver trace in Fig. 11 are identified on the tape record as "lightning." Considerable noise appears in the IF output of the satellite loop receiver at the same time. About 3 seconds (and 14 miles) later, another, much larger, noise burst at the satellite largely obliterates signals in the vicinity of the fourth time pulse, even though the ground station trace here shows less noise than associated with lightning in the previous burst. It appears likely that the source of the second burst was much closer to the satellite.

SIGNAL AND NOISE DURING A DAYLIGHT PASS

Fig. 12 shows the *apparent* electric field intensity[12] (and the corresponding magnetic field intensity) of the NBA signal and of the ambient noise level as indicated by LOFTI's loop receiver output during a 6400 km long (4000 mile) daylight pass (March 21, 1961, orbit number 421). The principal sources of TRANSIT interference were silent. Over 800 data points provided the

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[12] The electric field intensity (microvolts per meter) which produced the same output from LOFTI's loop receiving system when the satellite was on the ground as was subsequently observed by telemetry when the satellite was in orbit. The ratio of electric field intensity in the ionosphere to that in free space for the same power level is a function of the ratio of the wave impedances in the two media. The loop antenna responds only to change of magnetic flux ($d\phi/dt$) in either medium. The magnetic field intensity (microampere-turns per meter) may be determined for each medium by dividing the apparent microvolts per meter values by the impedance of free space (377 ohms). However, the relative power level of the signals in the two media cannot be calculated until measurements of wave impedance in the ionosphere have been made.

second-by-second outline of each of the three curves, which respectively define the quasi-maximum apparent signal intensity (upper level), the quasi-minimum apparent signal intensity (lower level) and the quasi-peak apparent intensity of ambient noise at the satellite (dotted curve). The curved abscissa at the one microvolt-per-meter level of the ordinate defines the ground track of the satellite on the earth's surface as it moves in the ionosphere from out over the Pacific Ocean into the Atlantic region.

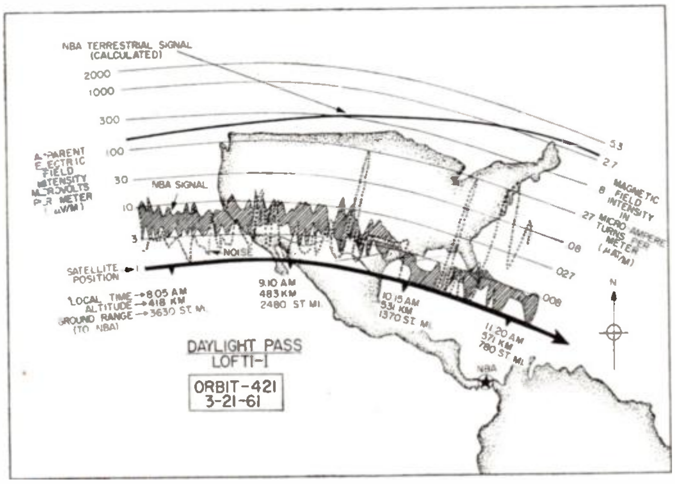

Fig. 12—VLF signal and noise field intensity at the satellite during 17 minutes of a forenoon pass. Upper boundary of signal trace is second-to-second quasi-maximum level; lower boundary is quasi-minimum level. Noise values are quasi-peak levels.

Local mean sun time is marked at the five minute elapsed time points along the track. The satellite is moving in daylight all the way, approaching high noon at the eastern end of the pass. Its altitude increases from 420 km (260 miles) at the west end to 565 km (350 miles) at the east end. At the same time, its ground range to the transmitter decreases from over 5800 km (3600 miles) at the first time marker to less than 1300 km (300 miles) at the last marker.

The field intensity of NBA's signal on the terrestrial surface directly under the satellite, as computed from a modified form of the Baldwin-McDowell equation, is shown by the smooth curve in the upper part of the graph. At the western end of the pass, the ratio of the apparent electric field intensity of the NBA signal at the satellite to the NBA computed electric field intensity at the Earth's surface is about −26 db (±5 db). At this point, the atmosphere is in a state of ionization corresponding to about two hours of solar irradiation. At the eastern end of the pass, the ionization is approaching the noon values; here the ratio is about −55 db (±5 db).

The quasi-peak noise level is below the signal level by roughly 6 to 12 db during most of the transcontinental part of this pass. Occasional large increases of noise appear in the data but persist for only a few seconds. However, beyond the Gulf of Mexico, over the Caribbean area, the noise level is much higher than the NBA signal level much of the time.

SIGNAL AND NOISE DURING A DAWN PASS

Fig. 13 is a presentation of the same type as Fig. 12 but for a dawn[13] pass (March 22, 1961, orbit number 434). The apparent signal level at the satellite relative to the computed terrestrial surface value drops rapidly as LOFTI and its companions move from the west toward sunrise at the satellite (from about −26 db ±5 db to about −46 db ±5 db). Immediately after it passes the sunrise line, the satellite moves into a region of rapidly increasing VLF field intensity. Loop receiver saturation level (indicating signals of over 100 μv/m apparent electric field intensity) is reached in about two minutes elapsed time. The apparent electric field intensity is then less than 15 db below the computed surface value; the satellite is at a position corresponding to about six hours before noon. From there on, the signal decreases with further satellite motion to the east, reaching a relative value of about −40 db ±5 db at the end of the pass. The position then corresponds to about four hours before noon.

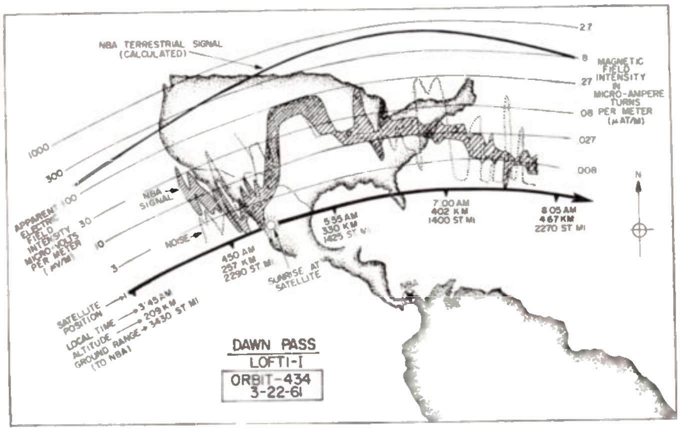

Fig. 13—VLF signal and noise field intensity at the satellite during 15 minutes of a dawn pass.

The ambient noise level is higher throughout most of this pass than in the daylight pass of Fig. 12, exceeding the signal level very considerably shortly before and after the satellite passes the sunrise position. Again the noise level exceeds the NBA signal level much of the time over the Caribbean area.

EFFECT OF DIRECTION OF SATELLITE FROM TRANSMITTER

Comparison of NPG signal data to NBA signal data affords some experimental confirmation of the effect of difference in angle θ of the wave normal relative to the geomagnetic field on attenuation. NPG is about 40° farther north than NBA. Limited data available for a period when the two stations were operating simultaneously at 18 kc indicates that the NPG signal in the ionosphere was attenuated relative to its computed ground signal intensity roughly 10 db less than NBA's signal.

[13] Similar data for night and dusk passes were not yet available at the time of this writing.

The computed effect of θ on power loss through the model ionosphere of Fig. 2 is shown below.

Angle θ Between Wave Normal and Geomagnetic Field	Computed Transmission Loss Through Ionosphere	
	Night	*Day*
0° (or 180°)	2.0 db	27 db
30 (or 150°)	3.5	33
60 (or 120°)	15	75
90	>100,000	>100,000

TIME DISTRIBUTION OF SIGNAL INTENSITY

The time distribution of the apparent field intensity of the NBA signals observed by LOFTI in a particular portion or volume of the atmosphere is shown in Fig. 14 relative to computed ground signal intensity. To develop the data in this form, the second-by-second values for a number of passes of the satellite through this volume were reduced to one-minute mean values.[14]

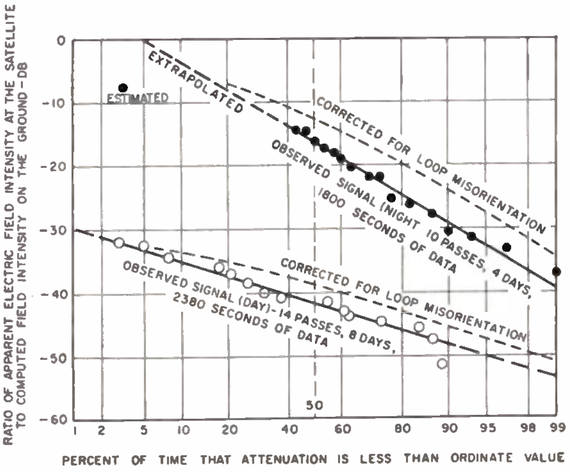

Fig. 14—Cumulative time distribution of NBA signal intensity at the satellite in a space to the north of NBA approximately 4400 km long (west to east) by 890 km wide (north to south) by 555 km high, and 165 km above the terrestrial surface at closest approach. Satellite at extreme line-of-sight distance from transmitter most of the time.

The volume to which the distribution curves of Fig. 14 apply was to the north of station NBA and extended from 20 to 28° N latitude and 60° to 100° W longitude. The satellite was at altitudes between 380 and 720 km (225 and 445 statute miles) during the daytime passes and between 165 and 360 km (105 and 224 miles) during the night-time passes. The particular passes and this volume were chosen because the greatest amount of reduced data for transmitter to satellite distances close to the extreme line-of-sight range was available for them, as of this writing.

The night curve in Fig. 14 presents data from 10 passes over a period of four days. Altogether, these passes provided 1800 seconds of telemetered information. Essentially all of the NBA pulses are visible in the telemetry recordings of these passes, except for a few moments when noise bursts obscured the signals briefly or when telemetry reception failed momentarily. Similarly, the day curve is derived from 14 passes on 8 days that together provide 2380 seconds of data. About 10 per cent of the time ticks are completely masked by VLF noise in these recordings. The designation "night" as applied in Fig. 14 is the interval of one and a half hours after sunset to one and a half hours before sunrise. Likewise, "day" is the period from one and a half hours after sunrise to one and a half hours before sunset.

The one-minute mean data points for both day and night cluster along lines of log-normal distribution. The night curve has no data points below 15-db attenuation because the loop receiver reached output saturation when the ionospheric attenuation was less than this value. The one point shown at about 8 db (marked "estimated") was derived from observations of output signal magnitude during periods when the 40-db pad was inserted in the satellite's loop circuit. It is not considered to be as accurate as the rest of the data and the night curve has therefore been extended by straight line extrapolation. The lines marked "corrected for loop misorientation" represent the observed data compensated, on a probability basis, for the lower loop output resulting from the likely misalignment of the loop relative to the VLF wave front.[15]

DISCUSSION OF DISTRIBUTION CURVES

The range of amplitude variation of the observed night signal in the 3 to 99 per cent time range, including the estimated point at 8-db attenuation, is about 30 db. The 3 to 99 per cent range for the observed day signal (including the extrapolated section from 90 to 99 per cent) is about 21 db. These values compare fairly well with the range of fading observed in long-distance VLF ground signal propagation studies.

The difference in attenuation between the day and night curves at the 50 per cent time points is about 26 db. This is nearly the same as the difference between the computed day and night absorption loss (cumulative attenuation) above 160 km (100 miles) altitude in the model ionosphere (Fig. 4).

The attenuation of magnetic field intensity of the VLF radio wave is indicated directly by the ordinate in Fig. 14. It appears that 50 per cent of the time the magnetic field intensity of the signal in the ionosphere at night is 13 db less than the computed ground field and is 38 db less during the daylight period. These figures apply to the particular conditions under which

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[14] In general, the field intensity at the satellite loop, as gauged from final IF output level telemetered from the satellite, did not vary much more than about three to one in any one minute interval. Each minute of data was examined second by second and a microvolt-per-meter level was assigned to it which represented a best estimate of the mean level for that one minute interval. Although the satellite loop receiver saturated at 100 μv/m, programmed 2-second intervals of reduction of loop signal input to the receiver occurred once each 30 seconds. During this time, a 40-db attenuator was automatically inserted into the loop circuit by the satellite instrumentation. The resulting increase of input required to produce receiver saturation has made possible a rough estimate of the field intensity causing the saturated condition immediately prior to and succeeding these 2-second intervals.

[15] See Appendix IV.

the data was taken and include correction for probable loop misorientation.

Since the interface entry and the subsequent absorption losses are both dependent to some degree on the angle between the wave-normal and the geomagnetic field, becoming less as θ approaches 0 degrees, the attenuation of the signal by the ionosphere should, in general, be less at the higher latitudes.

TIME DELAY AND DOPPLER SHIFT

The low velocity of VLF wave propagation in the ionosphere indicated by the long delay time of the echoes in Fig. 11 is also apparent in the relatively much longer time required for the signal to arrive at the satellite as compared to arrival time at the ground VLF receivers. Time pulse delays ranging from about 10 to 200 msec appear throughout the data. As much as 8 cps Doppler shift of NBA's carrier frequency, which is held constant at the transmitter to 1 part in 10^{10} (about $\pm 2\times 10^{-6}$ cps), appears in the data so far examined.

EFFECT OF ALTITUDE

As of this writing, about one sixth of the LOFTI I data has been reduced. No correlation between altitude of the satellite and the apparent signal intensity has yet become evident. The data so far studied does not disagree with the concept that the absorption loss occurs almost entirely in the *D* region.

CONCLUSIONS

General conclusions which are derived before all the significant data from an experiment has been reduced and analyzed are necessarily tentative. However, the LOFTI I data and studies do indicate that:

$$n^2 = (\mu - i\chi)^2 = 1 - \frac{X}{1 - iZ - (1/2)Y_T^2/(1 - X - iZ) \pm \left[\frac{1}{4}Y_T^4/(1 - X - iZ)^2 + Y_L^2\right]^{\frac{1}{2}}} \qquad (2)^{16}$$

1) A considerable portion of the energy radiated by transmitters and other sources of VLF signals in the lower atmosphere penetrates the interface between the ionosphere and lower atmosphere and appears within the ionosphere. It seems likely that there is a transition loss at the interface, after which the signal energy is further reduced by attenuation due to absorption plus some spreading loss as the wave moves upward through the ionized air.

2) Most of the propagation loss between the ground transmitter and satellite receiver appears to occur below 100 km altitude.

3) The signal intensity data so far reduced to statistical form indicates that 50 per cent of the time the magnetic field intensity of the VLF wave is reduced less than 13 db at night and less than 38 db by day because of passage through the ionosphere. The corresponding figures for 10 per cent and 90 per cent of the time are 4 db and 29 db at night, and 33 and 45 db by day. This information is, however, based on only about 2000 seconds of data each for the night and day conditions, and applies only to 18-kc signals received by the satellite at latitudes less than 28° north and to the north of the transmitter.

4) The VLF signal appears to travel to its point of reception in the ionosphere much more slowly than in free space. Transmission delay as great as thirty times the free space delay appears in the data.

5) The echoes of station NBA signals observed in the experiment lend support to present theories of whistler propagation.

6) Satellite communication using VLF radio waves is feasible and affords some interesting possibilities for future application.

FUTURE PLANS

Further LOFTI experiments are planned for exploration of VLF reception at the higher latitudes and at other frequencies. Transmission from within the ionosphere is to be studied. It is expected that, in addition to other data, the next experiment will provide the information on electric dipole impedance in the ionized medium lost in the LOFTI I experiment.

APPENDIX I

ATTENUATION OF A RADIO WAVE PROPAGATING IN A MAGNETO-IONIC MEDIUM

The absorption coefficient χ for an electromagnetic wave in a magneto-ionic medium can be determined from the complex refractive index n.

where

μ = refractive index,
$X = \omega_N^2/\omega^2$,
$Y_L = \omega_L/\omega$,
$Y_T = \omega_T/\omega$, and
$Z = \nu/\omega$.

The quantities

$\omega_N^2 = 4\pi Ne^2/\epsilon_0 m$,
$\omega_L = \omega_H \cos\theta = (\mu_0 H_0 e/m) \cos\theta$,
$\omega_T = \omega_H \sin\theta = (\mu_0 H_0 e/m) \sin\theta$,
ν = collision frequency, and
ω = wave angular frequency;

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[16] J. A. Ratcliffe, "The Magneto-Ionic Theory and Its Application to the Ionosphere," Cambridge University Press, New York, N. Y., and London, Eng.; 1959.

in which

ω_N = plasma angular frequency,
ω_H = gyro angular frequency,
N = electron density,
H_0 = magnitude of the imposed magnetic field,
θ = angle between the direction of the magnetic field and the wave normal,
e = charge of the electron,
m = mass of the electron,
μ_0 = magnetic permittivity of free space,
ϵ_0 = electric permittivity of free space.

An expression for α, the power attenuation of the radio wave per unit distance in the ionosphere, may be derived from χ in equation (I-1). A greatly simplified form of this expression appears as (1).

APPENDIX II

VARIATION OF ORBITAL PARAMETERS

Orbit Attained

Certain major orbital parameters of LOFTI I and its attached companions four days after launch are given in Table I. This data was developed from Minitrack station observations by the NASA Computing Center and issued by the Goddard Space Flight Center.

TABLE I

MAJOR ORBITAL ELEMENTS OF SATELLITE 1961 ETA FOR THE EPOCH 0353 UT (10:53 PM, EST), FEBRUARY 25, 1961

Anomalistic Period (Time for one revolution around Earth, perigee to perigee)	95.92030	Minutes
Inclination (Plane of orbit relative to Earth's equatorial plane)	28.37	Degrees
Argument of Perigee (Angle between ascending node and perigee in plane of orbit)	60.392	Degrees
/Motion Plus (Progression of perigee)	10.676	Degrees per day
Eccentricity (Departure of orbit from perfect circle)	0.05731	
Semi-Major Axis (Half-length of major axis)	1.08826	Earth Radii
Perigee [Least altitude relative to equatorial radius of earth (=3963.3 statute miles)]	103	Statute Miles
Apogee (Greatest altitude relative to equatorial radius of earth)	597	Statute Miles
Velocity at Perigee	17,953	Miles per hour
Velocity at Apogee	16,007	Miles per hour

Progression of the Apogee

Although 1961 Eta's perigee became lower by only a few miles during its short life, its apogee progressively fell from almost 600 miles to nearly 250 miles, as shown in the upper part of Fig. 15.

Table I gives the rate of progression of the perigee as very nearly $10\frac{2}{3}$ degrees per day at that particular epoch in time. This rate did not change greatly during the nearly $36\frac{1}{3}$ days of the satellite's life and the apogee progressed similarly. The change in sub-orbital position[17] of the apogee is shown in the lower part of Fig. 15.

The progression in position of the apogee and perigee over a complete cycle in latitude allowed LOFTI I to sample the VLF field at varying altitude over all the areas covered by telemetry before 1961 Eta perished, presumably in a burst of fire, on March 30, 1961.

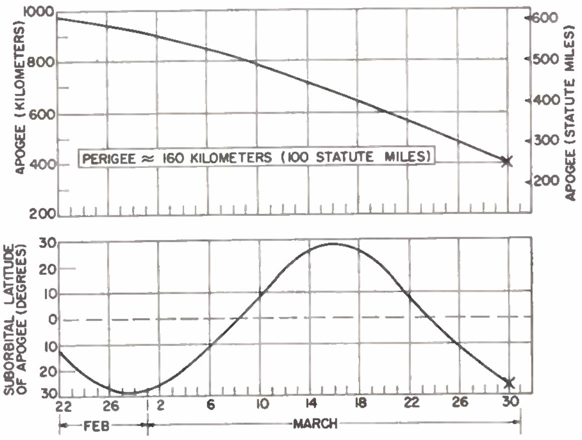

Fig. 15—Reduction in altitude of the apogee and variation of the apogee in latitude during the lifetime of the satellite. The altitude of the perigee remained close to 160 km until the end of life.

APPENDIX III

SATELLITE INSTRUMENTATION[18]

The Satellite

LOFTI I (Fig. 16) was a 20-inch diameter sphere that weighed 57 pounds complete with instrumentation. A loop antenna of approximately elliptical shape, which had a mean diameter of 20 inches, was mounted on the sphere. The satellite's two 15-foot long whip antennas were wound in strip form on separate motor-driven reels inside the sphere. The strips were to be unreeled upon radio command, assuming a hollow-rod form as they extended slowly through the upper guide tube and an orifice in the base of the sphere, respectively.[19] Six sets of solar cells assembled in patches on the surface of the sphere provided primary power for the electrical and electronic equipment. Eight photo-electric sensors were mounted on the sphere, four to provide data on satellite attitude relative to the sun and four to provide similar information on attitude relative to the Earth. Four rods, which were folded up during launch, were

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[17] Point on Earth's surface vertically under the satellite.

[18] The Radio Techniques Branch, Radio Div., USNRL, planned and directed the experiment, and provided the satellite VLF and ground station instrumentation. The Satellite Techniques Branch, Applications Res. Div., USNRL (M. Votaw, Branch Head) provided the satellite, including the VHF instrumentation, and was responsible for launch.

[19] H. R. Warren, "DeHavilland antenna erection unit," *Proc. of Fifth National Convention on Military Electronics*, Washington, D. C.; June 26–28, 1961.

the radiator elements of a turnstile antenna for the 136-Mc telemetry system.

Electronic Instrumentation

The block diagram (Fig. 17) shows the major electronic elements of LOFTI I. The telemetered information was carried by frequency variation of subcarriers in Inter-Range Instrumentation Group (IRIG) frequency bands 3 to 7 (inclusive), which amplitude-modulated the carrier of the 136.17-Mc telemetry transmitter. The divided-down (1057 cps) output of the individual sync-locked crystal oscillators of the two VLF receivers also amplitude-modulated the telemetry carrier, one of the 1057 cps outputs being considerably attenuated relative to the other to permit recognition in the telemetry in case of loss of synchronization. The term "25 cps IF output" in the diagram refers to the output of the final stage of the second (25 cps) IF amplifier of the VLF receiver, and the term "detected output" refers to the output of the diode detector following the first IF (1082 cps) amplifier.

Pickaback Mounting

Fig. 18 shows LOFTI I as it appeared mounted as a copassenger on the TRANSIT III-B satellite. The two-satellite assembly was attached to the second stage of the Thor/Able-Star launch vehicle and covered by a nose cone during launch.

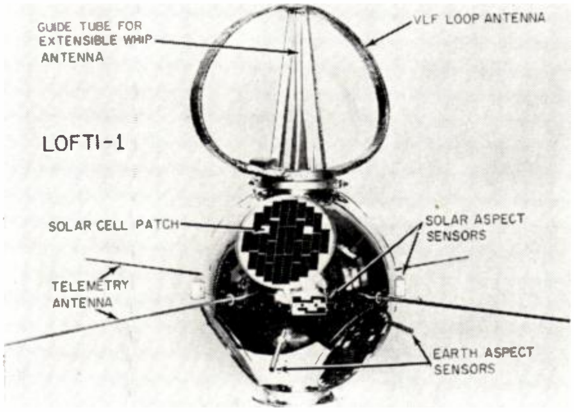

Fig. 16—Side view of the LOFTI I satellite.

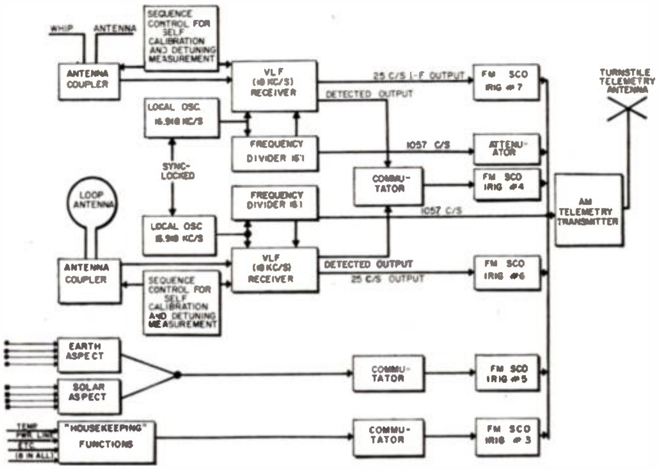

Fig. 17—Major electronic elements of LOFTI I satellite instrumentation which pertained to the VLF experiment.

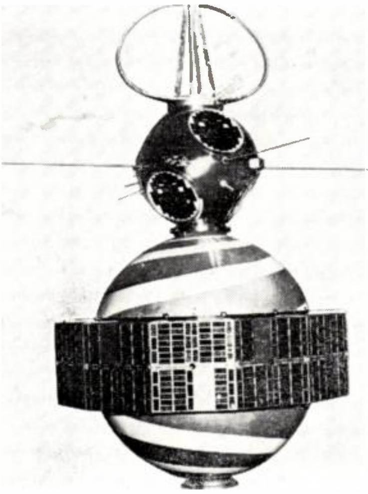

Fig. 18—Side view of LOFTI I mounted on TRANSIT III-B.

APPENDIX IV

DECREASE IN OUTPUT OF A SINGLE PLANE LOOP DUE TO MISORIENTATION RELATIVE TO THE WAVE NORMAL

Assuming circular polarization of the VLF wave in the ionosphere, the LOFTI loop would exhibit cosine-law response when rotated about any axis perpendicular to the wave normal. The loss due to misorientation would then be:

Misorientation (Per cent of 90°)	Corresponding Reduction in Loop Output Voltage
0 per cent	0 db
20	0.5
30	1
40	1.5
50	3
60	4.5
70	7
80	10
90	16
95	22
99	36

ACKNOWLEDGMENT

The LOFTI I program has been sponsored by the Special Projects Office through the Bureau of Ships and conducted by the Radio Division of the U. S. Naval Research Laboratory with the cooperation of the Applications Research Division. A major effort of this nature is very much dependent on the devotion, knowledge, and skill of many persons; unfortunately it is impractical to name them here, but the authors are deeply grateful to them all. The generous and expert cooperation of the National Aeronautics and Space Administration and its personnel is gratefully acknowledged. Discussions with R. A. Helliwell and L. H. Rorden of Stanford University and the Stanford Research Institute, respectively, have been very helpful in LOFTI planning.

Decreasing the Threshold in FM by Frequency Feedback*

L. H. ENLOE†, MEMBER, IRE

Summary—**The "frequency feedback demodulator" or "frequency compression demodulator" can be used to extend the threshold of signal-to-noise improvement in large index frequency-modulation systems. Previous papers have advanced the argument that the threshold occurs in the usual manner when noise peaks exceed the carrier envelope at the input to the frequency detector of a feedback demodulator. However, correlation between calculated values and carefully measured experimental values has been poor. The calculated values have been incorrect by orders of magnitude in many typical cases. In this paper it is shown that the threshold can also occur because of the feedback action of the demodulator. When this is taken into account, the threshold can be calculated accurately. It is shown that the threshold cannot be improved by the often suggested scheme of inserting a carrier of the proper phase at the input to the frequency detector. The phase-locked loop, a related device, is shown to have a threshold which is equal to or poorer than the threshold of the feedback demodulator for large modulation indices. It is shown that the stability of the feedback loop (and consequently the threshold) of the feedback demodulator is a function of modulation, but that by following the procedure given, the effect can be almost entirely eliminated. The results of this paper allow one to design feedback demodulators for the first time which will extend the threshold in a predictable manner.**

INTRODUCTION

THERE is widespread interest at the present time in modulation techniques which trade bandwidth of the transmitted signal for improved baseband signal-to-noise and signal-to-interference ratios. These techniques are of special interest in the field of space communications where distances are so large that even the use of masers and large low-noise antennas results in a signal-to-noise ratio, assuming single-sideband modulation, too small for commercial telephone service by a factor of about a hundred. In addition, the most efficient operation requires the use of a single-frequency channel by many satellites and ground stations simultaneously. This mode of operation demands a modulation technique which can increase the baseband signal-to-mutual-interference ratio considerably over that of single-sideband modulation. Large index frequency modulation satisfies both of these requirements and could be used but for the fact that it has a relatively poor threshold. *The threshold is defined as the minimum carrier-to-noise ratio yielding an FM improvement which is not significantly deteriorated from the value predicted by the usual small noise signal-to-noise formulas.* This paper discusses special demodulators using frequency feedback to extend the threshold and bring frequency modulation into an attractive position.

The frequency feedback demodulator originated with Chaffee [1] in the early thirties. His paper and a companion paper by Carson [2] contain interesting results concerning the distortion and small noise signal-to-noise characteristics of the demodulator. In the past few years various papers have been published [3], [4] to show that a form of this demodulator (a limiter was added) could be used to decrease the threshold of signal-to-noise improvement in large index FM. More recently, demodulators of this type were used in the Project Echo experiment for the same purpose [5]. In these demodulators the frequency deviation of the large index wave is compressed by using feedback, so that it may be passed through a narrow-band-pass IF filter before demodulation. Previous papers [3], [4] have advanced the theory that the system threshold occurs in the usual manner at the frequency detector of the feedback demodulator when noise peaks exceed the carrier, and is equal to that of a conventional FM demodulator having the same narrow-band-pass IF filter. This implies that, if the system were above threshold with no feedback applied, an unlimited signal-to-noise ratio could theoretically be obtained by increasing the frequency deviation of the carrier and the amount of feedback simultaneously. In practice it has been found that this behavior cannot be obtained, and it has also been found that the threshold can be different from the predicted value by orders of magnitude. The tendency has been to attribute these inconsistencies to the difficulties involved in obtaining a stable feedback system for large amounts of feedback. It will be shown in this paper that the more fundamental reason is that the threshold deteriorates as a function of the feedback action of the system. A mathematical analysis of the feedback demodulator when it is operating *below* threshold is a difficult and unsolved problem. The determination of where the threshold occurs, however, is a considerably less difficult problem, and it is this problem which is considered. In the past it has been suggested [4] that the threshold of the feedback demodulator could be eliminated by properly inserting a carrier in front of the frequency detector. In this paper it is shown that in a properly designed system the threshold would not be improved by carrier insertion. It is also shown that the feedback stability and the threshold of the feedback demodulator are both functions of modulation, but that by following the suggested procedure the effect can be almost entirely eliminated. The results of this paper allow one, for the first time, to design feedback demodulators which will extend the threshold in a predictable manner.

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I. The Concept of Frequency Feedback

The block diagram of the feedback demodulator is shown in Fig. 1. In order to understand its operation, imagine for the moment that the voltage controlled oscillator (VCO) is removed from the circuit and the feedback loop left open. Assume that a wide index FM wave is applied to the input of the mixer, and a second FM wave, from the same source but whose index is a fraction smaller, is applied to the VCO terminal of the mixer. The output of the mixer would consist of the difference frequency, since the sum frequency components are removed by the bandpass filter. The frequency deviation of the mixer output would be small, although the frequency deviation of both input waves is large, since the difference between their instantaneous deviations is small. Hence, the indexes of modulation would subtract and the resulting wave would have a smaller index of modulation. The reduced index wave may be passed through a filter, whose bandwidth need be only a fraction of that required for either large index wave, and frequency detected. It is now apparent that the second FM wave could be obtained by feeding the output of the frequency detector back to the VCO.

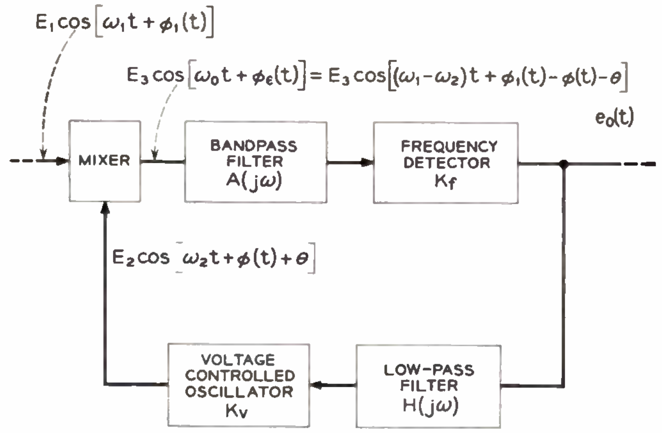

Fig. 1—Block diagram for a frequency feedback demodulator. K_f and K_v are the gain constants of the frequency detector and the VCO respectively. They relate radian frequency to voltage. $A(j\omega)$ and $H(j\omega)$ are the transfer functions of the band-pass IF filter and low-pass baseband filter in the feedback path respectively. Although the output is indicated as being immediately behind the frequency detector, it might equally well be considered to be any other point in the loop.

II. Small Index Response and Signal-to-Noise Ratio Above the Threshold

The "moving finger" or quasi-stationary behavior of the feedback demodulator is more or less obvious, and so our attention will be concentrated on the less obvious small index behavior. It is the small index response which may be used to determine the system stability, the effective or closed-loop noise bandwidth, the baseband signal-to-noise ratio and the threshold. The baseband analog of the demodulator, shown in Fig. 2, is valid as long as the modulation index of the wave entering the band-pass IF filter is small compared to unity.[1] Under this condition the IF filter may be represented by its low-pass equivalent.[2] The VCO and frequency detector may be replaced by an ideal integrator and an ideal differentiator respectively. The mixer is a frequency subtractor. The open-loop transfer function $K_v K_f A_L(j\omega) H(j\omega)$ must, of course, satisfy the usual Nyquist and other stability criteria.[2] This eliminates any possibility of using a so-called "rectangular" IF filter. Let us define the closed-loop transfer function as the function which relates the phase of the wave generated by the VCO to the phase of the wave at the signal input to the mixer, *i.e.*, $\phi(j\omega)\phi_1(j\omega)$ in Fig. 2. *From linear feedback theory it follows that the closed-loop bandwidth is unavoidably larger than the open-loop bandwidth.* In later sections we shall see that this fact plays an important role in the determination of the threshold.

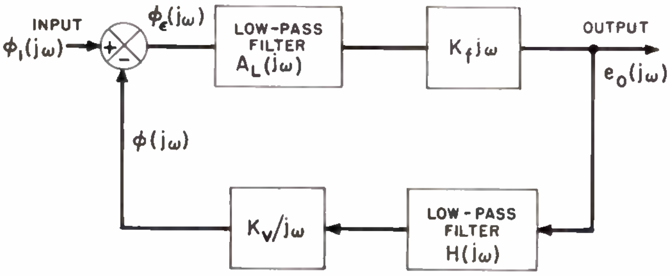

Fig. 2—Baseband analog of the feedback demodulator when the wave entering the IF band-pass filter has a modulation index much smaller than unity, *i.e.*, when $|\phi_\epsilon(t)| \ll 1$. $A_L(j\omega)$ is the low-pass equivalent of the band-pass IF filter $A(j\omega)$. $\phi_1(j\omega)$, $\phi_\epsilon(j\omega)$, $\phi(j\omega)$ and $e_0(j\omega)$ are the Fourier transforms of $\phi_1(t)$, $\phi_\epsilon(t)$, $\phi(t)$ and $e_1(t)$, respectively (shown in Fig. 1).

It will now be shown that the signal-to-noise ratio of a feedback demodulator is the same as that of a conventional FM demodulator receiving the same signal and noise density if the carrier-to-noise power ratio is sufficiently large. Assume for the moment that there is no feedback, *i.e.*, that $K_v = 0$. Let the unmodulated carrier and noise at the input to the frequency detector be given by [6]

$$
\begin{aligned}
e(t) &= \cos\omega_0 t + I_c(t)\cos\omega_0 t - I_s(t)\sin\omega_0 t \\
&= R_e[1 + I_c(t) + jI_s(t)]\epsilon^{j\omega_0 t} \\
&= R_e\sqrt{[1 + I_c(t)]^2 + I_s^2(t)} \\
&\qquad \epsilon\left[j\omega_0 t + j\tan^{-1}\frac{I_s(t)}{1 + I_c(t)}\right],
\end{aligned}
$$

---

[1] It should be noted that the modulation indexes of the waves at the VCO output and mixer signal input may be large. These waves have many sidebands for each baseband frequency component, but they are correlated in such a manner that when they are multiplied together in the mixer a cancellation process takes place which results in a carrier and a single pair of quadrature sidebands per baseband component.

[2] This will be discussed further in Section V and in Appendix C.

where $R_e(\ )$ stands for the real part of ( ). $I_c(t)$ and $I_s(t)$ are random Gaussian variables in phase quadrature with each other, normalized by the carrier envelope. The last equation places in evidence the envelope and phase of the composite wave.[3] The constant frequency term ω_0 can be subtracted out by a balanced frequency detector or eliminated by filtering, and so will be dropped. If the envelope of the noise is much smaller than the carrier envelope except for a negligible portion of time, *i.e.*, if

$$|I_c(t) + jI_s(t)| \ll 1,$$

then the phase and envelope of the composite wave are given by

$$\Psi_1(t) \doteq I_s(t)$$

and

$$E_1(t) \doteq 1 + I_c(t).$$

These expressions show that the composite wave at the input to the frequency detector is small index phase-modulated by the term $I_s(t)$, which is derived from the component of noise which is in phase quadrature with the carrier. The wave is also small index envelope-modulated by the term $I_c(t)$, which is derived from the component of noise which is in phase with the carrier. When feedback is applied, the VCO generates a wave which reduces the angle-modulation index of the wave in the IF, *i.e.*, the quadrature component of noise $I_s(t)$. Thus it is seen that, as long as the carrier-to-noise ratio is sufficiently large, the demodulator does not respond to the in-phase noise (envelope of the composite wave), but that it demodulates the quadrature noise in exactly the same fashion as it would demodulate signal modulation. Signal and quadrature noise are reduced in the same proportion by feedback, with the result that the baseband signal-to-noise ratio is independent of feedback. For large carrier-to-noise ratios the baseband signal-to-noise ratio of a feedback demodulator is then the same as that of a conventional FM demodulator [8].

III. QUALITATIVE DISCUSSION OF THE THRESHOLD AND EXPERIMENTAL RESULTS

The conventional FM receiver accepts a band of noise at the input to its frequency detector equal to the bandwidth required by the large index transmitted wave. When noise peaks exceed the carrier for any significant portion of time, a threshold occurs. This threshold limits, for a given carrier power and noise density, the baseband signal-to-noise improvement obtainable by increasing the deviation of the transmitted wave. It was shown in Section II that the frequency feedback demodulator reduced the quadrature noise $I_s(t)$ reaching the frequency detector by the use of feedback. While the in-phase noise $I_c(t)$ was not reduced by feedback, it was also not detected by the ideal frequency detector and consequently was not fed around the loop. Feedback then decreases the envelope of the noise at the input to the frequency detector. Based upon this observation, it has been *wrongly* suggested [3], [4] that the threshold is determined by the envelope *of* the noise at the frequency detector and hence should not be degraded by feedback. If this were true, the baseband signal-to-noise ratio could (in theory) be increased indefinitely by increasing the deviation of the transmitted wave and the amount of feedback simultaneously, as long as the system were above threshold with no feedback applied. Unfortunately, we shall see later that the threshold is degraded by the feedback action of the demodulator because of effects which are not at all obvious from the above oversimplified argument.

Let us study briefly the reasons why the feedback demodulator is able to extend the threshold. The conventional large index FM receiver fails to use a very important piece of *a priori* information, that even though the carrier frequency will have large frequency deviations, its rate of change will be at the baseband rate. Both the feedback FM demodulator and the phase-locked loop use this *a priori* information to extend the threshold. They are essentially "tracking filters" which can track only the slowly varying frequency of large index waves, and they consequently respond to only a narrow band of noise centered about the instantaneous carrier frequency. We would expect intuitively that the threshold of an ideal tracking filter would be equal to that of any other frequency detector receiving the same carrier and narrow band of noise. The bandwidth of noise to which the feedback demodulator and phase-locked loop respond is precisely the band of noise which the VCO tracks. The noise bandwidth of both systems is that of the closed-loop response function defined in Section II. While the thresholds of the phase-locked loop and feedback demodulator occur because of the same basic mechanism, the details by which they occur are, of course, different. Let us concentrate now on the feedback demodulator. Assume for the moment that the feedback loop is open and that the demodulator input consists of an unmodulated carrier wave and a band of Gaussian noise. The noise can be separated into components which are in phase and in phase quadrature with the carrier, as discussed in Section II. If the carrier-to-noise ratio at the input to the frequency detector is large, then the composite wave at that point is small index phase-modulated by the quadrature noise and small index envelope-modulated by the in-phase noise. For the purpose of this paper the threshold will be said to occur when the phase of the composite wave at the input to the frequency detector contains a significant amount of noise in addition to that derived from the quadrature noise, *i.e.*, in addition to the noise predicted by the FM improvement formula.

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[3] The phase and envelope of any general wave are uniquely defined by its pre-envelope [7].

Now close the feedback loop. The baseband noise angle modulates the VCO. If the root-mean-square phase deviation of the VCO wave is small, the spectrum consists primarily of a carrier "spike" and the first-order spectral sideband zone. These first-order sidebands are in phase quadrature with the VCO carrier spike. The mixer forms the product of the VCO wave with the demodulator input carrier and noise. The product of the VCO carrier spike and incoming carrier yields a new carrier in the IF. The product of the VCO carrier spike and incoming quadrature and in-phase noise terms yields new quadrature and in-phase noise terms in the IF. The quadrature noise becomes angle noise and the in-phase noise becomes envelope noise. The product of the quadrature term (first-order sidebands) of the VCO with the incoming carrier yields a second quadrature term in the IF which tends to cancel the first quadrature term. It is because of this cancellation that noise in the baseband is reduced by feedback. It is also this cancellation that allows the VCO wave to have a frequency deviation which may be wider than the IF filter bandwidth. If the products mentioned so far are the only ones of any significance, the system behaves in a linear fashion. Two products remain, however. The product of the quadrature term of the VCO wave with the incoming quadrature and in-phase noise terms yields additional angle and envelope noise. *The threshold occurs when this additional angle noise becomes significant.* This happens when the root-mean-square phase deviation of the VCO wave caused by noise is no longer small compared to unity. The exact relations are discussed quantitatively in Section IV. Experimentally it is found that, soon after the additional noise becomes noticeable, noise impulses appear in the baseband. Their onset serves as a convenient measure of the threshold in practical systems. They appear suddenly and increase rapidly in number as the carrier-to-noise ratio decreases. They first appear when the root-mean-square phase deviation of the VCO wave (caused by noise) is $\phi_{\text{rms}}=1/3.11$ radian, *i.e.*, approximately one-third of a radian. They are apparently the result of higher-order nonlinearities which occur in this region.

The threshold should be expressed in terms of the input carrier-to-noise ratio in the closed-loop noise bandwidth in order for its full implication to be appreciated. This is easily done by using the small index baseband analog shown in Fig. 2. The mean-square phase of the wave generated by the VCO, $\phi_{\text{rms}}{}^2$, can be obtained by integrating the product of the input phase spectral power density and the square of the absolute value of the closed-loop response function. Above threshold the input phase spectral power density caused by noise is equal to the spectral power density of the normalized quadrature noise. Thus

$$\phi_{\text{rms}}{}^2 = \sigma_s{}^2 \int_{-\infty}^{+\infty} \left| \frac{K_vK_fA_L(j\omega)H(j\omega)}{1+K_vK_fA_L(j\omega)H(j\omega)} \right|^2 df,$$

where $\sigma_s{}^2$ is the spectral power *density* of the normalized quadrature noise $I_s(t)$, which is also equal to the total input normalized white noise density. The closed-loop noise bandwidth[4] is

$$B_c = \left[\frac{1+K_vK_f}{K_vK_f}\right]^2 \int_{-\infty}^{+\infty} \left| \frac{K_vK_fA_L(j\omega)H(j\omega)}{1+K_vK_fA_L(j\omega)H(j\omega)} \right|^2 df,$$

and the carrier-to-noise ratio in front of the mixer in a bandwidth equal to the closed-loop noise bandwidth is $\rho=1/2B_c\sigma_s{}^2$. Consequently, the input carrier-to-noise ratio in a bandwidth equal to the closed-loop noise bandwidth is given, in terms of the mean-square phase of the VCO wave, by

$$\rho = \frac{1}{2\phi_{\text{rms}}{}^2}\left[\frac{K_vK_f}{1+K_vK_f}\right]^2.$$

As determined by experiment, the threshold occurs at

$$\rho_T = \frac{1}{2(1/3.11)^2}\left[\frac{K_vK_f}{1+K_vK_f}\right]^2 = 4.8\left[\frac{F-1}{F}\right]^2, \qquad (1)$$

where $F=1+K_vK_f=$ amount of feedback or frequency compression. For large feedback $\rho_T=4.8$ or 6.8 db.

Let us dwell upon the full implications of (1), for they are probably the most important results in this paper. This equation tells us where the *feedback threshold* occurs in the demodulator as a function of the feedback factor and the input carrier-to-noise ratio in a bandwidth equal to the closed-loop noise bandwidth of the system. Recall that the development of this equation assumed that the system was above threshold on an open-loop basis. Therefore (1) must be used only after assurance that this condition is satisfied. For instance, the result $\rho_T=0$ when $F=1$ means only that a system without feedback does not have a threshold produced by feedback. *In order for a feedback demodulator to be above threshold, it must be above both the open-loop and the feedback thresholds independently.* Observe that the closed-loop noise bandwidth is not uniquely related to the IF filter bandwidth, baseband bandwidth or the bandwidth required by the large index wave in front of the mixer. However, once the open-loop transfer function $K_vK_fA_L(j\omega)H(j\omega)$ is specified, the closed-loop noise bandwidth is determined uniquely. It will always be larger than the open-loop bandwidth, and for large feedback it will be much larger. Consequently, if the system were designed by placing all the selectivity in the IF filter, then the carrier-to-noise ratio in the IF band in front of the frequency detector would be much larger than that in a bandwidth equal to the closed-loop noise bandwidth (by the ratio of the bandwidths). As a result, the system threshold would be caused entirely by the feedback threshold.

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[4] Notice that power is assumed to reside in both positive and negative frequencies, and that the closed-loop bandwidth is two sided.

Since the closed-loop noise bandwidth does not depend on whether the filtering is done at IF or baseband or a combination of both, it is immediately clear how to improve the performance of the demodulator. First, note that full feedback should be maintained over all baseband frequencies to be transmitted so that the frequency deviation in the IF filter is fully compressed. With this restriction, the closed-loop noise bandwidth should be as small as possible for a given amount of feedback, in order to minimize the feedback threshold. The filtering should be proportioned so that the IF filter is as wide as possible without allowing the open-loop threshold to predominate. This allows the largest possible frequency deviation in the IF and therefore yields the largest possible baseband signal-to-noise ratio for a fixed system threshold carrier-to-noise ratio. The remaining filtering required to realize the desired closed-loop response function is done at baseband in the feedback path.

The concepts discussed in this Section have been thoroughly tested experimentally. Demodulators were deliberately designed to have the feedback threshold predominate. A typical set of data is shown in Table I. The threshold was defined as the point where the noise impulses first appear (actually, in order to obtain a consistent measuring point, the value where the impulses were occurring at an average rate of one per second was used). The closed-loop noise bandwidths were varied by using different IF filter shapes and/or by placing different filters in the baseband feedback path. The IF filter 3-db bandwidths were 6 to 7 kc in all cases. The mean value is $\phi_{rms} = 1/3.11$. All values fall within a spread of pulse or minus 7 per cent of the mean value, which is within the estimated accuracy of the measurements. Although this data is for a 3-kc baseband system, C. L. Ruthroff (to be published) has obtained similar results for a 1-Mc baseband system.

TABLE I

THE ROOT-MEAN-SQUARE PHASE OF THE VCO WAVE AT THE THRESHOLD OF TWO FEEDBACK DEMODULATORS FOR DIFFERENT AMOUNTS OF FEEDBACK AND DIFFERENT CLOSED-LOOP NOISE BANDWIDTHS

	$F=1+K_rK_f$ (Feedback) DB	B_c (Closed-Loop Noise Bandwidth) KC	ϕ_{rms} Radians
Receiver No. 1	20	172	1/2.89
	20	147	1/2.92
	20	39	1/3.20
	14	68	1/3.20
Receiver No. 2	20	45	1/3.27
	30	135	1/3.20

A typical set of experimental curves in Figs. 3 and 4 illustrate that the IF bandwidth of a feedback demodulator designed using the old theory can indeed be greatly increased without increasing the over-all system threshold. The closed-loop frequency response functions for the two different filter arrangements are shown in Fig. 3. In case A the IF filter was a single pole with a 6-kc, 3-db bandwidth, and the baseband filter in the feedback path was a single pole with a 15-kc 3-db bandwidth. In case B the 3-db IF bandwidth was increased to approximately 30 kc, and the 3-db baseband bandwidth was decreased to approximately 3 kc. Curves of the output baseband noise in a 3-kc bandwidth vs the input carrier-to-noise ratio in an arbitrary but fixed bandwidth are shown in Fig. 4. It can be seen that the system threshold was the same in both cases. The important point is that while the thresholds were equal, the wideband IF system could receive an FM signal with an

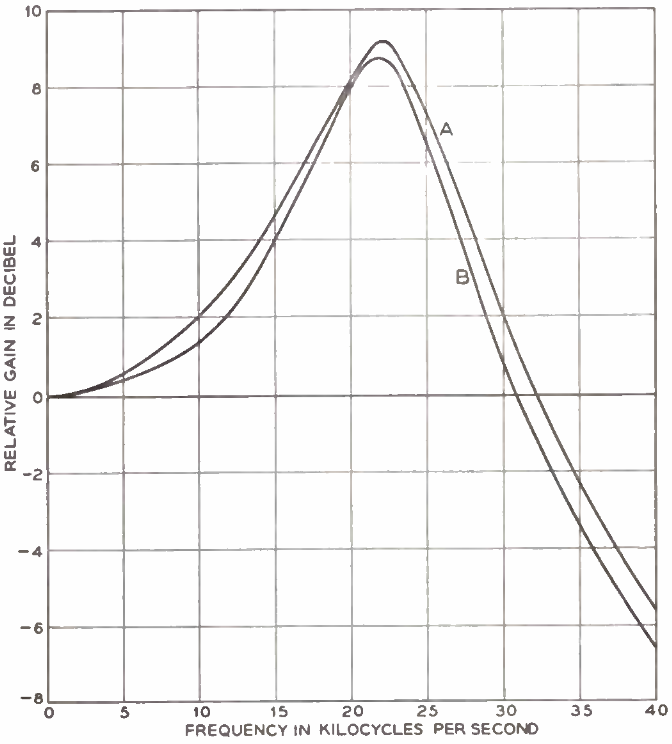

Fig. 3—Measured closed-loop system response. Curve A: 6-kc single-pole IF filter, 15-kc single-pole baseband filter. Curve B: 30-kc single-pole IF filter, 3-kc single-pole baseband filter.

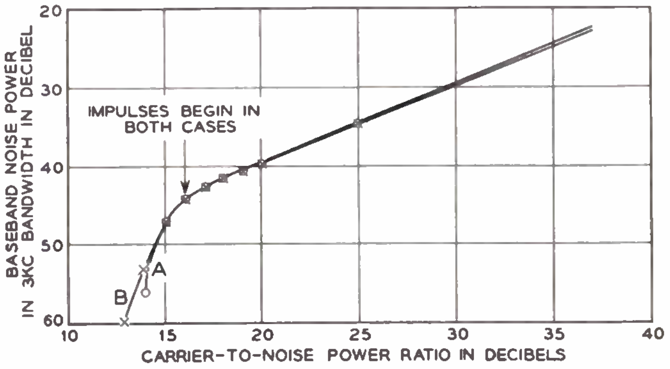

Fig. 4—Plot of the relative baseband noise power in a 3-kc bandwidth vs the input carrier-to-noise power ratio in an arbitrary but fixed bandwidth for the two cases in Fig. 3. The input noise was white for all practical purposes.

index approximately 5 times larger than that of the narrow IF system, yielding an increase in signal-to-noise ratio of 25.[6]

IV. Analysis of the Threshold (Carrier Unmodulated)

A mathematical analysis of the feedback demodulator when it is operating *below* threshold is a difficult and unsolved problem. The determination of where the threshold occurs, however, is a considerably less difficult problem, and it is this problem which will now be considered. We have seen that when the feedback demodulator is above threshold, the baseband noise is derived entirely from the quadrature component of the input noise. This condition must be satisfied in order for the FM improvement formulas to be valid. The threshold will be assumed to occur when a significant amount of additional noise appears in the baseband. A series approach will be used to determine where this additional noise becomes significant. In order to assure the validity of the result, we will then have to check the envelope in front of the frequency detector to make sure that the probability of zero crossings is negligibly small.

Let the carrier and noise at the input to the mixer be given by

$$e_1 = \cos \omega_1 t + N_c(t) \cos \omega_1 t - N_s(t) \sin \omega_1 t,$$

where $N_c(t)$ and $N_s(t)$ are Gaussian random variables in time quadrature with each other, normalized with respect to the carrier magnitude. Let the wave generated by the VCO be given by

$$e_2 = 2 \cos \left[\omega_2 t + \phi(t) + \theta\right].$$

The problem is to find the conditions which must be satisfied so that the phase of the wave entering the frequency detector will be essentially

$$\Psi_1(t) \doteq \overline{N}_s(t) - \bar{\phi}(t),$$

where $\overline{N}_s(t)$ and $\bar{\phi}(t)$ represent the output of the baseband equivalent of the narrow-band IF filter to $N_s(t)$ and $\phi(t)$ respectively. If only the difference frequency terms are retained, the mixer output is

$$\begin{aligned} e_3(t) = e_1 e_2 = {} & \left[\cos \phi(t) + N_c(t) \cos \phi(t) \right. \\ & \left. + N_s(t) \sin \phi(t)\right] \cos (\omega_0 t - \theta) \\ & - \left[-\sin \phi(t) - N_c(t) \sin \phi(t) \right. \\ & \left. + N_s(t) \cos \phi(t)\right] \sin (\omega_0 t - \theta), \end{aligned}$$

where $\omega_0 = \omega_1 - \omega_2$. When this wave is passed through the IF filter and $\sin \phi(t)$ and $\cos \phi(t)$ are expressed in series form, then

$$\begin{aligned} e_3(t) = {} & \left\{ \sum_{m=0}^{\infty} (-1)^m \frac{\overline{\phi^{2m}}}{(2m)!} + \sum_{m=0}^{\infty} (-1)^m \frac{\overline{N_c \phi^{2m}}}{(2m)!} \right. \\ & \left. + \sum_{m=0}^{\infty} (-1)^m \frac{\overline{N_s \phi^{2m+1}}}{(2m+1)!} \right\} \cos (\omega_0 t - \theta) \\ & - \left\{ - \sum_{m=0}^{\infty} (-1)^m \frac{\overline{\phi^{2m+1}}}{(2m+1)!} - \sum_{m=0}^{\infty} (-1)^m \frac{\overline{N_c \phi^{2m+1}}}{(2m+1)!} \right. \\ & \left. + \sum_{m=0}^{\infty} (-1)^m \frac{\overline{N_s \phi^{2m}}}{(2m)!} \right\} \sin (\omega_0 t - \theta). \end{aligned}$$

The envelope and phase of a wave in the form $A(t) \cos \omega_0 t - B(t) \sin \omega_0 t$ are given respectively by

$$E_1(t) = \sqrt{A^2(t) + B^2(t)}$$

and

$$\Psi_1(t) = \tan^{-1} \frac{B(t)}{A(t)} .$$

Neglecting all third-order and higher terms, we have

$$\begin{aligned} \Psi_1(t) \doteq {} & \overline{N}_s(t) - \bar{\phi}(t) - \overline{N_c(t)\phi(t)} \\ & - \overline{N}_c(t)\overline{N}_s(t) + \overline{N}_c(t)\bar{\phi}(t) \end{aligned}$$

and

$$\begin{aligned} E_1(t) \doteq {} & 1 + \overline{N}_c(t) + \overline{N_s(t)\phi(t)} - \tfrac{1}{2}\overline{\phi^2(t)} + \tfrac{1}{2}\overline{N}_c^{\,2}(t) \\ & + \tfrac{1}{2}\overline{N}_s^{\,2}(t) + \tfrac{1}{2}\bar{\phi}^2(t) - \overline{N}_s(t)\bar{\phi}(t). \end{aligned} \tag{2}$$

Pay particular attention to the difference between terms which were multiplied and then filtered, such as $\overline{N_s(t)\phi(t)}$, and terms which were filtered and then multiplied, such as $\overline{N}_s(t)\bar{\phi}(t)$. The relative importance of these second-order terms will now be discussed. Assume that there is no feedback and that the resulting open-loop system is above threshold. $\phi(t)$ is then zero, and only the first and fourth terms remain in the expression for the phase $\Psi_1(t)$. The first term represents the "above threshold noise" or quadrature noise. The fourth term is negligible because the system is assumed above threshold. Now assume that feedback is applied, and that the closed-loop system is also above threshold. If there is a significant amount of feedback, $\bar{\phi}(t)$ approaches $\overline{N}_s(t)$. The first two terms represent the "above-threshold" noise of the feedback demodulator. The fifth term, which is introduced by feedback, is almost equal to and tends to cancel the fourth term, which is present with or without feedback. They may both be neglected. The third term, which is also introduced by feedback, is considerably larger than the other second-order terms. It represents phase noise

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[6] The system parameters should not be taken as optimum in any sense. This design served only to verify theory and nothing more.

which adds to the above-threshold noise, *i.e.*, the first two terms. The threshold is assumed to occur when this additional noise becomes significant. The third term is a distortion term since it is a function of the "signal" $\phi(t)$. In the analysis of feedback systems having distortion, it is often possible to represent the complete system as a linear one in which the distortion components are introduced by generators [9]. This approach is valid as long as the fed-back distortion components are small; *i.e.*, all significant distortion must be caused by the signal. An important aspect of such an analysis is that one can tell to a fair degree of approximation where the results of the linearized model cease to be accurate. This general approach proves to be a useful one in the present problem. Let $\phi_s(t)$ be the VCO phase contributed by the quadrature noise $N_s(t)$, and let $\phi_d(t)$ be the phase contributed by the distortion component. Then if $\phi_s(t) \gg \phi_d(t)$, the significant distortion would be caused by $\phi_s(t)$, and the linearized baseband model shown in Fig. 5 is valid. The quadrature noise and distortion are introduced into the system in parallel and are treated identically by feedback. Consequently $\phi_s(t) \gg \phi_d(t)$ if the spectral density of the distortion input $\phi_s(t) N_c(t)$ is much smaller than the spectral density of the quadrature term $N_s(t)$. In Appendix A it is shown that for large feedback the ratio of the spectral density of $\phi_s(t) N_c(t)$ to the density of $N_s(t)$ is equal to the mean-square value of $\phi_s(t)$, *i.e.*, $\phi_{\mathrm{rms}}{}^2$. Thus, the linear model in Fig. 5 is valid as long as $\phi_{\mathrm{rms}}{}^2 \ll 1$. In this region of operation the phase term in (2) becomes approximately

$$\Psi_1(t) \doteq \left\{ \overline{N}_s(t) - \bar{\phi}_s(t) \right\} - \left\{ \overline{N_c(t)\phi_s(t)} - \bar{\phi}_d(t) \right\},$$

where the second bracketed term is much smaller than the first. This second bracketed term represents the excess noise which is not derived from the quadrature input noise. One would expect intuitively that $\phi_{\mathrm{rms}}{}^2 \doteq 1/10$ would be close to the upper limit of validity for the model of Fig. 5, for much beyond that the nonlinear action on the feedback distortion would produce sufficient additional distortion to result in a cumulative situation. This result agrees quite well with the measured threshold value of $\phi_{\mathrm{rms}} = 1/3.11$.

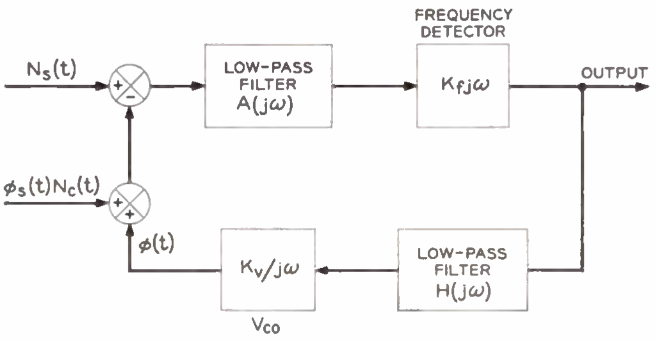

Fig. 5—Block diagram for representing the effect of excess phase noise or "noise distortion."

In order for the results in the above paragraph to be correct, it is necessary that the number of zero crossings of the envelope be negligibly small for $\phi_{\mathrm{rms}}{}^2 \leq 1/10$. Calculation of the actual probability of zero crossings is prohibitively difficult. In lieu of this calculation it will be shown that the root-mean-square fluctuation of the envelope is negligibly small compared to its mean value. The first term in the envelope expression of (2) is the normalized carrier envelope. The second, fifth, and sixth terms represent the envelope noise when the loop is open. Since the system is above threshold on an open-loop basis these terms are sufficiently small. The last three terms tend to cancel each other and can be neglected if $\bar{\phi}(t) \doteq \overline{N}_s(t)$. Demonstration that the third and fourth terms are small compared to the normalized carrier envelope, *i.e.*, unity, requires the specification of a particular closed-loop transfer function. We shall select the two-pole low-pass function, having a damping ratio less than unity, because of its practical importance. It is felt that the results are typical of most other transfer functions of interest. From (2) it can be seen that the third and fourth terms are formed by multiplying two variables together and then passing the result through the low-pass equivalent of the IF filter. In Appendix B it is shown that the total power lying within a rectangular band of frequencies equal to *twice* the noise bandwidth of the closed-loop response is $P = 11/4\phi_{\mathrm{rms}}{}^4$. Since the power passed by the IF filter is smaller still,[6] the envelope noise of the composite wave at the input to the frequency detector is small compared to unity for $\phi_{\mathrm{rms}}{}^2 \leq 1/10$.

In summary, the mean-square phase deviation of the VCO wave $\phi_{\mathrm{rms}}{}^2$ caused by noise must be small compared to unity in order for the noise appearing in the baseband to be primarily derived from the quadrature input noise, *i.e.*, predictable from the FM improvement formula. This, then, is also the condition which must be satisfied for a system to be above the feedback threshold.

V. MODULATION

In Section IV the carrier was assumed to be unmodulated. Here it is shown that the system *stability*, and hence the closed-loop bandwidth and feedback threshold, are all functions of *modulation*. It is then shown that the dependence may be virtually eliminated by proper design.

Assume that the carrier frequency is varying slowly so that it may be represented quasi-statically. Noise can then be separated into components in phase and in quadrature with the carrier. The quadrature component is the angle noise, and it is the response of the system to this noise that is of interest. The nature of this re-

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[6] Notice that, for a fixed closed-loop noise bandwidth, increasing the resonant peak of the closed-loop response function allows more envelope noise to pass through the IF filter.

sponse may be determined from the small index response of the system. It is desired to find a transfer function which will relate the signal component of the instantaneous phase at the output of a mistuned high-Q band-pass filter to the phase of the small index modulated carrier applied to its input. The mistuning, of course, represents the deviation of the quasi-statically varying carrier frequency caused by signal modulation. Distortion will not be considered since it does not affect stability. In Appendix C[7] it is shown that the baseband analog of the mistuned high-Q band-pass filter is

$$Y(j\omega) = \frac{1}{2R(\omega_d)}\left[Y_x(j\omega) + Y_x^*(-j\omega)\right],$$

where

$$Y_x(j\omega) = Y_L[j(\omega + \omega_d)]\epsilon^{-j\theta(\omega_d)},$$

$$Y_L[j(\omega + \omega_d)] = R(\omega + \omega_d)\epsilon^{j\theta(\omega+\omega_d)},$$

ω_d = difference between the carrier frequency and the filter center frequency,

$Y_L(j\omega)$ = conventional low-pass equivalent of the band-pass circuit.

In Figs. 6 and 7 are shown magnitude and phase plots for a single-pole and Bode-type filter [10]. The conventional low-pass equivalent is given by the $\omega_d = 0$ curve. Modulation manifests itself as a variation in ω_d, and so its effect on the phase margin of the system is very apparent. A slow roll-off filter such as the single-pole filter is obviously required in the IF. The single-pole filter introduces only a slight excess phase shift for carrier frequency deviations, *i.e.*, variations in ω_d, within the 3-db bandwidth. Consequently, only slight variations occur in the closed-loop bandwidth and therefore in the feedback threshold.

This instability can have real nuisance value because of the likelihood that it will be overlooked in routine tests on experimental systems. The reason for this is that the system is unstable only on the the peaks of the modulation. Further, the oscillations are generally at a frequency which is high compared to the baseband bandwidth and consequently do not appear *after* the baseband filter which usually follows FM demodulators to remove noise lying outside of the baseband bandwidth. The effect which is noticed is a degradation of the threshold as a function of modulation. The carrier frequency in the IF breaks into oscillation at the modulation peaks. The IF filter behaves like a slope circuit and produces envelope modulation on the carrier, making it much easier for noise peaks to exceed the carrier envelope in front of the frequency detector and produce the threshold.

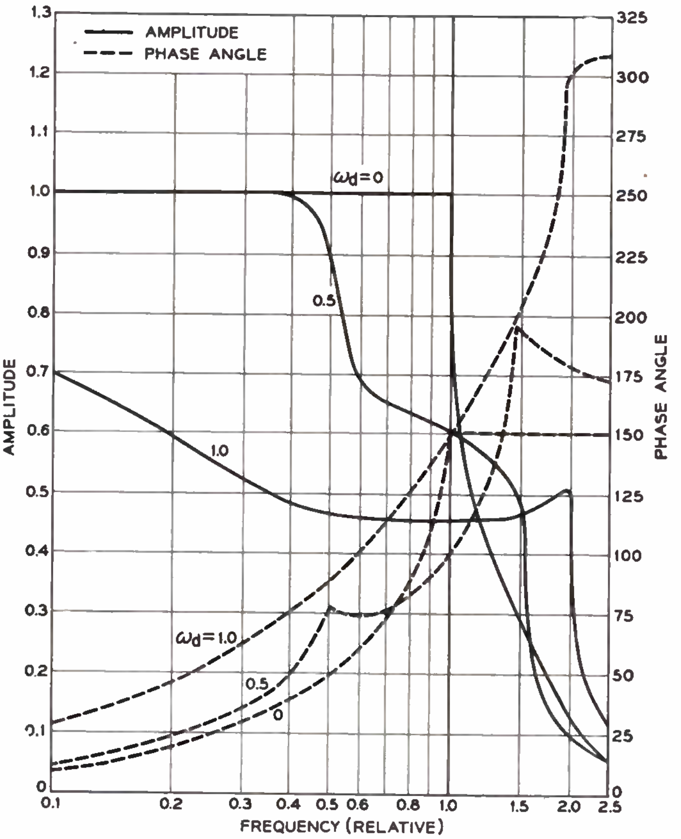

Fig. 6—Amplitude and phase of the transfer function relating the signal component of the instantaneous phase at the output of a Bode-type high-Q band-pass filter to the phase of the small index modulated carrier applied to its input. The carrier is mistuned by ω_d.

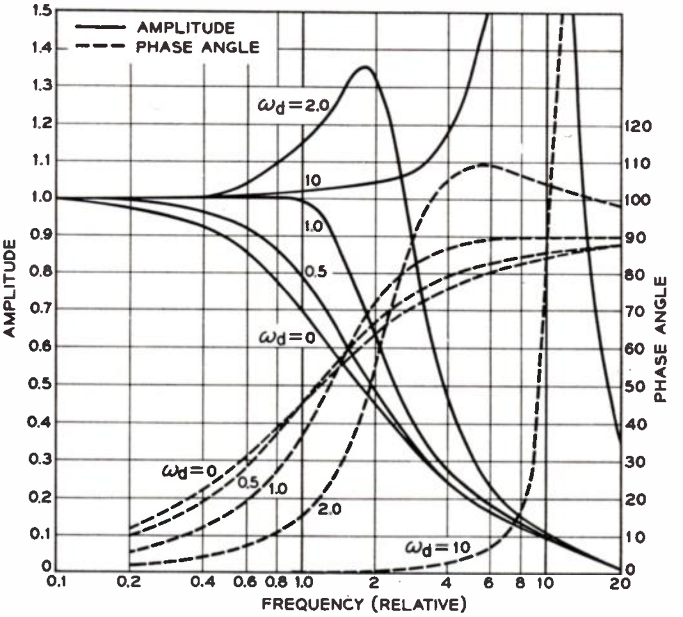

Fig. 7—Amplitude and phase of the transfer function relating the signal component of the instantaneous phase at the output of a single-pole high-Q band-pass filter to the phase of the small index modulated carrier applied to its input. The carrier is mistuned by ω_d.

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[7] The equation developed can also be obtained from a much more general analysis of distortion presented in an unpublished memorandum in 1954 by S. Doba.

The optimum IF filter has not yet been determined; however, it is felt that a single-pole filter will be close. Additional filtering is done at baseband in order to accomplish the following objectives:

1) Full feedback should be maintained over all baseband frequencies to be transmitted, so that the frequency deviation in the IF is fully compressed.
2) The closed-loop noise bandwidth should be as small as possible for a given amount of feedback in order to keep the feedback threshold at a minimum.
3) The IF filter should be as wide as possible without allowing the open-loop threshold to predominate. This allows maximum frequency deviation in the IF and thus maximum baseband signal-to-noise ratio.

VI. Sample Design

The material presented above will now be illustrated by means of an example. This example is not offered as THE way to design a system; it is meant only to consolidate thinking. In particular, we shall use the experimental feedback threshold value given in Table I. In practice the "operating threshold" would depend upon the intended use, *i.e.*, television, data, multiplex telephony, etc.

Assume the following data:

$f_b = 3$ Mc = baseband bandwidth,
$\Delta f = \mp 60$ Mc = peak frequency deviation.

The design will use a Bode-type [10] open-loop characteristic with a phase margin of 50°. It can be seen from Fig. 8 that this phase margin yields a closed-loop noise bandwidth very close to the minimum value for values of feedback less than 30 db. The characteristic will be realized in two parts: A single-pole filter in the IF and the rest at baseband. Now we must determine the required amount of feedback. Fig. 8 illustrates graphically that we must use as little as possible in order to keep the closed-loop bandwidth minimized. However, we must use enough to compress the deviation of the IF wave sufficiently for it to pass through the single-pole IF filter. The IF filter must be narrow enough to prevent the open-loop threshold from being dominant. Thus, we must know the feedback threshold, which can not be calculated until the amount of feedback is determined. We see that we have gone in a full circle. The result is that we consider only the baseband bandwidth as being known, estimate a value of feedback, and then solve for the frequency deviation. If we do not obtain the required deviation, we make another estimate and try again.

Let us choose 20 db of feedback for our first estimate, *i.e.*, $F = 10$. The closed-loop bandwidth is, from Fig. 8, $B_c \approx 11.6\ f_b = 34.8$ Mc. From (1), the carrier-to-noise ratio at the input to the mixer in this bandwidth at the feedback threshold is

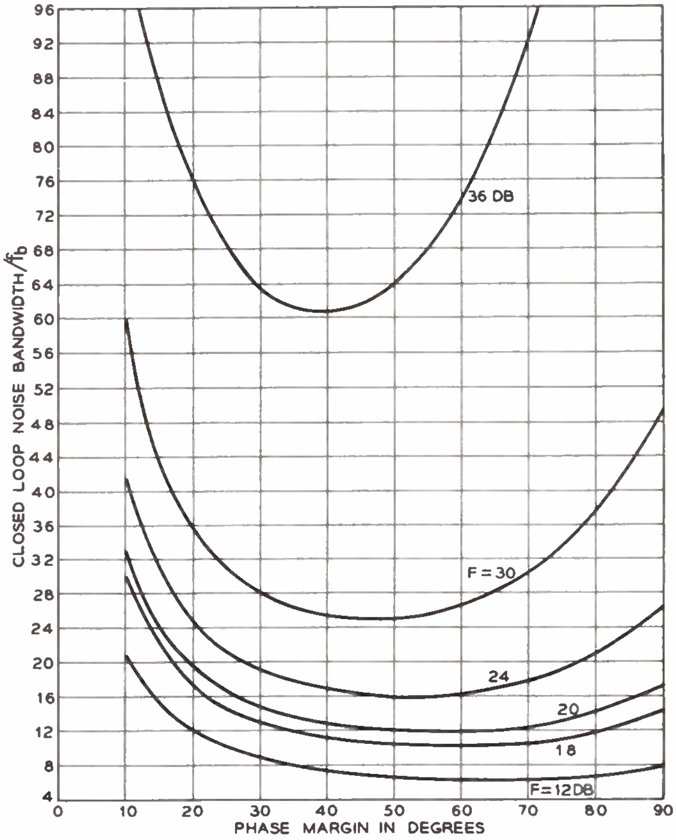

Fig. 8—The closed-loop noise bandwidth of systems having a Bode-type open-loop characteristic for different amounts of feedback and phase margin.

$$\rho_T = 4.8\left[\frac{9}{10}\right]^2 = 3.92 \quad \text{or} \quad 5.94 \text{ db}.$$

The single-pole filter in the IF must have a noise bandwidth small enough so that the open-loop threshold does not predominate. The threshold of a frequency detector depends upon the ratio of the input noise bandwidth to baseband bandwidth. See Fig. 9.[8] Therefore we are forced to guess and check our guess.

Guess at a value of 6 for the ratio of the IF filter noise bandwidth to baseband bandwidth. From Fig. 9 it is seen that for $B/f_b = 6$ the signal-to-noise ratio is proportional to the carrier-to-noise ratio for all carrier-to-noise ratios greater than 8.5 db. 8.5 db will be used as the open-loop threshold. Now for the check. The noise density which yields a carrier-to-noise ratio of 5.94 db in a 34.8-Mc bandwidth (closed-loop noise

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[8] These curves were derived for an ideal frequency detector with a carrier and a narrow band of Gaussian noise at its input. Since in the feedback demodulator the quadrature noise has been reduced by feedback, the curves do not apply exactly. However, experience has shown that they may be used to estimate the frequency detector threshold for systems typified by those in Table I.

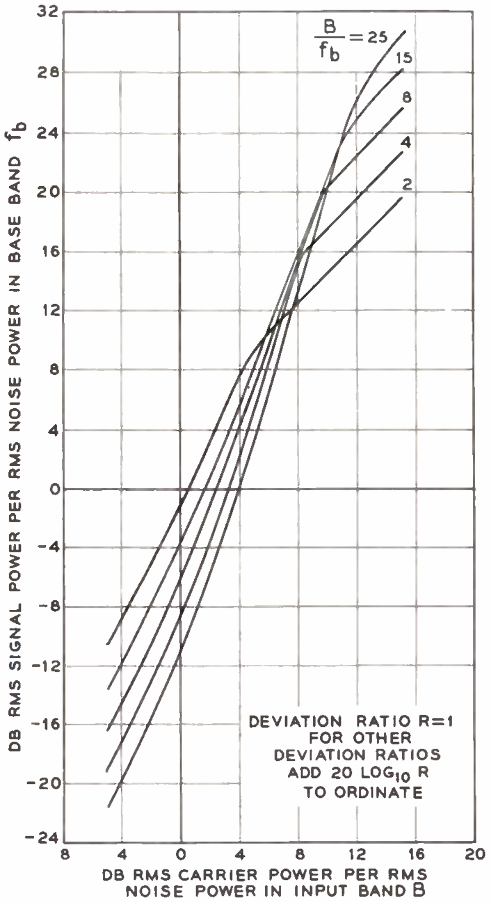

Fig. 9—Threshold curves. (Presented by F. J. Skinner in an unpublished memorandum in 1954; derived from the results of Rice [6]. The threshold phenomenon generates a more or less flat power spectrum which is superimposed on the parabolic power spectrum (triangular voltage spectrum) at baseband. Thus, the threshold effects are noticed sooner in the lower frequencies, *i.e.*, for large B/f_b ratios. For similar curves see Stumpers [11].

bandwidth) also yields a carrier-to-noise ratio of 8.5 db in a 19.35-Mc bandwidth (IF noise bandwidth). Using 19.35 Mc for the IF bandwidth means that B/f_b is slightly greater than 6 but not enough to matter. The 3-db bandwidth of a single-pole filter with a noise bandwidth of 19.35 Mc is $B = 19.35/1.57 = 12.3$ Mc. The compressed wave could have an index[9] of 2 in this filter and not have sufficient distortion to be detectable on an oscilloscope (this distortion is reduced by feedback) [1]. The peak frequency deviation in front of the mixer would be

$$\Delta f = 10 \times 2 \times 3 = 60 \text{ Mc}.$$

From Fig. 9 a conventional demodulator receiving the same transmitted wave would have a threshold greater than 12 db in the required 120-Mc rf bandwidth. Thus, the feedback demodulator has a threshold which is better by at least

$$12 - 5.94 + 10 \log_{10} \frac{120}{34.8} \doteq 11.5 \text{ db}.$$

VII. Carrier Insertion and the Phase-Locked Loop

In amplitude modulation systems a locally generated sine wave, whose phase and frequency are the same as the carrier, may be inserted to extend the threshold indefinitely, at least theoretically [12]. It has been proposed [4] that such a scheme be incorporated with the feedback demodulator. If the large index transmitted wave were phase modulated instead of frequency modulated, feedback could be used to reduce the deviation in the IF until only the first-order sidebands were important. A locally generated sine wave of the proper phase and frequency could then be inserted to eliminate the threshold of the phase detector. It will be shown in the next few paragraphs that the thresholds of the phase-locked-loop and the feedback demodulator with carrier insertion are equal to or poorer than that of a feedback demodulator without carrier insertion. The reason for this stems from the fact that in the first two cases the instantaneous phase deviation of the wave behind the mixer (multiplier) must be small compared to unity in order to prevent severe distortion. For large transmitter modulation indexes this requires more feedback than would be required by a straightforward feedback demodulator, and consequently results in a larger closed-loop noise bandwidth. The larger closed-loop noise bandwidth yields a poorer feedback threshold. First let us consider the carrier insertion scheme. Let the carrier be inserted directly after the mixer. The combined output is then

$$\begin{aligned} e_3(t) = \{&C_2 + \cos\phi(t) + N_c(t)\cos\phi(t) \\ &+ N_s(t)\sin\phi(t)\}\cos(\omega_0 t - \theta) \\ - \{&-\sin\phi(t) - N_c(t)\sin\phi(t) \\ &+ N_s(t)\cos\phi(t)\}\sin(\omega_0 t - \theta), \end{aligned} \quad (3)$$

where $C_2 \cos(\omega_0 t - \theta)$ is the inserted carrier, normalized by the incoming carrier magnitude.

The envelope and phase may be found exactly as in the development of (2). They are

$$\begin{aligned} E_1(t) = 1 &+ \frac{1}{1 + C_2}\left\{\overline{N_c(t)} + \overline{N_s(t)\phi(t)} - \tfrac{1}{2}\overline{\phi^2(t)}\right\} \\ &+ \frac{1}{(1 + C_2)^2}\left\{\tfrac{1}{2}\overline{N_s^2}(t) + \tfrac{1}{2}\overline{N_c^2}(t) + \tfrac{1}{2}\overline{\phi^2}(t) - \overline{N_s}(t)\bar{\phi}(t)\right\}, \end{aligned}$$

$$\begin{aligned} \Psi_1(t) &= \frac{1}{1 + C_2}\left\{\overline{N_s}(t) - \bar{\phi}(t) - \overline{N_c(t)\phi(t)}\right\} \\ &- \frac{1}{(1 + C_2)^2}\left\{\overline{N_c}(t)\overline{N_s}(t) - \overline{N_c}(t)\bar{\phi}(t)\right\}. \end{aligned}$$

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[9] In practice one could not realize a closed-loop noise bandwidth as small as this example suggests. We have selected a rather idealized open-loop characteristic and have completely neglected excess linear phase shift caused by parasitics. It is for closed-loop noise bandwidths much greater than the baseband bandwidth where the ability to widen the IF bandwidth pays dividends.

The terms proportional to $1/(1+C_2)^2$ tend to cancel just as did the corresponding terms in (2).

There are two points to notice:

1) The inserted carrier reduces the percentage envelope modulation by the factor $1+C_2$, decreasing system susceptibility to envelope noise.

2) The above-threshold phase noise $\overline{N}_s(t) - \bar{\phi}(t)$ and the "threshold-producing" phase noise $\overline{N_c(t)\phi(t)}$ are reduced by the same factor $1+C_2$, as modulation would be if it were present. This reduction produces the same effect on the baseband as a decrease in the gain constant of the phase detector. Hence, the effect is simply to reduce the amount of feedback. The feedback threshold of this demodulator is not affected by carrier insertion.

Now consider the phase-locked loop. In the phase-locked loop the IF frequency ω_0 is zero and $\theta = \pi/2$. The output of the mixer (multiplier) is a baseband signal proportional to the quadrature component of (3). Therefore, there is no need for a separate phase detector or carrier insertion. Notice that the quadrature term is $Q(t) \doteq N_s(t) - \phi(t) - N_c(t)\phi(t)$. Again the threshold producing noise is present and undiminished, yielding the feedback threshold.

In summary, both the phase-locked loop and the feedback demodulator with carrier insertion require the instantaneous phase deviation of the wave behind the mixer (multiplier) to be small compared to unity. For large transmitted modulation indexes this requires more feedback on the highest baseband frequency than would be required by a straightforward feedback demodulator, and consequently a larger closed-loop noise bandwidth is obtained. The larger closed-loop noise bandwidth yields a poorer feedback threshold.

VIII. CONCLUSIONS

The over-all threshold of the frequency feedback demodulator was determined. The results show that two thresholds can occur. The first occurs in the usual manner at the frequency detector; the second occurs when the root-mean-square phase of the VCO wave (caused by noise) is no longer small compared to unity. Experimentally it was found that the second threshold occurs when $\phi_{\rm rms} = 1/3.11 \pm 7$ per cent radians. This was related quantitatively to the carrier-to-noise ratio in the closed-loop noise bandwidth. It was pointed out that for maximum baseband signal-to-noise ratio the feedback threshold and open-loop threshold should occur in the same neighborhood. It was also shown that sharp cutoff filters in the IF can cause a serious loss in threshold under modulated conditions, and for this reason slow roll-off filters such as the single-pole filter should be used. Realizing these facts can improve the baseband signal-to-noise ratio by one or two orders of magnitude, for a given carrier power and noise density, over that which would be obtained in systems designed be previous theories [3], [4]. It was shown that the phase-locked loop and the feedback demodulator with carrier insertion have a poorer threshold than the standard feedback demodulator for large index modulation.

The author would like to point out several problems that need further study. The "operational" threshold, which determines the limit of usefulness of a demodulator, varies according to the service for which it is used, *i.e.*, multiplex telephony, television, etc. This threshold needs to be determined subjectively for each of these services. It has been pointed out in this paper that feedback demodulators should be designed to have the feedback and open-loop thresholds occur in the same neighborhood. Experimental results for the various services are also needed here in order to be more specific. Intermodulation distortion produced by the IF filter should be studied to determine the maximum allowable frequency deviation. Last but not least, it would be useful if a general analysis of the behavior of the feedback demodulator below threshold were available. However, the fact that the threshold of an ideal frequency detector, in the absence of feedback, is not understood to a satisfactory degree gives some insight into the difficulties.

APPENDIX A

It is desired to show that the ratio of the spectral power density of the term $\phi_s(t)N_c(t)$ to the density of $N_s(t)$ is equal to the mean-square value of $\phi_s(t)$ for large feedback. $N_c(t)$ and $N_s(t)$ are the in-phase and quadrature components of the input noise, normalized by the carrier magnitude. $\phi_s(t)$ is the component of the phase of the VCO wave derived from the quadrature component of the input noise.

The spectral density of the term $\phi_s(t)N_c(t)$ is found by determining the covariance and then taking the Fourier transform. The covariance is

$$R_1(\tau) = E[\phi_s(t_1)N_c(t_1)\phi_s(t_2)N_c(t_2)],$$

where $E(\ )$ is the statistical average or expectation of $(\ )$. Now $\phi_s(t)$ is derived from $N_s(t)$ by linear filtering; it follows that $\phi_s(t)$ and $N_c(t)$ are independent processes as long as the RF and IF filters are symmetrical [13]. Thus

$$R_1(\tau) = R_{\phi_s}(\tau)R_{N_c}(\tau),$$

where R_{ϕ_s} and R_{N_c} are the covariances of ϕ_s and N_c, respectively. The spectral density can now be found by convolving the individual densities.

$$S_1(f) = \int_{-\infty}^{+\infty} S_{N_c}(f_1 - f)S_{\phi_s}(f_1)df_1.$$

We are interested in the power density only within the bandwidth of the closed-loop response function. If the feedback is equal to or greater than 15 db or so, the bandwidth of the RF band of noise is equal to or greater than twice the 3-db bandwidth of the closed-loop response function. Under these conditions

$$S_{N_c}(f_1 - f) \doteq S_{N_c}(f)$$

for f and f_1 within the desired range. Thus

$$\frac{S_1(f)}{S_{N_s}(f)} \doteq \int_{-\infty}^{+\infty} S_{\phi_s}(f_1)df_1 \doteq \phi_{\text{rms}}{}^2,$$

since

$$S_{N_c}(f) = S_{N_s}(f).$$

APPENDIX B

It is desired to determine an upper bound to the total power passed by an ideal rectangular filter, whose bandwidth is equal to twice the closed-loop noise bandwidth, when the input is the first two second-order terms in the envelope expression of (2), *i.e.*,

$$E_2(t) = N_s(t)\phi(t) - \tfrac{1}{2}\phi^2(t).$$

A two-pole low-pass transfer function will be chosen as representative of the closed-loop transfer function. Let the impulse and transfer functions be, respectively,

$$h(t) = \frac{\omega_n}{\sqrt{1 - \zeta^2}} \epsilon^{-\omega_n \zeta t} \sin\{\omega_n \sqrt{1 - \zeta^2} t\}$$

and

$$H(j\omega) = \frac{\omega_n{}^2}{\omega_n{}^2 - \omega^2 + 2j\omega\omega_n\zeta},$$

where $\zeta < 1$.

The covariance can be expanded in the following form [14]:

$$\begin{aligned} R_{E_2}(\tau) = {} & R_{N\phi}{}^2(0) + R_{NN}(\tau)R_{\phi\phi}(\tau) + R_{N\phi}(\tau)R_{\phi N}(\tau) \\ & - R_{N\phi}(0)\, R_{\phi\phi}(0) - R_{N\phi}(\tau)R_{\phi\phi}(\tau) - R_{\phi N}(\tau) \\ & R_{\phi\phi}(\tau) + \tfrac{1}{4}R_{\phi\phi}{}^2(0) + \tfrac{1}{2}R_{\phi\phi}{}^2(\tau), \end{aligned}$$

where $R_{N\phi}(\tau) \equiv E[N_s(t_1)\phi(t_2)]$, etc. Now the individual terms must be calculated. The term $R_{\phi\phi}(\tau)$ may be found in Middleton [15]:

$$\begin{aligned} R_{\phi\phi}(\tau) = {} & \frac{\sigma_s{}^2\omega_n}{4\zeta} \epsilon^{-\zeta\omega_n|\tau|} \cos\{\omega_n\sqrt{1 - \zeta^2}\tau\} + \\ & + \frac{\zeta}{\sqrt{1 - \zeta^2}} \sin\{\omega_n\sqrt{1 - \zeta^2}\,|\tau|\}, \end{aligned}$$

where $\sigma_s{}^2$ is the spectral density of $N_s(t)$. The term for $R_{NN}(\tau)$ is

$$R_{NN}(\tau) = \sigma_s{}^2\delta(\tau),$$

where $\delta(\tau)$ is the "delta" function. The term $R_{\phi N}(\tau)$ is simply the product of the impulse function $h(\tau)$ and $\sigma_s{}^2$.

$$R_{\phi N}(\tau) \left| \begin{array}{ll} = \dfrac{\sigma_s{}^2\omega_n}{\sqrt{1 - \zeta^2}} \epsilon^{-\omega_n\zeta\tau} \sin \omega_n \sqrt{1 - \zeta^2}\,\tau, & \tau \geq 0, \\ 0, \tau \leq 0. & \end{array} \right.$$

The spectral density corresponding to the inverse Fourier transform of $R_{E_1}(\tau)$ contains a dc term, a term of uniform density and a number of terms whose energy is concentrated within a bandwidth twice that of the function $H(j\omega)$. The total power which would be passed by a rectangular filter whose bandwidth is equal to twice the closed-loop noise bandwidth ($B_c = \omega_n/4\zeta$) is smaller than

$$P = \frac{11}{4}\left(\frac{\sigma^4 s^{\omega^2} n}{16\zeta^2}\right) = \frac{11}{4}\phi_{\text{rms}}{}^4.$$

P is calculated from $R_{E_1}(\tau)$ by letting $\tau = 0$ in all terms except the one containing $R_{NN}(\tau)$ and adding them. The term $R_{NN}(\tau)$ yields a uniform spectral density. Only the power within a bandwidth equal to twice the closed-loop noise bandwidth was included.

APPENDIX C

MISTUNED ANGLE-MODULATED CARRIER

It is desired to find a transfer function which will relate the signal component of the instantaneous phase at the output of a mistuned high-Q band-pass filter to the phase of the small index modulated carrier applied to its input. Distortion will not be considered since it is only the signal transfer function which is important to stability. Let the small index signal be

$$\begin{aligned} e_1(t) = \cos[\omega_0 t + \phi(t)] &= R_e\left(\epsilon^{j\omega_0 t} \sum_0^\infty \frac{[i\phi(t)]^n}{n!}\right) \\ &\doteq R_e([1 + j\phi(t)]\epsilon^{j\omega_0 t}). \end{aligned}$$

The result is a "baseband" signal which has been translated in frequency by ω_0. If $Y_L(j\omega)$ is the transfer function of the conventional low-pass equivalent of the filter, then $Y_L[j(\omega + \omega_d)]$ is the transfer function seen by the baseband signal because of the carrier being mistuned by ω_d. The output signal will, in general, be both envelope and phase modulated. Only the signal component of the phase modulation is desired. Write

$$\begin{aligned} Y_L[j(\omega + \omega_d)] &\equiv R(\omega + \omega_d)\epsilon^{j\theta(\omega + \omega_d)} \\ &= \epsilon^{j\theta(\omega_d)} Y_x(j\omega), \end{aligned}$$

where

$$Y_x(j\omega) \equiv R(\omega + \omega_d)\epsilon^{j\theta(\omega + \omega_d) - j\theta(\omega_d)}.$$

$Y_x(j\omega)$ is not the transfer function of a physically realizable network, and consequently its impulse function is complex, *i.e.*,

$$h(\tau) = h_1(\tau) + jh_2(\tau),$$

where $h_1(\tau)$ and $h_2(\tau)$ are real. The filter output may be found by convolution:

$$\begin{aligned} e_0(t) &= \epsilon^{j\theta(\omega_d)} \int_{-\infty}^{+\infty} [h_1(\tau) + jh_2(\tau)][1 + j\phi(t - \tau)d\tau \\ &= \epsilon^{j\theta(\omega_d)} R(\omega_d) + \int_{-\infty}^{+\infty} [h_1(\tau) + jh_2(\tau)][j\phi(t - \tau)]d\tau. \end{aligned}$$

The signal component of phase is

$$\Psi(t) = \operatorname{Im} \ln e_0(t)$$

$$= \theta(\omega_d) + \frac{1}{R(\omega_d)} \int_{-\infty}^{+\infty} h_1(\tau)\phi(t-\tau)d\tau$$

assuming

$$\left| \int_{-\infty}^{+\infty} \phi(t-\tau)[h_1(\tau) + jh_2(\tau)]d\tau \right| \ll R(\omega_d),$$

i.e., if the distortion is not too severe. $\theta(\omega_d)$ is a constant and can be ignored. The desired transfer function has an impulse function of $h_1(\tau)/R(\omega_d)$. $h_1(\tau)$ is the real part of $h(\tau)$, and consequently its transfer function consists of the even real part and odd imaginary part of the minimum phase function $Y_z(j\omega)$. The desired transfer function is then

$$Y(j\omega) = \frac{1}{2R(\omega_d)} [Y_z(j\omega) + Y_z^*(-j\omega)].$$

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REFERENCES

[1] J. G. Chaffee, "The application of negative feedback to frequency-modulation systems," *Bell Sys. Tech. J.*, vol. 18, pp. 403–437; July, 1939.

[2] J. R. Carson, "Frequency modulation—theory of the feedback receiving circuit," *Bell Sys. Tech. J.*, vol. 18, pp. 395–403; July, 1939.

[3] M. O. Felix and A. J. Buxton, "The performance of FM scatter systems using frequency compression," *Proc. Natl. Electronics Conf.*, vol. 14, pp. 1029–1043; 1958.

[4] M. Morita and S. Ito, "High sensitivity receiving system for frequency modulated waves," 1960 IRE INTERNATIONAL CONVENTION RECORD, pt. 5, pp. 228–237.

[5] C. L. Ruthroff, "Project Echo: FM demodulators with negative feedback," *Bell Sys. Tech. J.*, vol. XL, pp. 1149–1157; July, 1961.

[6] S. O. Rice, "Properties of a sine wave plus random noise," *Bell Sys. Tech. J.*, vol. 27, pp. 109–157; January, 1948.

[7] J. Dugunji, "Envelopes and pre-envelopes of real waveforms," IRE TRANS. ON INFORMATION THEORY, vol. IT-4, pp. 53–57; March, 1958.

[8] H. S. Black, "Modulation Theory," D. Van Nostrand Co., Inc., New York, N. Y., pp. 218–234; 1953.

[9] J. C. West, J. L. Douce and B. G. Leary, "Frequency Spectrum Distortion of Random Signals in Non-linear Feedback Systems," IEE Monograph No. 419M; November, 1960.

[10] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y., pp. 451–529; 1945.

[11] F. L. H. M. Stumpers, "Theory of frequency modulation noise," PROC. IRE, vol. 36, pp. 1081–1092; September, 1948.

[12] D. B. Harris, "Selective modulation," PROC. IRE, vol. 33, pp. 565–572; June, 1945.

[13] D. Middleton, "Introduction to Statistical Communication Theory," McGraw-Hill Book Co., Inc., New York, N. Y., p. 399; 1960.

[14] *Ibid.*, p. 343.

[15] *Ibid.*, p. 168.

Theory and Design of Wide-Band Parametric Converters*

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Summary—This paper treats the parametric converter from the circuit theory point of view. A linear time-varying capacitance is imbedded in an arbitrary linear passive and time-invariant network. First, we review the steady-state analysis of such a circuit. The formula for transducer power gain is next derived. Based on this formula, the maximum gain-bandwidth product is studied using Bode's relationship.

We prove that for the noninverting type parametric converters, the optimum transducer power gain is that of Manley and Rowe, *i.e.*, ω'/ω. The maximum fractional bandwidth is given by $C_1/C_0(\omega_0'/\omega_0)^{1/2}$. Simple coupling circuits which approach the optimum situations are presented along with typical design examples.

I. INTRODUCTION

EXISTING work on parametric amplifiers deals mostly with the physical aspects of the device, the analysis and design of simple coupling circuits. Usually, a particular amplifier with a single tuned circuit is described in terms of the calculated and measured bandwidth and noise figure. However, a general approach to the analysis and design of a parametric amplifier is not available. Some attempts have been made recently in designing wide-band amplifiers by the use of a more complicated coupling circuit.[1,2] Yet, the ultimate limitations and potentials of a parametric device allowing a general circuit configuration have not been treated.

The present paper studies the parametric converter from the circuit theory point of view. The well-known small signal approach is used, and the nonlinear capacitance is approximated by a periodically varying capacitance. We consider a general circuit configuration where the variable capacitance is imbedded in an arbitrary passive network. The first part of the paper reviews the steady-state analysis of such a circuit based on works of Desoer, Duinker and Rowe.[3–5] Formula for the power gain is derived in terms of the variable capacitance and appropriate impedances of the passive imbedding network. From the general formula, the maximum gain-bandwidth product is then studied by utilizing Bode's familiar relationships.[6]

In this paper we are concerned with only the noninverting-type parametric converter. A forthcoming paper will deal with the inverting-type amplifiers and converters. A straightforward circuit analysis demonstrates that the optimum transducer power gain for a noninverting-type converter is ω'/ω (Manley-Rowe result[7]), where ω and ω', respectively, are the angular frequency of the input signal and the output-upper sideband signal. The maximum fractional bandwidth is approximately equal to $C_1/C_0(\omega_0'/\omega_0)^{1/2}$, where ω_0' and ω_0 are the center frequencies of the two bands. C_1 and C_0 are defined in Section II and are the familiar capacitances used by many authors. Synthesis procedures for coupling networks to approach the optimum gain bandwidth are presented. These are under the assumption that $\omega\omega' = \omega_0\omega_0'$ over the band. A design example indicates that even with the above assumption, a converter with 70 per cent bandwidth can be obtained to achieve a power gain within $\frac{1}{2}$ db from the optimum. The design of a real broad-band converter, where $\omega\omega' = \omega_0\omega_0'$ does not hold, requires the use of an elaborate approximation method.

II. STEADY-STATE ANALYSIS OF A GENERAL CIRCUIT

A. General Background

The circuit shown in Fig. 1 represents a general configuration of a parametric amplifier or converter. The network N is linear, passive and time invariant. The input signal is represented by a sinusoidal current source of angular frequency ω_e, and is applied at port 1 of the network.

$$i_S(t) = \mathrm{Re}[Ae^{j\omega_e t}]. \qquad (1)$$

The output voltage $v_L(t)$ is taken at port 3 and the variable capacitance $C(t)$ is connected to the network at port 2. Our problem is to determine the steady-state voltage $v_L(t)$. However, if the steady-state current i_c

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[1] H. Seidal and G. Herrmann, "Circuit aspects of parametric amplifiers," 1959 IRE WESCON CONVENTION RECORD, pt. 2, pp. 83–90.

[2] G. L. Matthaei, "A study of the optimum design of wide-band parametric amplifiers and up-converters," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-9, pp. 23–38; January, 1961.

[3] C. A. Desoer, "Steady-state transmission through a network containing a single time-varying element," IRE TRANS. ON CIRCUIT THEORY, vol. CT-6, pp. 244–252; September, 1959.

[4] S. Duinker, "General properties of frequency-converting networks," *Philips Res. Repts.*, vol. 13, pp. 37–78, February, 1958; also pp. 101–148, April, 1958.

[5] H. E. Rowe, "Some general properties of nonlinear elements II. Small signal theory," PROC. IRE, vol. 46, pp. 850–860; May, 1958.

[6] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y.; 1945.

[7] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements—part I. General energy relations," PROC. IRE, vol. 44, pp. 904–913; July, 1956.

going through the variable capacitance is known, $v_L(t)$ can be solved, since by the superposition theorem we have

$$V_L(\omega) = Z_{31}(\omega)I_S(\omega) + Z_{32}(\omega)I_c(\omega). \tag{2}$$

$V_L(\omega)$, $I_S(\omega)$ and $I_c(\omega)$ are, respectively, the Fourier transforms of $v_L(t)$, $i_S(t)$ and $i_c(t)$. $Z_{31}(\omega)$ and $Z_{32}(\omega)$ are the transfer impedances of the open circuit 3-port network, N. Thus our problem is reduced to the determination of $i_c(t)$.

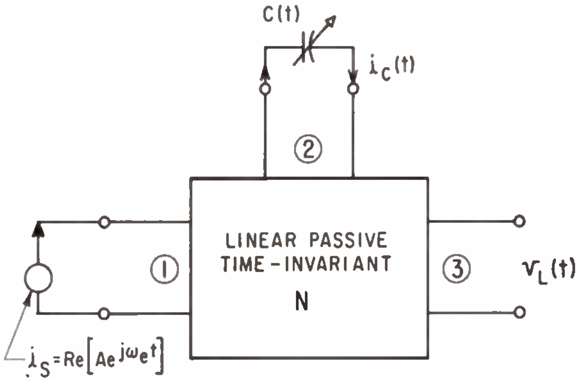

Fig. 1—A general configuration of a parametric amplifier or converter.

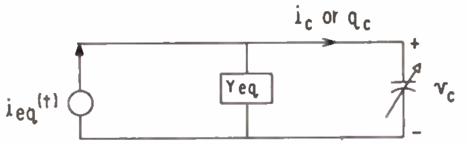

Fig. 2—The Norton's equivalent circuit of the network N looking back from port 2.

The Norton's equivalent circuit of the network N looking back from port 2 is drawn in Fig. 2. Clearly,

$$I_{eq}(\omega) = I_c(\omega) + Y_{eq}V_c(\omega) \tag{3}$$

where Y_{eq} is the equivalent admittance looking back from the variable capacitance, and $I_{eq}(\omega)$ and $V_c(\omega)$ are the Fourier transforms of $i_{eq}(t)$ and $v_c(t)$. The variable capacitance is expressed in terms of a complex Fourier series with a fundamental angular frequency $\bar{\omega}$ (pump frequency).

$$C(t) = \sum_{n=-\infty}^{\infty} C_n e^{jn\bar{\omega}t}, \tag{4}$$

where

$$C_n = \overline{C_{-n}} \qquad (n = 1, 2, \cdots). \tag{5}$$

From (1), (3) and (4) it can be shown that the steady-state components of $i_c(t)$ and $v_c(t)$ contain only frequencies at $\pm\omega_e + m\bar{\omega}$ for all integers of m both positive and negative. The following treatment uses the familiar Fourier series analysis rather than Fourier transforms. Thus capital letters are phasors at single frequencies. Let the charge across the variable capacitance be

$$q_c(t) = \sum_{m=-\infty}^{\infty} \left[Q_{cm}e^{j(\omega_e+m\bar{\omega})t} + Q_{cm}'e^{j(-\omega_e+m\bar{\omega})t}\right], \tag{6}$$

and

$$v_c(t) = \sum_{k=-\infty}^{\infty} \left[V_k e^{j(\omega_e+k\bar{\omega})t} + V_k' e^{j(-\omega_e+k\bar{\omega})t}\right], \tag{7}$$

where

$$Q_{cm}' = \overline{Q_{c-m}} \quad \text{and} \quad V_k' = \overline{V_{-k}} \tag{8}$$

To clarify the various frequency components and notations, a plot of phasors vs frequencies is shown in Fig. 3.

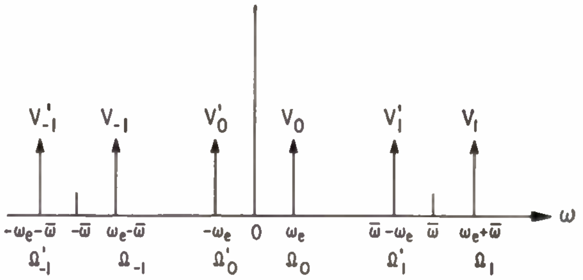

Fig. 3—A plot of various frequency components of interest.

Since

$$q_c(t) = C(t)v_c(t) \tag{9}$$

we can combine (6), (7) and (9) to obtain

$$Q_{cm} = \sum_{i=-\infty}^{\infty} C_i V_{m-i}$$

$$Q_{cm}' = \sum_{i=-\infty}^{\infty} C_i V_{m-i}'. \tag{10}$$

The current i_c can then be expressed as

$$i_c = \frac{dq_c}{dt} = \sum_{m=-\infty}^{\infty} \left[I_{cm}e^{j(\omega_e+m\bar{\omega})t} + I_{cm}'e^{j(-\omega_e+m\bar{\omega})t}\right], \tag{11}$$

where

$$I_{cm} = j(\omega_e + m\bar{\omega})Q_{cm} = j\Omega_m \sum_{i=-\infty}^{\infty} C_i V_{m-i} \tag{12}$$

and

$$I_{cm}' = j(-\omega_e + m\bar{\omega})Q_{cm}' = j\Omega_m' \sum_{i=-\infty}^{\infty} C_i V_{m-i}', \tag{13}$$

where $\Omega_m = m\bar{\omega} + \omega_e$ and $\Omega_m' = m\bar{\omega} - \omega_e$. It is clear from the above that

$$I_{cm}' = \overline{I_{c-m}} \tag{14}$$

and

$$\Omega_m' = \Omega_{-m}. \tag{15}$$

Sometimes it is convenient to express (12) in the matrix form as

$$\begin{bmatrix} \cdot \\ \dfrac{I_{cm-1}}{j\Omega_{m-1}} \\ \dfrac{I_{cm}}{j\Omega_m} \\ \dfrac{I_{cm+1}}{j\Omega_{m+1}} \\ \cdot \\ \cdot \\ \cdot \end{bmatrix} = \begin{bmatrix} \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ \cdots C_2 & C_1 & C_0 & C_{-1} & \cdots & \cdots \\ \cdots & C_2 & C_1 & C_0 & C_{-1} & \cdots \\ \cdots & \cdots & C_2 & C_1 & C_0 & \cdots \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \end{bmatrix} \begin{bmatrix} \cdot \\ V_{m-1} \\ V_m \\ V_{m+1} \\ \cdot \\ \cdot \\ \cdot \end{bmatrix} \tag{16}$$

Similarly, (13) for I_{cm}' can be expressed in a matrix form.

B. Parametric Circuit Analysis

We next consider the special case where $C(t)$ contains only the constant term and the fundamental component as shown in Fig. 4.

$$C(t) = C_0 + 2C_1 \cos \bar{\omega} t, \tag{17}$$

i.e.,

$$C_1 = C_{-1} \quad \text{and} \quad C_n = C_{-n} = 0 \qquad \text{for } n \geq 2.$$

For this simple case, a typical form of (12) is

$$\frac{I_{cm}}{j\Omega_m} = C_1 V_{m-1} + C_0 V_m + C_1 V_{m+1}. \tag{18}$$

The Norton's equivalent circuit of Fig. 2 is redrawn as shown in Fig. 5. We include the fixed capacitance C_0 with Y_{eq} and let

$$Y = j\omega C_0 + Y_{eq}. \tag{19}$$

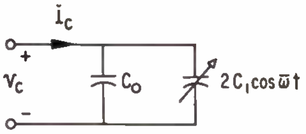

Fig. 4—The variable capacitance, $C(t)$.

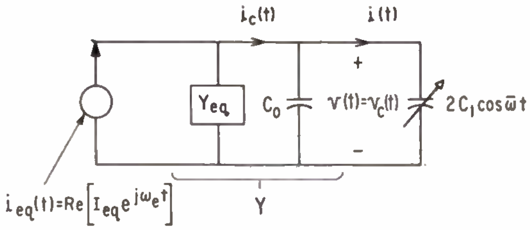

Fig. 5—The Norton's equivalent circuit and the variable capacitance.

The current through the variable capacitance $2C_1 \cos \bar{\omega} t$ and the voltage across it are now designated as $i(t)$ and $v(t)$ as shown in Fig. 5. We put

$$i(t) = \sum_{m=-\infty}^{\infty} \left(I_m e^{j\Omega_m t} + I_m' e^{j\Omega_m' t} \right)$$

and

$$I_m' = \bar{I}_{-m}$$

where I_m and I_m' (without the subscript c) represent the phasors of $i(t)$ and $i'(t)$ at frequencies Ω_m and Ω_m', respectively. The matrix equation relating the phasors of $i(t)$ and $v(t)$ at different frequencies becomes

$$\begin{bmatrix} \cdot \\ \cdot \\ \cdot \\ \dfrac{I_{m-1}}{j\Omega_{m-1}} \\ \dfrac{I_m}{j\Omega_m} \\ \dfrac{I_{m+1}}{j\Omega_{m+1}} \\ \cdot \\ \cdot \\ \cdot \end{bmatrix} = \begin{bmatrix} \cdots & \cdots & \cdots & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ & & & & \\ \cdot\, C_1 & 0 & C_1 & 0 & 0 \\ \cdot\, 0 & C_1 & 0 & C_1 & 0 \\ \cdot\, 0 & 0 & C_1 & 0 & C_1 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ \cdots & \cdots & \cdots & \cdots & \cdots \end{bmatrix} \begin{bmatrix} \cdot \\ \cdot \\ \cdot \\ V_{m-1} \\ V_m \\ V_{m+1} \\ \cdot \\ \cdot \\ \cdot \end{bmatrix}. \tag{20}$$

The corresponding equation pertaining to the I_m' and V_m' need not be written since, for all m, $I_m' = \bar{I}_{-m}$ and $V_m' = \bar{V}_{-m}$. A set of equations relating phasors at different frequencies can be written immediately. With reference to Fig. 5, let

$$Y_i = Y(j\Omega_i). \tag{21}$$

At $\Omega_{-1} = \omega_e - \bar{\omega}$

$$0 = j\Omega_{-1} C_1 V_{-2} + Y_{-1} V_{-1} + j\Omega_{-1} C_1 V_0; \tag{22}$$

at $\Omega_0 = \omega_e$

$$I_{eq} = j\Omega_0 C_1 V_{-1} + Y_0 V_0 + j\Omega_0 C_1 V_1; \tag{23}$$

at $\Omega_1 = \omega_e + \bar{\omega}$

$$0 = j\Omega_1 C_1 V_0 + Y_1 V_1 + j\Omega_1 C_1 V_2. \tag{24}$$

An equivalent circuit can be drawn to represent (22)–

(24). This is shown in Fig. 6, where the node voltages are the phasors V_i and each 2-port coupling network is described by the admittance matrix

$$\begin{bmatrix} 0 & j\Omega_m C_1 \\ j\Omega_{m+1} C_1 & 0 \end{bmatrix}.$$

Thus, the infinite ladder type equivalent circuit describes completely the circuit of Fig. 5. It is important to point out for the capacitance assumed in (17), the above analysis is exact if the circuit is stable.

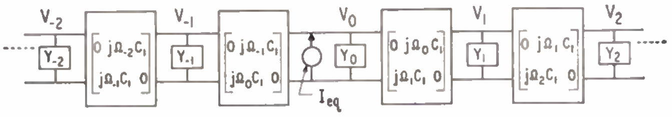

Fig. 6—The exact equivalent circuit which indicates the relationship of voltage phasors at various frequencies.

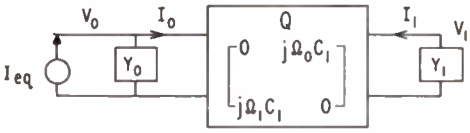

Fig. 7—The equivalent circuit for a noninverting-type converter.

With the above analysis, we consider now the operations of parametric converters and amplifiers. In the usual treatment, it is assumed that there are only two frequency bands present. If the two frequencies of interest are $\Omega_0=\omega_e$ and $\Omega_1=\omega_e+\bar{\omega}$, we have a noninverting-type converter because the output is taken at a frequency Ω_1, which is the sum of the signal and pump frequencies. The assumption made is $V_{-1}=V_2=0$. This is true if $Y_{-1}=Y_2=\infty$. The reason for this can be seen either from Fig. 6 or from (22) at Ω_{-1} and a similar equation at Ω_2. Since for any practical circuit Y_{-1} and Y_2 are not equal to infinity, the assumption that there are only two frequency bands present is only approximately true. However, the effects of such an approximation are usually negligible and will be omitted in this paper. With this assumption, the equivalent circuit for a noninverting-type converter is shown in Fig. 7. The relationships between the current and voltage phasors of the network Q of Fig. 7 are given by

$$\begin{bmatrix} I_0 \\ I_1 \end{bmatrix} = \begin{bmatrix} 0 & j\Omega_0 C_1 \\ j\Omega_1 C_1 & 0 \end{bmatrix} \begin{bmatrix} V_0 \\ V_1 \end{bmatrix}. \tag{25}$$

Thus, the four unknown quantities V_0, V_1, I_0 and I_1 can be solved immediately from (23)–(25).

Similarly, if the frequencies of interest are $\Omega_0=\omega_e$ and $\Omega_{-1}=\omega_e-\bar{\omega}$, we have an inverting-type circuit. The assumption here is $V_{-2}=V_1=0$ or $Y_{-2}=Y_1=\infty$. It turns out for the inverting circuits, amplification can be obtained at both the signal frequency Ω_0 and the converted frequency Ω_{-1}. Thus we have either an amplifier or a converter, depending on the output frequency.

III. Power Gain Formula

In Section II, we have indicated that the current $i(t)$ passing through the variable capacitance $2C_1 \cos \bar{\omega}t$ can be found and represented as

$$i(t) = \text{Re}\left[I_0 e^{j\Omega_0 t} + I_1 e^{j\Omega_1 t}\right] \tag{26}$$

for the noninverting type circuit.

The output voltage v_L of Fig. 2 is of the form

$$V_L(t) = \text{Re}\left[V_{L0} e^{j\Omega_0 t} + V_{L1} e^{j\Omega_1 t}\right]. \tag{27}$$

In this section, we derive the transducer power gain for the noninverting-type converter. As shown in Fig. 8 the source is represented by $i_S=\text{Re}\left[Ae^{j\omega_e t}\right]$ and a source resistance R_S in parallel. The load resistance is R_L and the voltage phasor of interest is V_{L1} at frequency $\Omega_1=\omega_e+\bar{\omega}$.

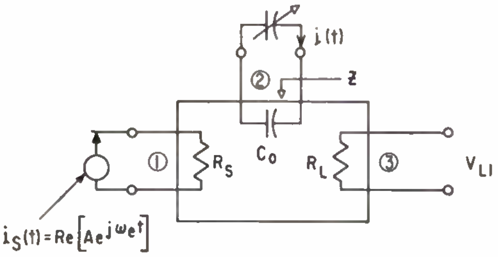

Fig. 8—A general configuration of a parametric converter including the source and the load.

Referring to (22) and (23), let $V_{-1}=V_2=0$, we find

$$V_0 = I_{eq}\frac{Z_0}{1+\Omega_0\Omega_1 C_1^2 Z_0 Z_1}, \tag{28}$$

$$V_1 = I_{eq}\frac{-j\Omega_0 C_1 Z_0 Z_1}{1+\Omega_0\Omega_1 C_1^2 Z_0 Z_1}, \tag{29}$$

where $Z_0=1/Y_0$ and $Z_1=1/Y_1$.

The I_{eq} can be expressed in terms of the source current and appropriate impedances of the fixed network including the capacitance C_0.

$$I_{eq} = \frac{AZ_{21}(\Omega_0)}{Z_0}. \tag{30}$$

Thus,

$$I_0 = j\Omega_1 C_1 V_1 = \frac{AZ_{21}(\Omega_0)}{Z_0}\cdot\frac{\Omega_0\Omega_1 C_1^2 Z_0 Z_1}{1+\Omega_0\Omega_1 C_1^2 Z_0 Z_1}, \tag{31}$$

and

$$I_1 = j\Omega_0 C_1 V_0 = \frac{AZ_{21}(\Omega_0)}{Z_0}\cdot\frac{j\Omega_1 C_1 Z_0}{1+\Omega_0\Omega_1 C_1^2 Z_0 Z_1}. \tag{32}$$

The transducer power gain is defined by

$$G_t = \left| \frac{2V_{L1}}{A\sqrt{R_S R_L}} \right|^2. \tag{33}$$

Since

$$V_{L1} = Z_{32}(\Omega_1) I_1, \tag{34}$$

we have from the above equations

$$G_t = \frac{4(\Omega_1 C_1)^2 \,|\, Z_{21}(\Omega_0) \,|^2 \,|\, Z_{32}(\Omega_1) \,|^2}{R_S R_L \,|\, 1 + \Omega_0 \Omega_1 C_1{}^2 Z_0 Z_1 \,|^2} \cdot \tag{35}$$

We now consider a special configuration as shown in Fig. 9, where network N is a lossless reciprocal 2-port. The resistance R acts as both the source and the load. In practice, a circulator or a band-pass filter is needed to separate out the input and output signals as shown in Figs. 10 and 11.

For convenience of analysis and design, we change the notation as follows: let $\omega = \Omega_0$ be a frequency in the input signal frequency band and $\omega' = \Omega_1$ be a frequency in the converted frequency band. Let I, V, Z be the phasor current, phasor voltage and impedance at frequency ω and I', V', Z' be the corresponding terms at frequency ω'. The transfer impedances, Z_{21} and Z_{32} of (35), can be simplified for the circuit of Fig. 9. In addition, if conservation of power and reciprocity theorem are used, we obtain

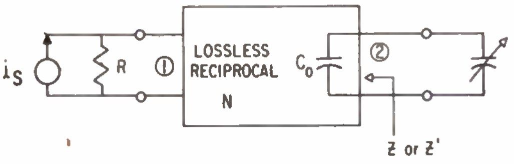

Fig. 9—A special configuration of parametric converter.

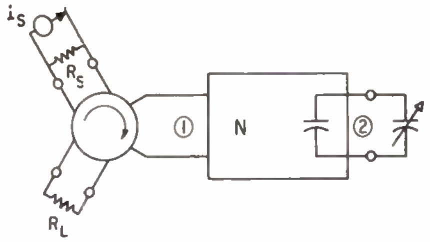

Fig. 10—A circulator-type parametric converter.

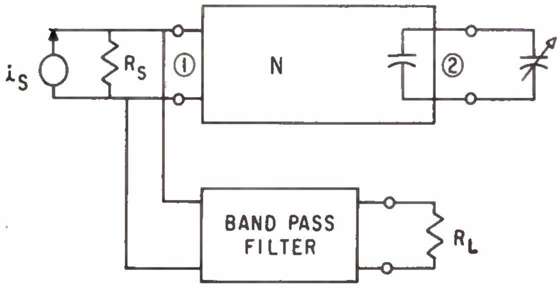

Fig. 11—A filter-type parametric converter.

$$|\, Z_{21} \,|^2 = |\, Z_{32} \,|^2 = R \cdot \operatorname{Re} Z. \tag{36}$$

Eq. (35) becomes

$$G_t = \frac{4(\omega' C_1)^2 \operatorname{Re} Z \operatorname{Re} Z'}{|\, 1 + \omega\omega' C_1{}^2 Z Z' \,|^2}, \tag{37}$$

where Z and Z' are the impedances looking back from the variable capacitance at frequencies ω and ω', respectively, as shown in Fig. 9. This equation can further be interpreted in the following manner: First we redraw the complete equivalent circuit of the noninverting-type converter by adding the source and the load to the circuit of Fig. 7. This is shown in Fig. 12. N and N' are the same network except that all computations pertaining to N are at ω and those pertaining to N' are at ω'. Let Z_{in} be the input impedance looking to the right of network N as shown in Fig. 12; then using the relations specified by the 2-port Q,

$$Z_{\text{in}} = \frac{1}{\omega C_1 \omega' C_1 Z'} \cdot \tag{38}$$

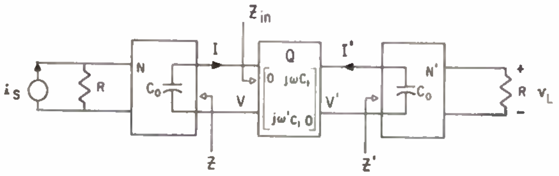

Fig. 12—The complete equivalent circuit for a non-inverting-type converter.

The 2-port Q can be interpreted as an impedance inverter. Substituting (38) in (37), we get

$$G_t = \left(\frac{\omega'}{\omega}\right) \frac{\operatorname{Re} Z \operatorname{Re} Z_{\text{in}}}{|\, Z + Z_{\text{in}} \,|^2} \cdot \tag{39}$$

The second factor in (39) has a maximum of unity for passive impedances Z and Z_{in} under the condition

$$Z_{\text{in}} = \bar{Z}. \tag{40}$$

This is the familiar conjugate matching condition. Therefore, we conclude that if

$$\frac{1}{\omega C_1 \omega' C_1 Z'} = \bar{Z}, \tag{41}$$

the transducer power gain is the maximum and is equal to ω'/ω, which is the familiar result of Manley and Rowe.

Let ω_0 and ω_0' be the center frequencies of the two bands. Thus

$$\omega = \omega_0 \pm \Delta \quad \text{and} \quad \omega' = \omega_0' \pm \Delta. \tag{42}$$

Clearly,

$$\frac{\omega'}{\omega} = \frac{\omega_0'}{\omega_0}\left(\frac{1 \pm \Delta/\omega_0'}{1 \pm \Delta/\omega_0}\right) \tag{43}$$

is a function of Δ. The maximum realizable gain is not a constant over a broad band. Equalization may be needed if the required tolerance in the pass band is important, even though (41) is satisfied over a broad band.

IV. GAIN-BANDWIDTH LIMITATION AND DESIGN ILLUSTRATIONS

In this section we derive an approximate formula for the maximum realizable percentage bandwidth and present the optimum design method based on the approximation that $\omega\omega' = \omega_0\omega_0'$ holds over the band. With the above approximation (41) is rewritten in the following form:

$$\bar{Z} Z' = \frac{1}{\omega\omega' C_1^2} \approx \frac{1}{\omega_0\omega_0' C_1^2} \cdot \tag{44}$$

Therefore in order to obtain the maximum gain over a given band, we need to design a passive coupling network such that (44) is satisfied. Moreover, the network is restricted to start with a shunt capacitance C_0, which limits the bandwidth. Let the driving point impedances at the two bands be written as

$$Z = |Z| e^{j\theta} \qquad Z' = |Z'| e^{j\theta'}. \tag{45}$$

From (44), we have

$$|Z| \, |Z'| \, e^{j(\theta'-\theta)} = \frac{1}{\omega_0\omega_0' C_1^2} \cdot \tag{46}$$

Since the right-hand side is purely real,

$$\theta = \theta', \tag{47}$$

(46) becomes

$$|Z| \, |Z'| = \frac{1}{\omega_0\omega_0' C_1^2} \cdot \tag{48}$$

The problem is then simplified and can be stated in terms of the design of a familiar 2-terminal interstage network. A driving-point impedance which starts with a shunt capacitance C_0 is to be designed to satisfy (48) over a maximum possible bandwidth. Bode has considered the following related problem:[8] A driving-point impedance Z which starts with a shunt capacitance C_0 has a maximum constant magnitude $|Z|$ over a low pass interval from zero to ω_c,

$$|Z| = \frac{2}{\omega_c C_0} \tag{49}$$

as shown in Fig. 13. The driving-point admittance for this optimum case is

$$Y = \frac{j\omega C_0}{2} + \frac{\omega_c C_0}{2}\sqrt{1 - \left(\frac{\omega}{\omega_c}\right)^2}. \tag{50}$$

The circuit as shown in Fig. 14 is actually the parallel combination of a capacitance $C_0/2$ and an infinitely long constant K image filter. It will be seen that this low-pass circuit can be used directly in our problem. Bode has also shown that if the magnitude of Z is to be maximized over more than one pass band, the phase must be ±90° outside the pass bands. For our problem we have two frequency bands of interest; the magnitude of the impedances over the two bands $|Z|$ and $|Z'|$ must satisfy (48). The bandwidths of the two pass bands are to be maximized. Therefore, we conclude that a typical phase characteristic of the 2-terminal impedance is given by Fig. 15. Since the phase is more or less specified except for $\theta = \theta'$ in the pass band, the magnitude is also more or less specified. In order to satisfy (48) it is clear that we have to choose ω_∞ such that

$$|Z| \approx |Z'| \approx \frac{1}{(\omega_0\omega_0')^{1/2} C_1} \cdot \tag{51}$$

This can be readily obtained by means of a low-pass to multiband-pass transformation of Bode's optimum low-

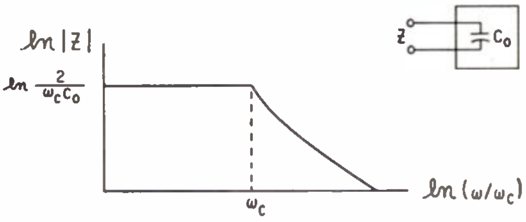

Fig. 13—Bode's optimum low-pass interstage behavior.

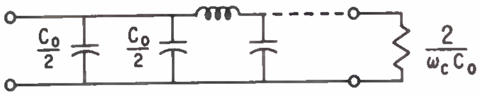

Fig. 14—Bode's optimum low-pass 2-terminal interstage network.

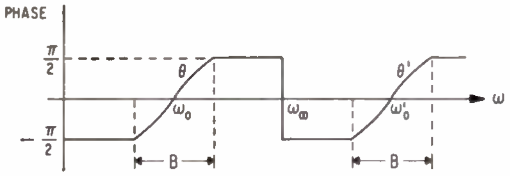

Fig. 15—Typical phase characteristics of an optimum 2-terminal impedance.

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[8] Bode, *op. cit.*, see ch. 17.

pass configuration. Let

$$p = \frac{(s^2+\omega_0^2)(s^2+\omega_0'^2)}{s(s^2+\omega_\infty^2)}, \tag{52}$$

$$\omega_\infty^2 = \frac{\omega_0^2+\omega_0'^2}{2}, \tag{53}$$

where p is the low-pass variable and s is the band-pass variable. Eq. (53) is obtained based on the condition $Z \approx Z'$. The transformation is shown in Fig. 16. Let the angular bandwidth of the two pass bands be B. From the conservation of bandwidth we have

$$B = \frac{\omega_c}{2}. \tag{54}$$

Combining (49), (51) and (54), we obtain the important result which states that the limitation of the fractional bandwidth is given by

$$\frac{B}{\omega_0} = \frac{C_1}{C_0}\left(\frac{\omega_0'}{\omega_0}\right)^{1/2}. \tag{55}$$

The gain-bandwidth product is thus

$$\sqrt{G_t}\,\frac{B}{\omega_0} = \frac{C_1}{C_0}\,\frac{\omega_0'}{\omega_0}. \tag{56}$$

A practical realization of the admittance of (50) or the circuit of Fig. 14 is by means of the usual M-derived termination. Bode has tabulated a few simple circuits which approximate the ideal case along with their magnitude behavior in his book.

To illustrate the above theory, the following examples are given: Let the center frequencies of the signal band and the converted band be

$$f_0 = 10^9 \text{ cycles},$$
$$f_0' = 10^{10} \text{ cycles}.$$

We want to have three designs which use three different variable capacitances with

$$C_0 = 10^{-12} f$$

and

$$C_1 = \begin{matrix}(a)\\(b)\\(c)\end{matrix}\begin{cases}0.05 \times 10^{-12} f\\ 0.15 \times 10^{-12} f.\\ 0.3 \times 10^{-12} f\end{cases}$$

From the formula for the optimum gain, we have at the center of the band

$$G_t = 10 \text{ or } 10 \text{ db}.$$

The percentage bandwidth limitations according to (55) for the three cases are

$$\text{per cent bandwidth} = \begin{matrix}a)\\b)\\c)\end{matrix}\begin{cases}15.8\\47.4\\94.8\end{cases}$$

The normalized equivalent low-pass circuit used is a simple M-derived $\frac{1}{2}$ section given by Bode as shown in Fig. 17. The actual circuit after the transformation specified by (52) and (53) and a frequency denormalization is shown in Fig. 18. The calculated results of the transducer power gain are plotted in Figs. 19–22. It is seen that for case a), the actual gain follows almost exactly the optimum in the complete pass band. For cases b) and c) the actual gain differs from the optimum only near the edge of the band. The realized bandwidth is about 75 per cent of the maximum. This is because of the mismatch of the simple M-derived circuit around the cutoff frequency. As Bode has shown, a more complicated low-pass circuit such as shown in Fig. 23 can almost realize the maximum possible percentage bandwidth.

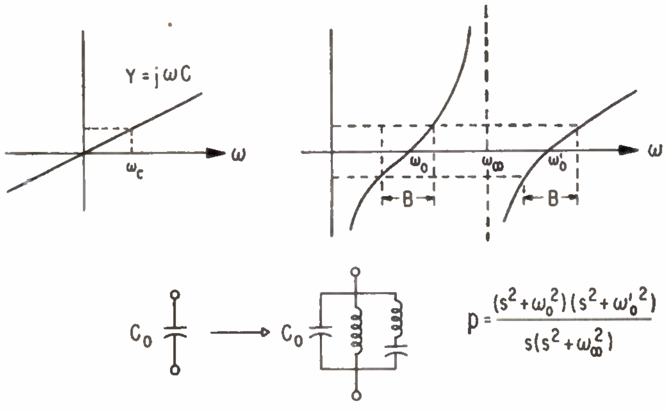

Fig. 16—Low-pass to multiband-pass transformation.

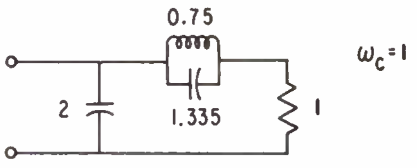

Fig. 17—Normalized low-pass coupling circuit.

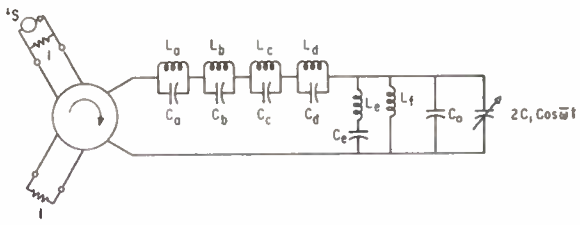

Fig. 18—The actual circuits for the three specified parametric converters.

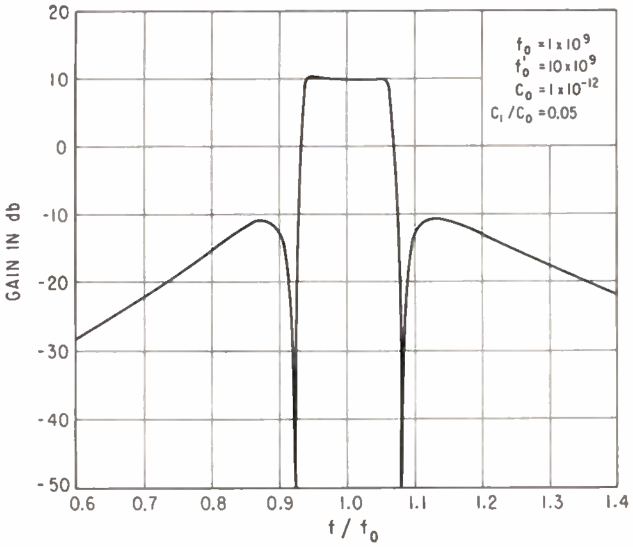

Fig. 19—Transducer power gain vs ω/ω_0 for $C_1/C_1=0.05$.

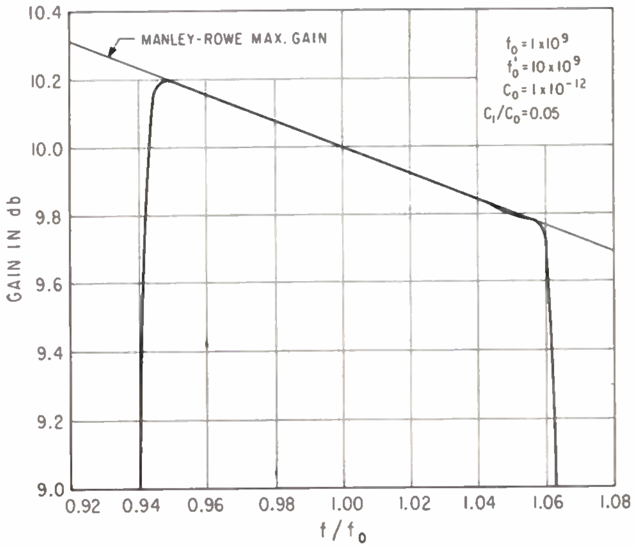

Fig. 20—Transducer power gain vs ω/ω_0 for $C_1/C_1=0.05$.

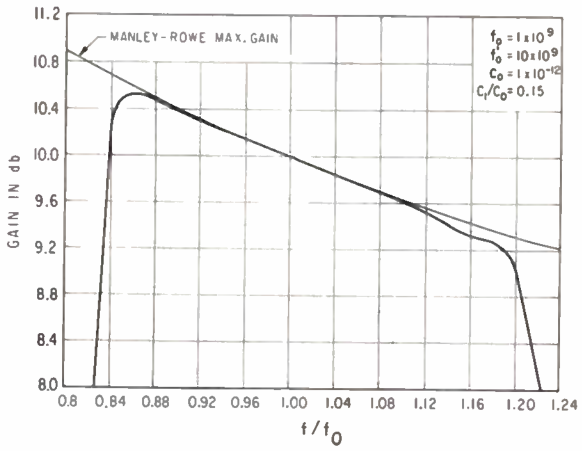

Fig. 21—Transducer power gain vs ω/ω_0 for $C_1/C_1=0.15$.

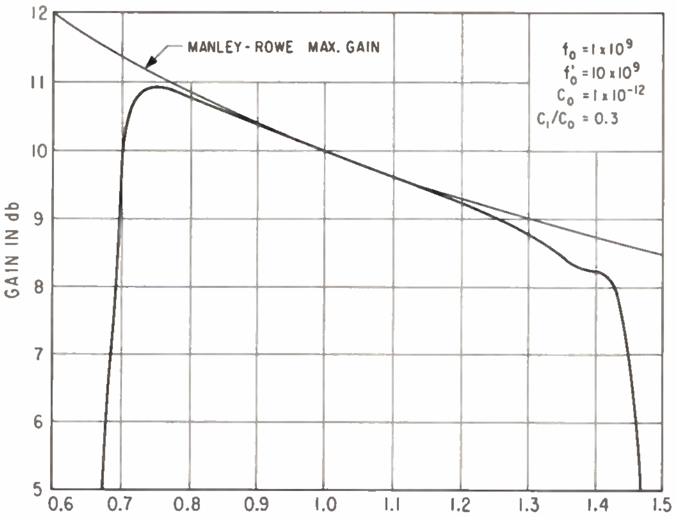

Fig. 22—Transducer power gain vs ω/ω_0 for $C_1/C_1=0.3$.

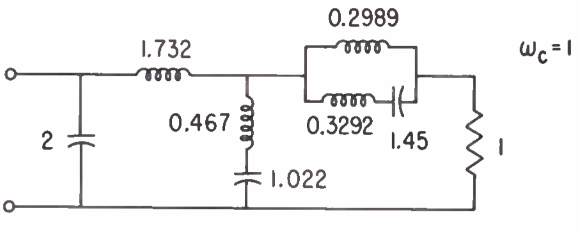

Fig. 23—A normalized low-pass coupling circuit which approaches the optimum behavior.

In conclusion, it must be pointed out again that the above theory and design technique are based on the approximation $\omega\omega'=\omega_0\omega_0'$. For a real broad-band situation where the above approximation does not hold, we have to refer back to (44), where $\bar{Z}Z'$ is a function of frequency. An elaborate network approximation procedure is necessary to determine the poles and zeros of the driving-point impedance.

ACKNOWLEDGMENT

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A New Precision Low-Level Bolometer Bridge*

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Summary—**A low-level RF bridge has been developed for the purpose of calibrating directly the outputs of signal generators over a wide frequency band. Using standard commercial bolometers giving an input resistance of 50 ohms and an error detection scheme based on raidometer techniques, the bridge has a measured sensitivity ($S/N=1$) of 3.0×10^{-10} w. The basic bridge design is described in this paper and various applications are discussed. Although designed primarily for RF measurement, the bridge can be applied to the detection of any kind of electromagnetic radiation provided suitable bolometer elements and modulating means can be arranged.**

INTRODUCTION

THE bridge described in this paper was developed during the course of an investigation carried out for the Signal Corps aimed at developing techniques to permit rapid, accurate, and smiplified measurements of extremely low-level RF voltages and powers normally associated with signal generators. The ultimate goal was to measure low-level RF powers to 10^{-16} w with 3 per cent accuracy over as wide a frequency range as possible with one instrument. At these levels, major problems are created by the requirements of low ambient noise and high sensitivity.

To cope with these problems, two approaches were investigated: 1) superconducting and semiconducting bolometers designed to exploit the benefits of cryogenic temperatures, and 2) the application of radiometer principles to the measurement of low-level signals in noise at ambient temperatures. Various aspects of the cryogenic phase of the work have been described previously [1], [2]. The ambient temperature phase has resulted in the development of a prototype bridge described herein.

BRIDGE DESCRIPTION

General

Fig. 1 is a block diagram of the bridge which is designed to compare the heating effect of a pulse of RF power to that of a dc pulse on a bolometer element. The RF and dc pulses are supplied alternately at an audio rate (100 cps) the period of which is long compared to the thermal time constant of the bolometer. Under these conditions, if the power supplied by the RF source is not equal to that due to the dc pulse, the resistance of the bolometer will vary periodically at the audio rate. A resistive bridge is used to detect this variation of bolometer resistance. The bridge is excited by a dc source which is automatically controlled in order to maintain the average resistance of the bolometer constant. The effect of bolometer instability is thus reduced to a second-order error in the measurement, since the substituted power is not affected by these variations. With this method, it is possible to detect differences in power of the order of magnitude of the noise output of the bolometer. The value of the substituted power can be calculated to an accuracy dependent only upon techniques available for dc measurements. The RF power will differ from this only by a factor determined by the efficiency of the bolometer, mount, modulator, and RF transmission system. Because it is necessary to use a bias power very much larger than the substituted power, in order to detect small changes in the bolometer resistance, the bias power must be known to within the percentage accuracy required for the RF measurement. This is a result of the fact that the substituted power is determined not by the change in current through the bolometer alone, but by both the bias current and the change in current. With this system, the maximum sensitivity is determined by the bandwidth of the detection circuits (which can be made quite small by the use of a long integration time). The factors which will ultimately limit the sensitivity are the choice of a practical observation time and the stability of the RF source.

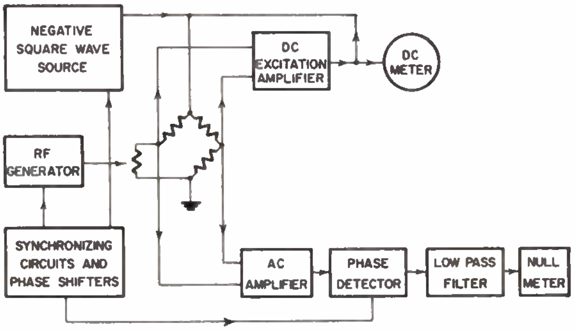

Fig. 1—Block diagram of bridge.

Basic Bridge

A symmetrical design has been used for the bridge with each arm having a resistance of 200 ohms. Two 100-ohm bolometers are arranged in a Narda Model 560 mount so that they are in parallel for the RF power (providing a 50-ohm termination) and in series for the dc power (200 ohms). The other three arms of the bridge consist of 200-ohm precision wirewound resistors. This type was chosen for stability and low noise. Tests

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† The Franklin Institute Laboratories for Research and Development, Philadelphia, Pa.

showed that the predominent sources of noise were the Johnson noise of the fixed resistors and bolometers and temperature fluctuation noise of the bolometers.

DC Square Wave Source

The value of the dc power which is substituted for the RF power depends upon the bias current in the bridge, which in turn is a function of the bolometer characteristic. Several methods are possible for adding the dc pulse to the bias current. The method chosen was to shunt the bridge with a variable resistance as shown in Fig. 2, thereby bleeding off a small portion of the bias current. This has the advantage that the value of the dc power is determined completely by the bias current and the value of the shunt. There is no need for a calibrated square wave source. Variation of the shunt resistance is obtained by shorting out a small portion of the shunt by means of a mechanical chopper. The accuracy of this method depends upon the contact resistance of the chopper, the accuracy of the resistors, and the accuracy of the dc voltmeter which measures the bridge voltage.

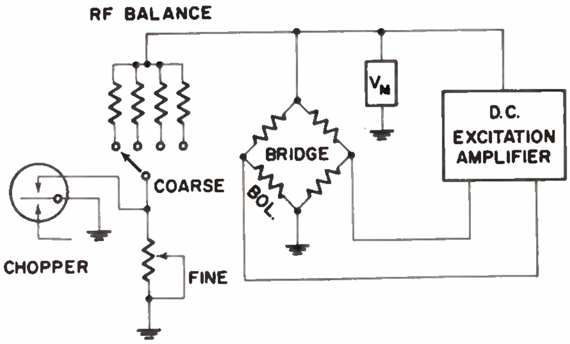

Fig. 2—DC pulse insertion.

DC Excitation Amplifier

The dc excitation amplifier maintains the dc balance of the bridge by controlling the bias current in such a way as to keep the average resistance of the bolometers equal to 200 ohms. It consists simply of a Kin Tel Model 204A dc amplifier, followed by a pentode current amplifier. A pentode was chosen since it is desirable that the amplifier have a large known output impedance to limit the current through the bolometers to a safe value in the event of failure of any of the components. Filtering was added to the plate circuit to reduce noise from the dc power supply, and a variable cathode resistor was used to provide a control of the bias current. This control permits the error voltage of the dc feedback loop to be adjusted to zero, thus reducing the required gain.

AC Amplifier and Phase Detector

To detect the 100 cps unbalance signal, a tuned ac amplifier was used. The input was transformer coupled in order to improve the noise figure and provide an ungrounded input. The final noise figure was estimated to be 1.8 db, the bandwidth about 6 cps. In order to avoid the use of tuned circuits within the amplifier, which would be bulky and cause coupling problems, selectivity was obtained by means of positive feedback around RC band-pass networks. Sufficient negative feedback was used in the amplifiers to maintain reasonable gain and bandwidth stability. Each stage was adjusted for a Q of between 10 and 15. The principal need for narrow bandwidth was to reduce hum pickup, which otherwise would have saturated the amplifier. The input transformer was shielded against electrostatic and magnetic pickup and was shock-mounted to avoid microphonics. The secondary was tuned to the chopper frequency of 100 cps by means of a parallel capacitor, and a series capacitor was used in the primary to avoid a dc short across the bridge. The filaments were supplied from a dc source to reduce hum pickup as much as possible.

The phase detector, Fig. 3, consisted of a current amplifier, the output of which was switched alternately to two storage capacitors by a reference derived from the 100-cps modulation signal. The output was measured between the two storage capacitors.

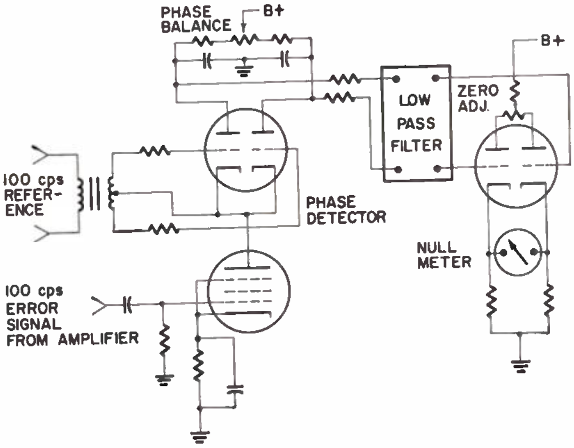

Fig. 3—Phase detector and null meter.

RF Modulation

Several methods of modulating the output of the signal generator were tried. In order to obtain the desired accuracy, small insertion loss and a large on-off ratio are required. Best results for internal modulation were obtained with a mechanical RF chopper, driven at the 100-cps rate. Most electronic or solid-state modulators did not have either a sufficient on-off ratio or required calibration of the insertion loss or gain. For signal generators which have a provision for internal square wave modulation, an output was also provided to supply the required input signal. However, this method is not preferred, since most signal generators cannot be modulated with a 100-cps square wave with any degree of accuracy. Furthermore, the output meters generally indicate average power and must be corrected for the modulation duty cycle. This means that the pulse widths must

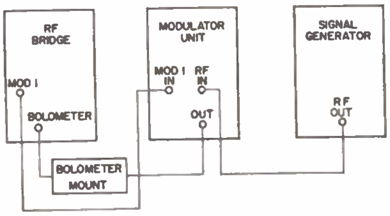

Fig. 4—Power bridge interconnections.

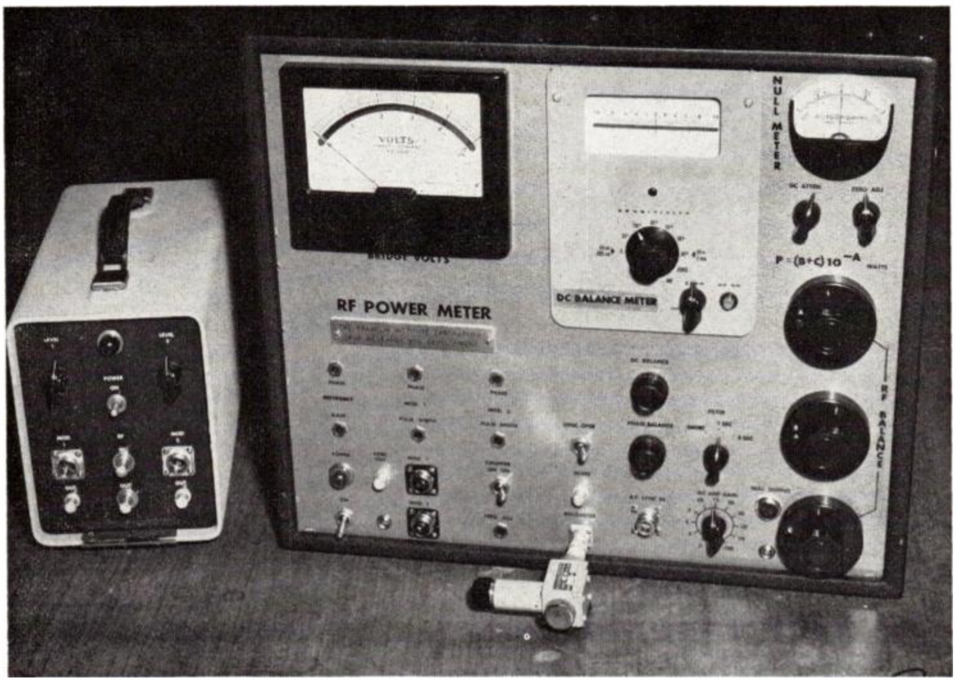

Fig. 5—Photo of bridge and modulator.

be measured precisely. Fig. 4 demonstrates the manner in which the bridge and modulator would be connected to the signal generator for calibration, and Fig. 5 is a photograph of the complete bridge and modulator unit.

BRIDGE PERFORMANCE

The detectability of the bridge is limited mainly by noise fluctuations in the output caused by Johnson and temperature fluctuation noise at the bolometer. The indicated mean square temperature fluctuation of the bolometer is given by

$$\overline{\theta^2} = \frac{2kT^2}{\beta\tau}$$

where

θ = temperature deviation from ambient
k = Boltzmann's constant
β = coefficient of heat transfer
τ = time of averaging
T = temperature of the bolometer.

The solution to the steady heat balance equation, $E = \beta\theta$, provides a means of relating this temperature fluctuation to an equivalent fluctuation in input power. Substituting, we have for the mean square noise power $\overline{p_t^2}$, equivalent to the temperature fluctuation

$$\overline{p_t^2} = \beta^2 \overline{\theta^2} = \frac{2kT^2\beta}{\tau}.$$

A similar treatment can be made for the Johnson noise voltage by means of the expression for bolometer sensitivity which relates output voltage to input power. If S is the bolometer sensitivity, the equivalent mean square power fluctuation p_j^2 due to Johnson noise is

$$\overline{p_j^2} = \frac{2kTR}{\tau S^2}.$$

By definition,

$$S = V/E = \frac{i\delta R}{\beta\delta T} = \frac{ibR\delta T}{\beta\delta T} = \frac{ibR}{\beta},$$

where V is the change in voltage across the bolometer due to the change in the bolometer resistance, δT is the mean temperature rise caused by detector current i, and b is the temperature coefficient of resistance, $b = 1/R(dR/dT)$. Substituting for S,

$$\overline{p_j^2} = \frac{2kT\beta^2}{\tau i^2 b^2 R}.$$

Therefore, the total mean square noise equivalent power is

$$\begin{aligned}\overline{p^2} &= \overline{p_t^2} + \overline{p_j^2} \\ &= \frac{2kT^2\beta}{\tau} + \frac{2kT\beta^2}{\tau i^2 b^2 R}\end{aligned}$$

and this will be minimum detectable signal defined as $(S/N=1)$. To take account of the fact that the bolometer temperature T will be different from the ambient temperature T_0 when bias current is applied to the bolometer, we should write the above expression as

$$\overline{p^2} = \frac{k\beta}{\tau}(T^2 + T_0^2) + \frac{2kT\beta^2}{\tau i^2 b^2 R}.$$

In the current model of the bridge, two Narda N610B bolometers with the following characteristics are used:

$$S = 4500 \text{ ohms per watt} = \frac{bR}{\beta}$$

$$R = 100 \text{ ohms}$$

$$b = 0.0038$$

$$\beta = \frac{bR}{S} = 8.43 \times 10^{-5} \text{ watt per degree.}$$

The rated current is 8.75 ma making the mean temperature rise $\bar{\theta} = 90.8°\text{C}$. Substituting these values in our equation for equivalent noise power, we find that the detection limit of the bridge $(S/N=1)$ should be 1.5×10^{-11} w for a time constant of 8 sec.

By giving careful attention to the reduction of stray pickup in the detection circuits, we have been able to achieve an over-all detection limit of 3×10^{-10} w. This approaches the ultimate limit to within one order of magnitude. This figure can be improved by increasing the integration time, but long time constants, in addition to being very annoying to work with, are limited by the stability of the source being measured. In the prototype bridge, an 8-sec time constant was found to be optimum. A 1-sec output was also provided for more rapid initial adjustments and for work at higher signal levels.

In the checks made thus far with the bridge, excellent agreement has been obtained in comparison with the other standards down to 10^{-8} w at frequencies ranging up to 100 Mc. The present RF chopper, James, FTC-159-1171, used in the modulator unit is limited to frequencies below 500 Mc. This is mechanical two-position switch driven by an electromagnetic actuator which switches the input of the bolometer mount alternately between the RF source and a terminating resistance. The high-frequency performance is limited by RF leakage across the switch and by the discontinuity introducd into the RF line due to the switch. With suitable detectors and modulators, however, the bridge is not subject to any frequency limitation. In fact we hope later to use the bridge in the measurement of low-level radiation in the optical range as well as at radio frequencies. Also, with the bridge providing a means of establishing a low-level RF signal of high accuracy, the use of attenuators and a sensitive radio detector should permit measurements to be carried out by comparison techniques at the level of 10^{-16} w as desired in the above-mentioned goals. The accurate low-level source will greatly reduce errors due to leakage and high attenuation formerly present in previous comparison techniques.

Figures on absolute accuracy of the bridge are not available since an exact calibration of the bolometer and mount efficiency have yet to be made. The comparison checks, however, definitely establish the accuracy as being well within the desired 3 per cent over the lower frequency end of the bolometer range. The main goals now are to extend the detection capability of the bridge through the use of vacuum mounted bolometers and some circuit modifications and, then, to couple the basic bridge to low temperature bolometers provided the proper characteristics can be achieved.

ACKNOWLEDGMENT

The authors would like to express their appreciation to M. M. Hirsch, formerly a member of the staff at The Franklin Institute, who conceived many of the features used in the design of the bridge and whose originality and perseverance contributed immeasurably to the success of the project. We are also indebted to R. G. Keyes of the Signal Corps for his stimulating discussions and helpful suggestions.

BIBLIOGRAPHY

[1] D. L. Birx and N. Fuschillo, "The theory of low-temperature bolometer detector applied to the measurement of low-level RF power," IRE TRANS. ON INSTRUMENTATION, vol. I-7, pp. 310–315; December, 1958.
[2] B. Lalevic, "Criteria for the choice of a superconducting bolometer," *J. Appl. Phys.*, vol. 31, pp. 1234–1236; July, 1960.
[3] A. Van Der Ziel, "Noise," Prentice-Hall, Inc., New York, N. Y., p. 409; 1954.

P-N Junction Charge-Storage Diodes*

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***Summary*—The design theory for a new range of *p-n* junction applications is presented. The applications include pulse generation, wave shaping, and harmonic generation. The diodes are characterized by a very abrupt interruption of reverse current in the turn-off transient and are approximately ideal nonlinear capacitors.[1] The abrupt interruption of current in the reverse transient is related to the impurity profile in the junction. An estimate is given of the duration of the abrupt phase. In addition, the role of parasitic elements such as inductance, capacitance, and series resistance is discussed in relation to a particular representative circuit. In typical cases, the abrupt turn-off phase lasts for a time of the order of 10^{-9} sec. Transitions in excess of 100 v or an ampere are readily obtained.**

I. INTRODUCTION

THE reverse recovery property of *p-n* junctions has been a subject of considerable interest in the application of diodes to switching and rectification.[2,3] The diodes that Kingston,[2] Lax and Neustadter[3] considered were step junctions of various geometries with the qualitative result that the reverse recovery consisted of a *storage phase* followed by a *decay phase.* In their idealized circuits, steps of voltage were applied with the result that the storage phase consisted of a period of constant reverse current which persisted until the carrier density near the junction reached zero. The decay phase represented the current due to the residual stored carriers. It was found that both the storage phase and the decay phase lasted for times comparable to carrier lifetime. The diodes that we are considering differ qualitatively in their recovery characteristic from the step junction. The storage phase lasts for a time comparable to lifetime, but the time for the decay may be several orders of magnitude less than carrier lifetime.

In switching applications, the ideal recovery time is zero. That is, any reverse conduction tends to degrade the performance of the device. There is a group of applications, however, where the required recovery time is finite and determined within limits by the circuit.[4,5] The diode is biased in the forward direction by either part of a sinusoidal wave or else by a steady bias current and turned off either by the negative portion of the sinusoid or by a negative pulse. In distinction to the ideal computer diode or switch, the diodes that we are considering here are designed to conduct for a period of time in the reverse direction. When the stored minority carriers are depleted, a very abrupt step in current occurs. This abrupt step is useful in rectangular wave-shaping as well as harmonic generation. The transients for the ideal conventional switching diode as well as the class of diodes under consideration are compared in Fig. 1.

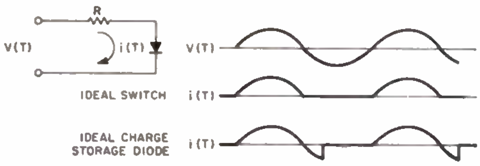

Fig. 1—Current transients for the ideal switch and the ideal storage diode. The ideal switch has no reverse conduction whatsoever. In the ideal storage diode, the reverse conduction is equal to the forward conduction, and the abrupt drop in current occurs in zero time.

The reverse conduction is obtained by having finite carrier lifetime (usually as long as is conveniently obtainable from the fabrication process). The abrupt step is obtained by designing the diode so that when the carriers at the junction are depleted, essentially all of the minority carriers have been removed. The junctions are built with retarding fields for minority carriers with the result that storage is constrained to the vicinity of the junction. Thus, we might define the class of diodes under consideration as *carrier storage diodes with constraining fields*, or, since the constraining fields are necessary to high performance (fast switching), just *storage diodes.* The ideal storage diode should have the following properties:

1) Infinite capacitance in forward direction (zero incremental voltage drop).
2) Infinitely fast switching from reverse storage conduction to cutoff.
3) Zero capacitance in reverse direction.
4) Long minority carrier lifetime.
5) No excessive parasitic elements. A small amount of series resistance is tolerable, and it will be seen that a small amount of series inductance may be desirable.

In the first part of this paper, we are concerned with the intrinsic limitations of the physical electronics on

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[1] D. Leenov and A. Uhlir, Jr., "Generation of harmonics and subharmonics at microwave frequencies with *p-n* junction diodes," PROC. IRE, vol. 47, pp. 1724–1729; October, 1959.

[2] R. H. Kingston, "Switching time in junction diodes and junction transistors," PROC. IRE, vol. 42, pp. 829–834; May, 1954.

[3] B. Lax and S. F. Neustadter, "Transient response of a *p-n* junction," *J. Appl. Phys.*, vol. 25, pp. 1148–1154; September, 1954.

[4] A. F. Boff, J. Moll and R. Shen, "A New High Speed Effect in Solid State Diodes," 1960 International Solid-State Circuits Conference—Digest of Technical Papers.

[5] A. F. Dietrich and W. Goodall, "Solid-state generator for 2×10^{-10} second pulses," PROC. IRE, vol. 48, pp. 791–792; April, 1960.

switching speed. We will see that the fast cutoff occurs in a time approximately x_0^2/D where D is an average diffusion constant for the injected carriers and x_0 is the distance from the center of the junction to the center of gravity of the injected charge distribution. To make the turn-off time small, we must have the injected carriers concentrated very close to the junction. This can be accomplished with a retarding field as is encountered in a graded junction.[6]

The calculation of the fast turn-off rate is most easily accomplished by first considering the dc conduction properties. Generally, if a retarding field is used to confine the carriers to the junction, the maximum carrier density for a given stored charge will be greater than if there is no retarding field. The increased carrier density will result in conductivity modulation and a decrease in the retarding field. For this reason, calculation of the center of gravity of the injected carriers is a nonlinear problem. Since charge distribution is closely related to the diode current-voltage characteristic, the observed static characteristics serve as an internal check on the calculations.

The qualitative result is that as stored charge is increased, the retarding fields decrease and the distribution of minority carriers becomes relatively more diffuse. That is, with increase in stored charge, the center of gravity of the stored minority charges moves away from the junction, and the cutoff transition becomes slower. The quantitative relations between stored charge per unit area and distribution of the stored charge is basic to a quantitative design theory for the maximum rate of turn-off.

It is necessary, in any given circuit application, to compromise between diode capacitance which acts as a passive limitation on the fast transition and the stored charge per unit area which limits the fast rate of recovery of minority carriers. For given operating conditions, the junction capacitance varies directly and stored charge density inversely with junction area. Section II is concerned with the distribution of the stored charge and its limitation on turn-off rate. We then consider the limitation of the passive parameters, and finally the compromise that must be made between diode capacitance and stored charge per unit area to obtain optimum transition speed.

II. Distribution of Stored Charge

Experiment and theory on the forward characteristic of the storage diode serve to provide an internal check on the calculation of the charge distribution and thus give confidence in the calculation of the transient. Shockley[6] has given an extensive discussion of the fields that exist at equilibrium in a graded junction. In this reference it is shown that a junction with large gradient develops a depletion region, whereas an extremely gradual junction remains essentially neutral throughout. The steepness of the junction is measured by

$K = L_D/2L_a$

$L_D = (kT\epsilon/2q^2n_i)^{1/2}$ = Debye length

$L_a = n_i/a$ = length in which impurity density changes by an amount equal to the intrinsic density

a = concentration gradient

n_i = intrinsic carrier density. (1)

If $K \gg 1$ there is space change, and if $K < 1$ the junction is neutral. The assumption of constant quasi-Fermi level in the junction gives the same formal behavior in the nonequilibrium case except that it is necessary to substitute a new *intrinsic* density n_i' for n_i. Thus if

$$\begin{aligned} L_D' &= (kT\epsilon/2q^2n_i')^{1/2} \\ L_a' &= n_i'/a \\ n_i' &= n_i\epsilon^{qV/2kT} \\ K' &= L_D'/2L_a'. \end{aligned} \tag{2}$$

Then if $K' < 1$ the junction is neutral.

Substitution for n_i' in L_D' and L_a' results in

$$K' = K\epsilon^{-3qV/4kT}. \tag{3}$$

At equilibrium, $(V=0)K'$ reduces to K and rapidly decreases from K as forward voltage is applied. It is possible, and in fact is commonly encountered, that the junction starts at zero bias with a depletion region $(K \gg 1)$ and goes through an intermediate range, finally ending up as a neutral junction $(K' < 1)$. This is the case for all of the junctions that we will be discussing in this paper. Morgan and Smits[7] have extended Shockley's calculations to include nonequilibrium conditions with the sole limitation that the quasi-Fermi level is constant throughout the central region of the junction. Their calculation shows that if the factor K is less than unity, the potential distribution obtained from the neutral approximation is valid. Sah[8] has obtained some approximate formulas for total junction capacitance for linear graded junctions, including diffusion capacitance under forward bias. In Sah's treatment, he assumes that generation-recombination current[9] in the transition region accounts for all of the conduction current. Detailed calculations[10] suggest that, in many practical instances, diffusion current and recombination in neutral regions account for most of the conduction current in silicon

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[6] W. Shockley, "The theory of p-n junctions in semiconductors and p-n junction transistors," *Bell Sys. Tech. J.*, vol. 28, pp. 435–489; July, 1949.

[7] S. P. Morgan and F. Smits, "Potential distribution and capacitance of a graded p-n junction," *Bell. Sys. Tech. J.* vol. 39, pp. 1573–1602; November, 1960.

[8] C. T. Sah, "Effects of electrons and holes of the transition layer characteristics of linearly graded p-n junctions," Proc. IRE, vol. 49, pp. 603–618; March, 1961.

[9] C. T. Sah, R. N. Noyce and W. Shockley, "Carrier generation and recombination in p-n junctions and p-n junction characteristics," Proc. IRE, vol. 45, pp. 1228–1243; September, 1957.

[10] J. Moll, unpublished work.

junctions with forward bias greater than about 0.5 to 0.5 v. Transition region recombination and generation account for most of the current for bias up to about 0.4 to 0.5 v forward bias. We shall be dealing with a forward bias range of 0.7 to 0.8 v and hence will ignore generation-recombination current. A further difference in our treatment of the junction as compared to that of Sah[8] is the terminology relating to transition region. When the depletion or space charge region has disappeared, we will simply treat the junction as a neutral region. For formal reasons, Sah defines a transition region that extends essentially through the region of modulation of the equilibrium conductivity. Our approximations can be extended to give the same result as that obtained by Sah[8] for the case of sufficient forward bias that the junction be neutral (usually about 0.5 v).

The important storage diode parameters are switching time, carrier lifetime, and values of parasitic elements such as transition region capacitance, series resistance and inductance.

In order to obtain the highest possible retarding fields, the storage diodes are fabricated by the diffusion of boron and phosphorus into opposite faces of a thin slice of silicon. Using a slice thickness in the range of 25 to 50 μ it is possible to obtain significantly higher effective retarding fields than would be obtained in a linear graded junction with equal junction gradient. The devices may be given a *long lifetime* treatment[11] to minimize recombination. The diffusion surface concentrations and times are set so that the resulting impurity profile is very nearly symmetrical. The impurity density in such a diode can be approximated as

$$N_D - N_A \cong 2N_0 \sinh bx, \tag{4}$$

where x is distance measured from the junction and N_0 and b are determined from the surface concentration and diffusion temperature and time. For a completely symmetrical diffusion,

$$b \cong \frac{W}{4Dt} \qquad (W > 4\sqrt{Dt}) \tag{5}$$

W = wafer thickness

D = diffusion constant

t = diffusion time.

Also,

$$N_0 = N_s \operatorname{erfc}(W/4\sqrt{Dt})$$

N_s = surface concentration. (6)

At normal injection densities, we can assume that we have charge neutrality, or

$$\rho = q(p - n + N_d - N_a) \cong 0. \tag{7}$$

The assumption (7) of charge neutrality makes possible the calculation of potential in the small region of constant quasi-Fermi level near the junction. We can make the substitution

$$p(x) = p^* \epsilon^{-q\Psi(x)/kT}$$
$$n(x) = p^* \epsilon^{+q\Psi(x)/kT}, \tag{8}$$

where $\Psi(x)$ is the potential with respect to the potential at $x=0$ and p^* is the electron or hole density at the junction ($x=0$).

The boundary conditions require that

$$(p^*)^2 = pn = n_i^2 \epsilon^{qV/kT}, \tag{9}$$

where V is the applied voltage. Substitution of (8) into (7) results in

$$p^* \sinh(q\Psi/kT) = N_0 \sinh bx. \tag{10}$$

Fig. 2 shows the impurity density distribution and potential variation. Fig. 3 shows the variations of $\partial\Psi/\partial x$ and p with distance under different injection densities. The electric field is

$$-\frac{\partial\Psi}{\partial x} = \frac{kT}{q}\,\frac{N_0 b}{p^*}\,\frac{\cosh bx}{\cosh \beta\Psi} \tag{11}$$

where

$$\beta = \frac{q}{kT}\,.$$

It is seen from (11) that at $x=0$, the field is inversely proportional to the injection level p^*, while at a certain distance away from the center, it becomes constant and independent of p^*.

It is very difficult, analytically, to deal with a variable field, so we will replace the variable field by an effective constant field that results in the same amount of charge storage as the variable field. If $\langle E\rangle$ is the average field, the amount of stored hole charge is

$$\frac{Q_{p0}}{A} = qp^*kT/q\langle E\rangle, \tag{12}$$

where

Q_{p0} = stored charge (holes)

p^* = hole density at the junction.

Eq. (12) is valid if the quasi-Fermi level is constant throughout most of the region of charge storage (or if $\sqrt{D_p\tau_p} \gg kT/q\langle E\rangle$) as is the case for the diodes under consideration. The current that flows will be the ratio of stored charge to lifetime. If the appropriate average field is the field near the junction (which varies inversely with p^*) the forward current would vary as $\exp(qV/kT)$; whereas, if the field far from the junction (which is independent of injection level) is most important, the forward current would vary as $\exp(qV/2kT)$. Measurements of forward current characteristics showed

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[11] G. Bemski, "Recombination in semiconductors," Proc. IRE, vol. 46, pp. 990–1004; June, 1958.

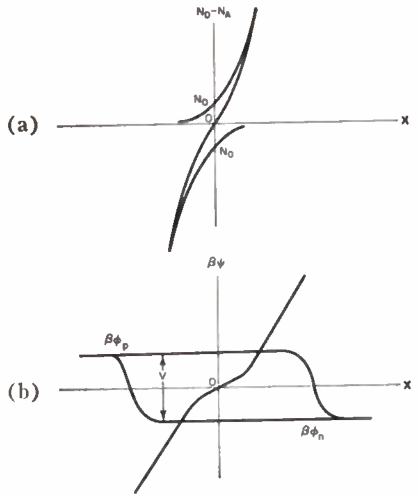

Fig. 2—Impurity density (a) and potential variation (b) in a storage diode with forward bias applied.

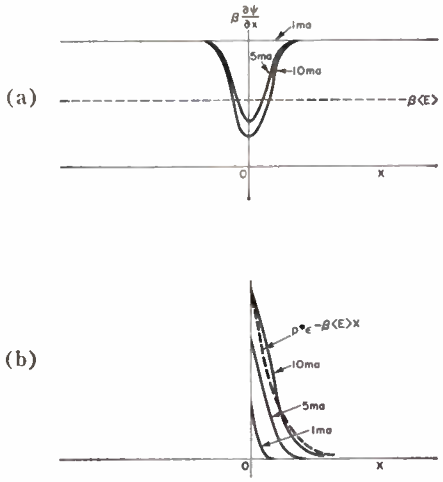

Fig. 3—Electric field (a) and carrier density (b) for a storage diode as a function of forward current.

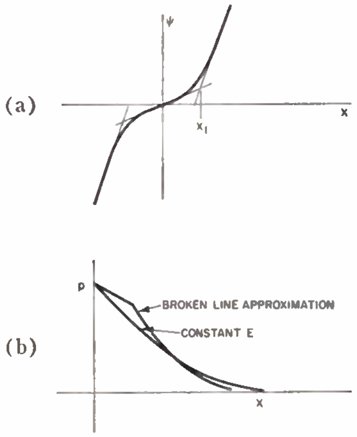

Fig. 4—Broken line approximation to potential in junction (a), and resulting hole-density distribution (b).

exp $(3qV/4kT)$ which indicates that the average field should be the geometric mean of the field at the junction and the field far away. This empirical result can be justified by approximating the potential distribution of Fig. 2 with straight lines as in Fig. 4 such that

$$
\begin{aligned}
\Psi &= E_1 x \qquad 0 \lll x \lll x_1\\
\Psi &= \frac{kT}{q}\ln\frac{E_1}{E_2} + E_2 x \qquad x_1 \lll x \lll \infty,
\end{aligned}
\tag{13}
$$

where

$$E_1 = \frac{kT}{q}\,\frac{2N_0 b}{p^*}; \qquad E_2 = \frac{kT}{q}\,b$$

$$x_1 = \frac{kT}{q}\,\frac{\ln(E_1/E_2)}{(E_1 - E_2)};$$

then the minority hole distribution is approximately

$$
\begin{aligned}
p &= p^*\epsilon^{-qE_1x/kT} \qquad 0 \lll x \lll x_1\\
p &= p^*\epsilon^{-q(E_1-E_2)x_1/kT}\epsilon^{-qE_2x/kT} \qquad x_1 \leq x \leq \infty.
\end{aligned}
\tag{14}
$$

The stored charge of holes Q_{p0}/A for constant quasi-Fermi level is

$$
\begin{aligned}
\frac{Q_{p0}}{A} &= q\int_0^{x_1} p^*\epsilon^{-(qE_1x/kT)}dx\\
&\quad + q\int_{x_1}^{\infty} p^*\epsilon^{-(q/kT)(E_1-E_2)x_1}\epsilon^{-(qE_2x/kT)}dx.
\end{aligned}
\tag{15}
$$

A is the diode area and Q_{p0} is the total stored charge of holes. There is a similar charge of electrons on the p-side. By equating the stored charge of (12) to (15) the effective constant field $\langle E\rangle$ can be expressed in terms of E_1 and E_2 as

$$\langle E\rangle = f(E_1/E_2)\sqrt{E_1E_2}, \tag{16}$$

where

$$f(y) = \frac{1}{\sqrt{\frac{1}{y}} + \left(\sqrt{y} - \sqrt{\frac{1}{y}}\right)y^{\left(\frac{y}{1-y}\right)}}$$

and is plotted in Fig. 5.

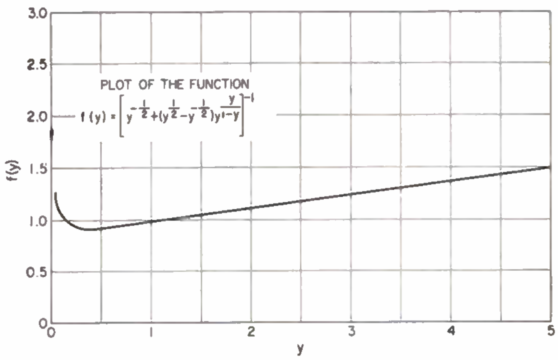

Fig. 5—Plot of $f(y)$ vs y.

It is seen that for the range of operation where $0.1E_2 < E_1 < 4E_2$

$$\langle E \rangle \approx \sqrt{E_1 E_2} = \frac{kTb}{q} \sqrt{\frac{N_0}{p^*}} \cdot \tag{17}$$

This range encompasses the range of impurity distributions and current that we normally use. We now substitute the approximate relation (17) into (12), and take the boundary condition into account with the result that

$$\frac{Q_{p0}}{A} = \frac{q}{b\sqrt{2N_0}} n_i^{3/2} \epsilon^{3qV/4kT} = \frac{Q_{n0}}{A} \cdot \tag{18}$$

The total current is $Q_{p0}/\tau_p + Q_{n0}/\tau_n$. Eq. (18) shows that the stored charge varies (in the forward direction) as exp $(3qV/4kT)$. The internal consistency of the theory of forward conduction is easily checked experimentally by making independent measurements of area, lifetime, voltage, current, junction gradient, and from knowledge of the diffusion profile. A practical junction has[12]

$N_0 = 6.7 \times 10^{15}$ cm^{-3}
$b = 7.5 \times 10^3$ cm^{-1}
$A = 1.47 \times 10^{-3}$ cm^2
$\tau = 3.3 \times 10^{-7}$ sec
$I = 10$ ma
$V = 0.75$ v
$n_i = 1.6 \times 10^{10}$/cm^3 (*i.e.*, silicon at 300° K).

From (17),

$$\frac{q}{kT} \langle E \rangle = b \sqrt{\frac{N_0}{p^*}} = 3.66 \times 10^3 \text{ cm}^{-1}. \tag{19}$$

From (12),

$$\frac{q}{kT} \langle E \rangle = \frac{2qAp^*}{Q_{p0} + Q_{n0}} = \frac{2qAp^*}{I\tau} = 3.9 \times 10^3 \text{ cm}^{-1}. \tag{20}$$

For (20), the electron and hole charges are simply added and the average lifetime is used. The lifetime was measured by an injection-extraction method.

The close agreement between the effective average field as calculated in (19) and (20) is good evidence that the calculated carrier distribution is very similar to the one that is actually obtained. Detailed calculation of the change in quasi-Fermi level shows that it is constant to within a few millivolts to distances as great as 10^{-3} cm from the junction for both electrons and holes. Most of the minority electrons and holes are contained in this region—hence the close agreement of (19) and (20).

The effect of recombination on the hole density is easily seen by assuming a constant field E in the junction. For this case the hole density for a few tenths of a volt or more forward bias is

$$p = p^* \epsilon^{-k_p x}$$

$$k_p = \frac{1}{2}\left(\left(\frac{q\langle E \rangle}{kT}\right) + \sqrt{\left(\frac{q\langle E \rangle}{kT}\right)^2 + \frac{4}{L_p^2}}\right), \tag{21}$$

where $L_p = \sqrt{D_p \tau_p}$ is the diffusion length for holes. In our practical example, $2/L_p \approx 10^3$cm^{-1} and is sufficiently small (L_p sufficiently large) so that the recombination term is insignificant in determining carrier distribution.

III. Storage Phase

The storage time T_s is the portion of the recovery transient during which the conductance remains high. The storage time T_s for the diode can be found most easily by assuming that all of the charge is pulled out or recombined during storage time. The storage time is obtained from the equation of continuity for charge. Thus, the total stored charge Q is obtained from

$$\frac{dQ}{dt} = I - \frac{Q}{\tau}, \tag{22}$$

where I is the conduction current across the junction and τ is the recombination lifetime. This form of charge continuity applies to step junctions as well as graded or retarding field junctions and *p-i-n* junctions. Eq. (22) is of particular utility for retarding field junctions (storage junctions) since these junctions have the property of containing very little residual stored charge at the end of the storage phase. The storage time for sinusoidal applied waves as well as rectangular waves can be obtained from (22). We will restrict ourselves to rectangular waves for this paper, however.

We assume that the diode is biased in the forward direction by a current I_f. If the duration of the current is several times the lifetime τ, a steady-state stored charge Q_0 will be obtained. If the duration is comparable to carrier lifetime or less, it will be necessary to solve (22) for Q_0, the stored charge at the beginning of the reverse transient. If at $t=0$ the current is reversed to $-I_r$, the storage time is

$$T_s = \tau \ln\left(1 + \frac{Q_0}{I_r \tau}\right), \tag{23}$$

where $Q_0 = I_f \tau$ if the forward current has been on a time $T_f \gg \tau$. In general, the initial charge is

$$Q_0 = I_f \tau (1 - e^{-T_f/\tau}), \tag{24}$$

where T_f = period of forward conduction.

Eqs. (23) and (24) are actually slight over estimations of storage time, since there is a finite residual charge left in the junction at the end of the storage phase, as illustrated in Fig. 6.

---

[12] In this case, the parameter K is approximately 7×10^6 at equilibrium and reaches a value of 10^{-3} at 0.75 v forward bias so the neutrality condition at the forward bias condition is well satisfied.

Lifetime is not a fixed constant of the diode. It tends to increase about 50 per cent for a 70°C temperature rise. Smaller junction diodes tend to have a lower lifetime and lower temperature coefficient because of an increased ratio of surface to body recombination. Lifetime also tends to decrease at high stored charge values because of nonlinear recombination and because the stored charge extends into lower lifetime regions. Lifetime may drop also at very low stored charge values because traps remain unfilled. Also, as the junction gradient is increased, there is a tendency for lifetime to decrease. A reduction in lifetime by as much as 50 per cent may be encountered when high storage is required.

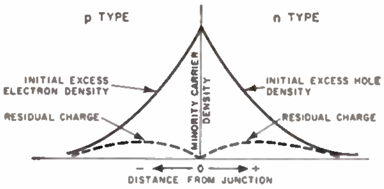

Fig. 6—Excess electron and hole density showing initial charge distribution and residual charge at the end of the storage phase.

IV. Transition Phase

The current that is conducted from the small bumps of residual charge can be calculated by a method that is analogous to the one used by Kingston[2] for step junctions. In the diode with retarding fields the reverse current is conducted by a combination of drift and diffusion during the storage phase, whereas the current in the step junction is conducted entirely by diffusion. At the end of the storage phase, however, all of the current in the retarding field diode must be diffusion current since the hole (and electron) density at the junction is approaching zero. Thus, we can calculate the gradient of carrier density at the junction. We will make the following analytical approximation to the residual carrier density:

$$\Delta p = \frac{I_r(\text{holes})}{qD_pA}\, x\, \epsilon^{-x/x_0} \qquad x > 0$$

$$\Delta n = \frac{I_r(\text{electrons})}{qD_nA}\, x\, \epsilon^{+x/x_0} \qquad x < 0, \tag{25}$$

where

$$x_0 = kT/q\langle E\rangle.$$

For a symmetrical junction it is a reasonable assumption that the electron and hole currents are equal. In this case $I_r(\text{holes}) = I_r(\text{electrons}) = \frac{1}{2}I_r$.

The analytical form (25) is suggested by more detailed calculations in the Appendix. It is valid only if (54) is satisfied, or alternately

$$\frac{x}{x_0} < 2\left(\frac{L_p}{x_0}\right)^2 \frac{I_f}{I_r}. \tag{26}$$

Since the lifetime is supposed to be very long, the value of $|x/x_0|$ may be quite large.

The actual drop of the reverse current follows a very complicated function of time. Since we are interested only in the approximate cutoff transition time, we assume for simplicity a reverse current decreasing exponentially with time.

$$i_r(t) = I_r\epsilon^{-t/T_q}. \tag{27}$$

The cutoff transition time is usually short compared to the lifetime; we can, therefore, neglect the recombination loss of carriers during the cutoff phase. We then have

$$\int_0^\infty i(t)dt = \Delta Q_p + \Delta Q_n, \tag{28}$$

where $\Delta Q_p + \Delta Q_n$ is the residual charge of electrons and holes, and time is measured from the end of the storage phase. It follows from (25) and the assumption that the electron and hole currents are equal that

$$I_rT_q = \frac{I_rx_0^2}{2}\left(\frac{1}{D_p} + \frac{1}{D_n}\right) \tag{29}$$

or

$$T_q = \frac{x_0^2}{D} \quad \text{where} \quad D = \frac{2D_pD_n}{D_p + D_n}. \tag{30}$$

It is interesting to note that the decay time that is calculated by our procedure reduces to the diffusion time from the original center of gravity of the injected carriers to the junction if the diffusion constants are equal. With the use of (12), (17), and (25) we obtain T_q as

$$T_q = \frac{1}{D}\left(\frac{(Q_0/A)}{(4q)(b)(N_0b)}\right)^{2/3}. \tag{31}$$

Note that N_0b is the gradient of impurities in the junction and to a large extent controls the breakdown voltage. As the gradient is increased the decay time as well as reverse breakdown voltage decreases. Also, if the gradient is held constant but the distance $1/b$ is decreased (b increased), the decay time decreases. Increase in b is accomplished in practice by using a very thin wafer and diffusing a relatively short time. A decrease in stored charge per unit area will also decrease the decay time. Thus, for a given circuit application, the diode capacitance which tends to slow down response must be compromised against the decay time.

V. Design Compromise for Specific Applications

Dynamic response of the storage diode depends upon a complex interaction between the diode and its associated circuit. The important diode design parameters are:

1) Junction gradient.
2) Junction area.
3) Minority carrier lifetime.
4) Package impedance (chiefly series inductance).

We will examine how circuit conditions determine the optimum values for these parameters. There is no single universally optimum design for the charge-storage diode, but only a best balance of factors for given operating conditions.

Behavior of the storage diode is demonstrated in the circuit of Fig. 7. Switch S is assumed initially open and so diode current $i_d = I_f$ and $V_d = V_f \approx \Psi_0$. (Ψ_0 is the forward applied voltage across the diode minus IR drop.) At time t_0 switch S is closed, causing the diode current polarity to reverse. A voltage spike may be apparent across the diode ($e_d = -L_d(di_d/dt)$) during this current change. Diode conductivity will be maintained for a time because of the stored minority carriers which have been injected by I_f. During this storage conduction phase, voltage across the diode will remain small (essentially the forward voltage minus a drop due to current and diode series resistance). Diode current will, therefore, be determined chiefly by the external circuit. When the stored charge becomes depleted, conduction of the diode will drop to the low value typical of its reverse biased state.

For any given application, interest centers upon the time at which the diode switch opens (storage phase) and the rapidity with which it opens (transition phase). The considerations relating to storage time have already been discussed. We will be concerned with optimizing the fast transition time for a given stored charge and breakdown voltage.

If the junction conductivity were assumed to drop instantaneously at the end of the storage phase (no residual charge), the observed transition characteristic would be determined only by impedances associated with the diode. These significant impedances are: the equivalent Thévenin source impedance looking out from the diode terminals, diode lead inductance, and diode shunt capacitance. This equivalent circuit is shown in Fig. 8 where cessation of diode conduction is associated with the opening of switch S.

Since the incremental capacitance of the diode is strongly voltage-dependent, this problem is nonlinear and so does not lead to a simple solution. The approximate contribution of the various impedance elements can be evaluated by considering each of the contributions separately.

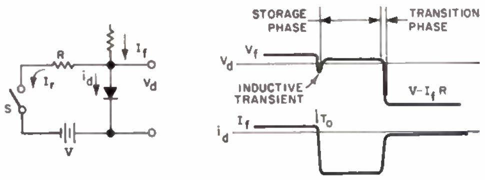

Fig. 7—Diode waveforms during switching.

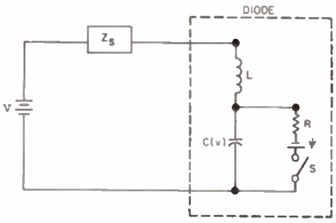

Fig. 8—Equivalent circuit for the charge-storage diode assuming an instantaneous drop in junction conductivity.

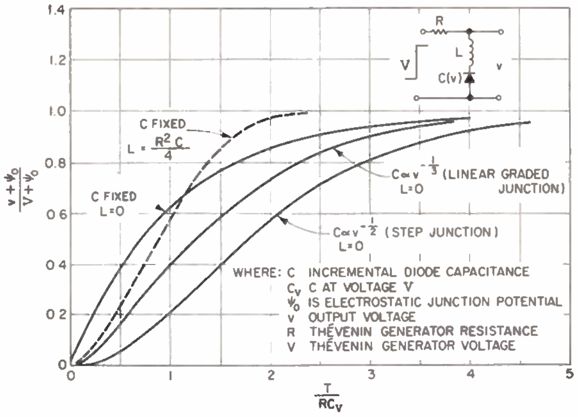

Fig. 9—Storage diode transition characteristic for *RC* limited operation.

For the effect of nonlinear capacitance, assume Z_s purely resistive and zero inductance. The resulting response is plotted in Fig. 9 for different assumed voltage dependencies of $C(V)$. The coordinates for this plot are simply the ones which normalize the analytic solutions. Transition is taken here as starting always at $v = -\Psi_0$ and ending at $v = V$. The horizontal time axis normalizes with respect to the incremental value of diode capacitance with full reverse voltage at the end of transition across the diode. It is apparent from Fig. 9 that very little difference in rise time results from the different capacitance voltage dependencies that are plotted. It seems plausible, therefore, that the effective diode capacitance can be represented by the constant value C_v. From this plot, 10–90 per cent transition time, when limited exclusively by RC considerations, would be

$$T_{RC} = 2.5\, RC_v. \tag{32}$$

It would be anticipated from this formulation that when diode capacitance limits transition speed, a high voltage transition will occur more rapidly than low voltage transition. This effect is observed.

Diode series inductance inevitably arises in conjunction with diode packaging. This is not necessarily a deleterious factor, but can serve to improve the transition speed considerably if its value is appropriately chosen.

The manner in which series inductance L is calculated to influence transition voltage is shown in Fig. 10. This plot assumes instantaneous cessation of diode conduction and a fixed value C_v for diode capacitance. The optimum value for L when C_v and R are given is the critically damped value

$$L_{\text{crit}} = \frac{R^2 C_v}{4}. \tag{33}$$

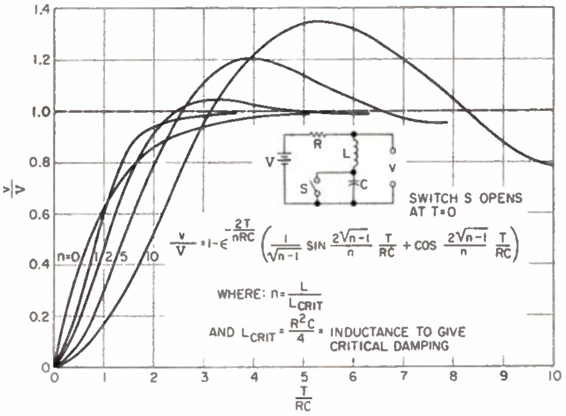

Fig. 10—Transient response for inductance variation.

This inductance about halves the pure RC transition time.

Assuming critical inductance and zero residual diode charge during transition, the optimum impedance limited transition time T_{RLC} is approximately

$$T_{\text{RLC}}\Big|_{10-90\text{ per cent}} \approx 1.25\, RC_v \qquad \left(\text{for } L = \frac{R^2C_v}{4}\right) \tag{34}$$

Assuming now that inductance is chosen at the critical value, we can find the optimum diode parameters for a given set of operating conditions. Minority carrier lifetime τ must be evaluated on the basis of storage phase requirements. It does not enter directly into transition phase consideration except insofar as it influences Q_0, the initial stored charge. The two remaining diode parameters that may then be optimized are the junction gradient and the junction area. These are determined from consideration of the transition phase and required reverse breakdown voltage V_B.

From (31)

$$T_q\Big|_{10-90\text{ per cent}} = 2.3T_q = \frac{1.15}{D}\left(\frac{Q_0/A}{(2q)(b)(N_0b)}\right)^{2/3} \tag{35}$$

(the factor 2.3 comes from definition of transition time as 10–90 per cent of the transition voltage).

This expression relates the gradient N_0b and junction area A to transition time when the influence of diode impedance may be ignored. Diode capacitance C_v may be related to other diode parameters as follows:

$$C_v = \frac{\epsilon A}{w} = \left(\frac{q\epsilon^2 N_0 b}{6V}\right)^{1/3} A, \tag{36}$$

so (34) may be expressed as

$$T_{\text{RLC}} = 1.25RA\left(\frac{q\epsilon^2 N_0 b}{6V}\right)^{1/3} \tag{37}$$

Minimum breakdown voltage establishes an upper limit upon the permitted gradient. An empirical expression for V_B in linear graded junctions in silicon is

$$V_B = \left(\frac{10^{25}}{3a}\right)^{0.4}, \tag{38}$$

where $a = 2N_0b$ is the gradient of impurities at the junction. With choice of maximum allowable gradient a, the junction area may be optimized for minimum time. Assuming that maximum permitted gradient has been chosen, we have

$$\left\{\begin{aligned} T_q\Big|_{10-90\text{ per cent}} &= G_qA^{-2/3} \\ &, \\ T_{\text{RLC}}\Big|_{10-90\text{ per cent}} &= G_{\text{RLC}}A \end{aligned}\right. \tag{39}$$

where

$$G_q = \frac{0.725}{D}\left(\frac{Q_0}{qb^2N_0}\right)$$

$$G_{\text{RLC}} = 0.66\left(\frac{q\epsilon^2 N_0 b}{V}\right)^{1/3} R.$$

The resulting transition time is taken conservatively as the sum of the impedance and residual charge transition time contributions considered individually.

$$\begin{aligned} T_t &= T_q\Big|_{10-90\text{ per cent}} + T_{\text{RLC}}\Big|_{10-90\text{ per cent}} \\ T_t &= G_qA^{-2/3} + G_{\text{RLC}}A. \end{aligned} \tag{40}$$

The value of A required to minimize T is

$$A = \left(\frac{2}{3}\,\frac{G_q}{G_{\text{RLC}}}\right)^{3/5} = 0.7\left(\frac{Q_0^2V}{(q\epsilon b)^2(N_0bR)^3}\right)^{1/5}, \tag{41}$$

and the calculated value for transition time under these conditions is

$$T_{t\,\text{min}} = 1.96G_q^{3/5}G_{\text{RLC}}^{2/5}$$

$$T_{t\,\text{min}} = 1.26\left[\left(\frac{\epsilon}{V}\right)\left(\frac{1}{N_0b}\right)^2\left(\frac{Q_0R}{qb}\right)^3\right]^{2/15} D^{-3/5}. \tag{42}$$

It is shown in the Appendix that the transition time may actually vary more slowly with injected charge than is indicated by (31), especially when the field far from the junction is much larger than the field near the junction. This modification results in a somewhat smaller transition time and, for a given breakdown voltage, a somewhat smaller optimum area.

VI. EXPERIMENTAL OBSERVATIONS

Figs. 11 and 12 show typical storage diode recovery characteristics as taken with a sampling oscilloscope driving an X-Y pen recorder. Extraction current was held constant and forward current was varied. The two test diodes were selected at capacitance values ($C_0 = 7$ and 50 pf) to demonstrate, respectively, the preponderance of stored charge density and capacitive loading on transition time limitation. Comparing these figures, the 7-pf diode exhibits a faster transition (due to its lower

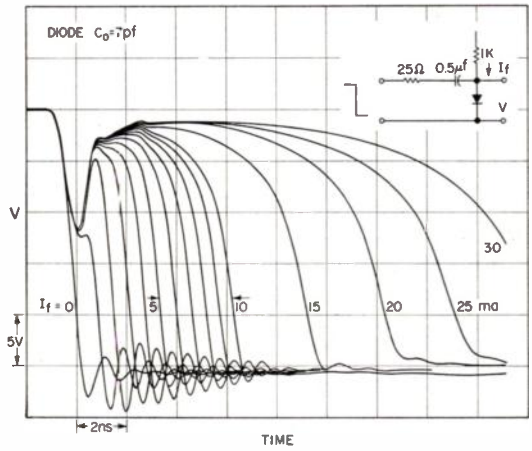

Fig. 11—Storage diode transition characteristic ($C_0=7$ pf) as a function of forward current, showing degradation due to charge dispersion at high storage density levels.

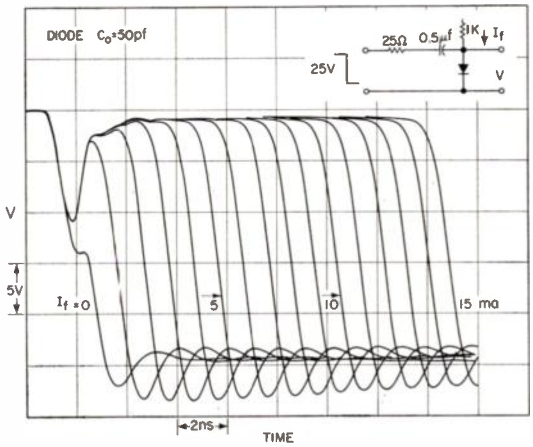

Fig. 12—Storage diode transition characteristic ($C_0=50$ pf) as a function of forward current, showing degradation due to capacitive loading.

capacitive loading) than does the 50-pf diode only at low forward current. Transition time of the 7-pf diode deteriorates very much more rapidly with increasing forward current than does the 50-pf diode because its stored charge density is higher. The negative spike at the upper left as well as the ringing are due chiefly to diode inductance. It can be seen that the lifetime of the higher capacitance diode is almost double that of the lower capacitance since storage time is almost double for the same forward current. Instrument rise time was 0.7 nsec.

VII. Conclusions

The design theory for a class of diodes which is useful for short pulse generation and wave shaping as well as harmonic generation has been presented. The design features of interest are the length of the storage phase of the reverse transient and the fast transition time at the end of the storage phase. The storage phase is very similar to the storage phase encountered in all *p-n* junctions, and its duration is determined by the lifetime of minority carriers, magnitude of stored charge and extraction rate. The lifetime of the minority carriers is not subject to design in a simple manner, but depends on the diode area, profile of impurity density, temperature, injection level, and recombination trap density.

The fast transition phase is independent of the minority lifetime but is dependent on impurity profile, magnitude of stored charge and passive circuit parameters. As the impurity gradient at the junction is made steeper to increase retarding fields, the transition becomes faster. As the stored charge is increased, there is a tendency to swamp out the built-in fields, and the transition becomes slower.

The passive circuit parameters tend to slow down the fast transition due to the *RC* discharging time of the junction capacitance. An appropriate value of series inductance can improve the fast transition, but the passive contribution is approximately the *RC* time. Minimizing the *RC* time by decrease of area will, with a given stored charge, increase the contribution to the fast turn-off of the residual stored minority carriers. The converse is also true: minimizing the contribution of the residual stored carriers for a given stored charge by increase of area will increase the passive *RC* contribution to the fast turn-off. Thus, for a given application there is an optimum area which gives maximum turn-off rate at the end of the storage phase.

The observed behavior is in complete qualitative agreement with the theoretical predictions. Where predictions can be made, the actual switching times are within a factor of about two of the predicted times. One feature of the transient that is not predicted by the theory is the slow initial segment of the turn-off phase which occurs at the high forward current levels in Fig. 7. A minor part of this rounding may be ascribed to modulation of series resistance and the high junction capacitance at low voltage, but the major part of the rounding is not understood.

Appendix

Calculation of the Shape of Carrier Distribution at the Beginning of Transition Phase

Assumptions

1) Constant retarding field E.
2) Negligible recombination.
3) Stored charge Q_0 at $t=0$ when current density suddenly becomes equal to J_r.
4) Forward current density is Q_0/τ but we assume that the distribution of Q_0 is determined by field only.

The appropriate differential equations are

$$J_n = qD_n \frac{dn}{dx} + q\mu_n nE \tag{43}$$

$$\frac{dn}{dt} = \frac{1}{q}\frac{dJ_n}{dx}. \tag{44}$$

Combination of (43) and (44) gives

$$\frac{d^2N}{dx^2} + \frac{qE}{kT}\frac{dN}{dx} - \frac{SN}{D} = 0. \tag{45}$$

Where N is the Laplace transform of $m = n(x, t) - n(x, o)$ at $t = 0$, $m(x, o) = 0$,

$$\frac{J_r}{q_s} = D_n \frac{dm}{dx} + \mu m E, \tag{46}$$

$N(S, x)$

$$= \frac{J_r}{qDS} \frac{\exp - \left(\frac{qE}{2kT} + \sqrt{\left(\frac{qE}{2kT}\right)^2 + \frac{S}{D}}\right)x}{\frac{qE}{2kT} - \sqrt{\left(\frac{qE}{2kT}\right)^2 + \frac{S}{D}}}, \tag{47}$$

let $S' = S + (qE/2kT)^2 D$. Then

$$m(x, t) = \frac{J_r}{qD^2} \exp\left[-\left(\frac{qE}{2kT}\right)^2 Dt - \frac{qE}{2kT}\right]$$

$$\cdot \mathfrak{L}^{-1} \frac{-\epsilon^{-\sqrt{S'/D}\,x}}{\left\{\frac{S'}{D_n} - \left(\frac{qE}{2kT}\right)^2\right\}\left\{\sqrt{\frac{S'}{D_n}} - \frac{qE}{2kT}\right\}}. \tag{48}$$

Carrying out the inversion from standard tables and simplifying, we obtain

$$m(x, t) = -\frac{J_r}{qD}\left\{\left(\frac{kT}{2qE} - \frac{x}{2} + \frac{qE}{2kT} Dt\right) e^{-qEx/kT} \operatorname{erfc} Z\right.$$

$$+ \frac{\sqrt{D_n t}}{\pi} \epsilon^{-qEx/2kT} \epsilon^{-Z2}$$

$$\left. - \frac{kT}{2qE} \operatorname{erfc}\left(\frac{x}{2\sqrt{D_n t}} + \frac{qE}{2kT}\sqrt{D_n t}\right)\right\} \tag{49}$$

where

$$Z = \frac{x}{3\sqrt{D_n t}} - \frac{qE}{2kT}\sqrt{D_n t}.$$

At $t = t_0$, the density reaches zero at

$$x = 0 \quad \text{or} \quad m(0, t_0) = -p^*$$

and

$$n(x, t_0) = p^* \epsilon^{-qEx/kT} + m(x\ \ t_0)$$

$$p^* = \frac{J_r}{qD}\left\{\left(\frac{kT}{2qE} + \frac{qE}{2kT} D_n t_0\right) \operatorname{erfc}\left(-\frac{qE}{2kT}\sqrt{D_n t_0}\right)\right\}$$

$$+ \frac{\sqrt{D_n t_0}}{\pi} \epsilon^{-q^2E^2Dt_0/4kT} - \frac{kT}{qE} \operatorname{erfc} \frac{qE}{2kT}\sqrt{D_n t_0}. \tag{50}$$

Now if

$$\frac{qE}{2kT}\sqrt{D_n t_0} > 2,$$

almost all of the contribution to electron density on the right-hand side will come from

$$\operatorname{erfc} - \frac{qE}{2kT}\sqrt{Dt_0}. \tag{51}$$

Eq. (51) requires

$$t_0 > \frac{8}{D}\left(\frac{kT}{qE}\right)^2. \tag{52}$$

From (23)

$$t_0 \cong \tau \ln\left(1 + \frac{Q_0}{I_r \tau}\right) \text{ or if } t_0 \ll \tau,$$

$$t_0 \cong \frac{Q_0}{I_r}. \tag{53}$$

Thus the requirement (52) is that

$$I_r < \frac{D_n}{q}\left(\frac{qE}{kT}\right)^2 Q_0 \tag{54}$$

or

$$\frac{I_r}{I_f} < \frac{D_n \tau}{8}\left(\frac{qE}{kT}\right)^2.$$

The condition (54) is usually satisfied since it is

$$\frac{I_r}{I_f} < \frac{L^2}{8\left(\frac{kT}{qE}\right)^2} \approx 10 \text{ in a typical case.} \tag{55}$$

If (55) is not satisfied, we actually do not expect a very drastic modification of turn-off time. When (55) is satisfied,

$$n(x, t_0) \cong \frac{J_r x}{qD} \epsilon^{-qEx/kT} \tag{56}$$

since

$$\operatorname{erfc}\left(-\frac{qE}{2kT}\sqrt{Dt_0}\right) \cong 2.$$

If the field E is not a constant, it does not appear to be possible to solve the problem of finding the total charge in the residual bumps. However, the form (56) suggests that we might make an approximation such as

$$n(x, t_0) \cong \frac{J_r x}{qD} \exp\left\{-\frac{q}{kT}\int_0^x E\,dx\right\}. \tag{57}$$

If we use the linear approximation to potential as in (13) we obtain

$$\Delta Q_n = \int_0^\infty q_n dx = -\frac{J_r}{2D}\left(\frac{kT}{q}\right)^2 \frac{g(y)}{E_1E_2}, \qquad (58)$$

where

$$g(y) = \frac{1}{y} + y^{\frac{y}{1-y}}\left(y - \frac{1}{y} + \ln y\right)$$

and

$$T_q = \left(\frac{1}{D_p} + \frac{1}{D_n}\right)\frac{1}{2}\left(\frac{kT}{q}\right)^2\left(\frac{1}{E_1E_2}\right)g(y). \qquad (59)$$

Eq. (59) is the same as (29) except that it is modified by the factor $g(y)$. Fig. 13 is a plot of $g(y)$ vs y. In the particular example that was discussed in Section II, $E_1 = 50$ V/cm, $E_2 = 200$ V/cm and $g(y) = 0.75$.

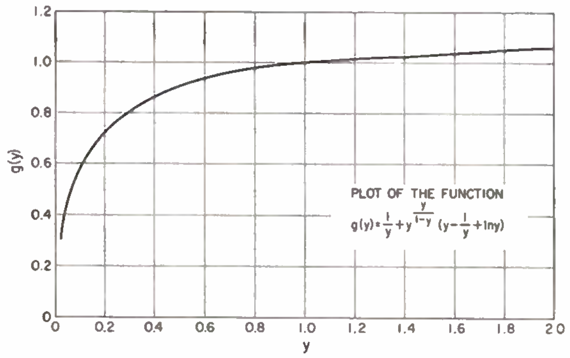

Fig. 13—Plot of $g(y)$ vs y.

ACKNOWLEDGMENT

The work reported here owes much to the skilled fabrication of these devices by H. Overacker and L. Madden.

Average Decay Laws for VLF Fields*

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***Summary*—A simplified discussion of propagation in the earth-ionosphere waveguide is given. It is shown that, if the square of field amplitude is averaged over the width of the waveguide, a very simple formula for the averaged intensity is obtained. This result is used to describe some of the broad features of VLF propagation in a relatively concise fashion.**

INTRODUCTION

THE representation of VLF field strength in terms of waveguide modes has been quite useful in explaining observed phenomena at great ranges.[1] Unfortunately, the convergence of the mode sum becomes very poor at short distances from the source. Recourse can be made to a ray or hop theory, but this is only convenient at the fairly short ranges. There is a distance range from about 400 to 4000 km where both ray theory and mode theory predict that the field strength is an undulating function of distance. The precise location of the maxima and minima in the interference patterns depends critically on the effective reflection height and to some extent on the electrical characteristics of the ionosphere and the ground. In many applications one is not too interested in the fine structure of the field strength vs distance, but rather one is more concerned with the average rate of decay of the field.

It is the purpose of this note to show that many of the broad characteristics of VLF fields, particularly in the frequency range from 10 to 25 kc, are compatible with a simple formula relating average field strength, distance, and frequency. The simplicity is achieved by making a number of reasonable approximations which can be justified on the basis of previous work[1] using more elaborate methods.[2]

FORMULATION

The space between the ground and the ionosphere is represented by a parallel plate waveguide of width h and whose effective surface impedances are Z_g and Z_i. The source is taken to be a vertical electric dipole located on the lower surface (*i.e.*, the ground). Following earlier work, the vertical field E at distance d and height

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[1] J. R. Wait, "The Mode Theory of VLF Ionospheric Propagation for Finite Ground Conductivity," presented at the VLF Symp., Boulder, Colo., January 23–25, 1957; also, PROC. IRE, vol. 45, pp. 760–767, June, 1957. (Additional references given here.)

[2] J. R. Wait, "Terrestrial propagation of very-low-frequency radio waves, a theoretical investigation," *J. Res. NBS*, vol. 64D, pp. 153–204; March/April, 1960.

z can be written[1,2]

$$E \cong E_0 W \tag{1}$$

where E_0 is the field of the dipole if the ground were flat and perfectly conducting, and if the ionosphere were not present (*i.e.*, $h \to \infty$).[3] The complex function W is thus a correction factor which contains the essence of the propagation problem. It is given by[1]

$$W = \frac{(d/\lambda)^{1/2}}{(h/\lambda)} e^{i[(2\pi d/\lambda)-(\pi/4)]} \cdot \sum_{n=0}^{\infty} \delta_n S_n{}^{3/2} e^{-i2\pi S_n d/\lambda} \cos(kC_n z) \tag{2}$$

where

$$\delta_n = \frac{1}{\left[1 + i\dfrac{\partial(R_i R_g)/\partial C}{4\pi(h/\lambda)R_i R_g}\right]_{C=C_n}} \tag{3}$$

$$R_i = \frac{C - \Delta_i}{C + \Delta_i}, \qquad \Delta_i = Z_i/\eta_0, \qquad \eta_0 = 120\pi$$

$$R_g = \frac{C - \Delta_g}{C + \Delta_g}, \qquad \Delta_g = Z_g/\eta_0,$$

C_n is a solution of

$$R_i R_g \exp[-i4\pi hC/\lambda] = \exp(-2\pi i n), \tag{4}$$

and finally $S_n = (1 - C_n{}^2)^{1/2}$.

Numerical values of C_n and S_n can be obtained from (4) using an iterative procedure. Extensive results[4] of this kind have been obtained by use of a digital computer and these have been used in evaluating the mode sum for W. Both a homogeneous and a stratified ionospheric model[2] have been considered in this way. More recent investigations[5] have dealt with the influence of earth curvature and the terrestrial magnetic field. These latter factors lead to a considerably more complicated mode equation. However, for present purposes, the parallel plate waveguide with homogeneous isotropic walls is adequate for this general discussion. Thus we may write

$$\Delta_g = G^{1/2} e^{i\pi/4}, \tag{5}$$

$$\Delta_i = \frac{\left[C^2 - \dfrac{i}{L}\right]^{1/2}}{1 - \dfrac{i}{L}} \tag{6}$$

where G and L are constants independent of C. In terms of the conductivity σ_g and dielectric constant ϵ_g of the ground,

$$G = \frac{\epsilon_0 \omega}{\sigma_g + i\epsilon_g \omega}.$$

In most cases $\epsilon_g \omega \ll \sigma_g$, so that

$$G \cong \frac{\epsilon_0 \omega}{\sigma_g}$$

is regarded as a real quantity in what follows. In terms of the (angular) plasma frequency ω_0 and the collisional frequency ν of the ionosphere,

$$L = \frac{\omega}{\omega_r} \quad \text{where} \quad \omega_r = \frac{\omega_0{}^2}{\nu}. \tag{7}$$

The square of the amplitude of the field in the waveguide can now be written

$$|E^2| = |E_0{}^2|\,|W^2| \tag{8}$$

where

$$|W^2| = WW^*$$

$$= \frac{d/\lambda}{(h/\lambda)^2} \sum_{n=0}^{\infty} \sum_{m=0}^{\infty} \delta_n \delta_m{}^* S_n{}^{3/2} (S_m{}^*)^{3/2} e^{-ik(S_n - S_m{}^*)d} \times \cos(kC_n z) \cos(kC_m{}^* z). \tag{9}$$

The double summation is not particularly useful in this form. However, the average value of $|W^2|$ can be represented in a very simple form if approximations are made which are valid for oblique incidence modes. A similar approach has been used by Brekhovskikh[6] in connection with acoustic propagation in the ocean. Specifically, if

$$C_n/\Delta_i \ll 1 \quad \text{and} \quad C_n/\Delta_g \gg 1$$

for all important modes, it follows that[1]

$$C_n \cong \frac{\pi(n - \frac{1}{2}) + e^{i3\pi/4} G^{1/2} (\overline{C}_n)^{-1}}{kh - i\left(1 - \dfrac{i}{L}\right)\left[(\overline{C}_n)^2 - \dfrac{i}{L}\right]^{-1/2}} \tag{10}$$

where

$$\overline{C}_n = \frac{\pi(n - \frac{1}{2})}{kh} = \frac{(n - \frac{1}{2})}{(2h/\lambda)}. \tag{11}$$

The restrictions stated above may now be replaced by

$$(\overline{C}_n)^2 \ll \frac{1}{L}, \; L \quad \text{and} \quad (\overline{C}_n)^2 \gg G.$$

Since L is of the order of 0.5 and $G \ll 1$ in the frequency range of 10 to 25 kc, these conditions are well satisfied for the important modes.

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[3] E_0 can be loosely described as the *inverse-distance field*.

[4] H. H. Howe and J. R. Wait, "Mode Calculations for VLF Ionospheric Propagation," presented at VLF Symp., Boulder, Colo.; January 23–25; 1957.

[5] 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. [See also (same authors) NBS Tech. Note 114, July, 1961, for extensive calculations.]

[6] L. M. Brekhovskikh, "Sound propagation in an underwater sound channel," *Dokl. Akad. Nauk. SSSR*, vol. 69, pp. 157–167; February, 1949.

The double summation for $|W^2|$ may thus be approximated by

$$|W^2| \cong \frac{d/\lambda}{(h/\lambda)^2} \sum_{n=1}^{\infty} \sum_{m=1}^{\infty} e^{-ik(S_n - S_m^*)d} \cdot \cos\left[\frac{\pi(n-\frac{1}{2})z}{h}\right] \cos\left[\frac{\pi(m-\frac{1}{2})z}{h}\right] \tag{12}$$

where δ_n and δ_m have been replaced by unity. In this case, the modes of lowest attenuation are of order one following earlier adopted conventions.[2]

The Averaging Process

We now consider the value of the intensity (or field amplitude)[2] averaged over the width of the waveguide. This average is obtained from $|\overline{E^2}| = |E_0{}^2|\,|\overline{W^2}|$ where

$$|\overline{W^2}| = \frac{1}{h}\int_0^h |W^2|\,dh. \tag{13}$$

In view of the relation

$$\frac{2}{h}\int_0^h \cos\left[\frac{\pi(n-\frac{1}{2})z}{h}\right] \cos\left[\frac{\pi(m-\frac{1}{2})z}{h}\right] dz = \begin{cases} 1 \text{ for } n = m \\ 0 \text{ for } n \neq m \end{cases} \tag{14}$$

it readily follows that

$$|\overline{W^2}| \cong \frac{(d/\lambda)}{2(h/\lambda)^2} \sum_{n=1}^{\infty} e^{-ik(S_n - S_n^*)d}. \tag{15}$$

It may be noted that

$$S_n - S_n^* = 2i \,\mathrm{Im}\cdot S_n \tag{16}$$

and from (10),

$$\mathrm{Im}\cdot S_n \cong -\frac{\lambda}{2\sqrt{2}\pi h}\left[(\overline{C}_n)^2\left(\sqrt{L} + \frac{1}{\sqrt{L}}\right) + \sqrt{G}\right]\cdot \tag{17}$$

Thus

$$|\overline{W^2}| \cong \frac{(d/\lambda)P_g}{2(h/\lambda)^2} \sum_{n=1}^{\infty} e^{-\alpha(n-1/2)^2 d} \tag{18}$$

where

$$\alpha = \frac{\lambda^2}{2\sqrt{2h^3}}\left(\sqrt{L} + \frac{1}{\sqrt{L}}\right) \tag{19}$$

and

$$P_g = \exp[-\sqrt{2G}(d/h)]. \tag{20}$$

The summation indicated above can be written explicitly in terms of the Jacobi theta function $\theta_2(\nu, \kappa)$ as defined by Jahnke, Emde, and Lösch.[7] Thus

$$|\overline{W^2}| \cong \frac{(d/\lambda)P_g}{2(h/\lambda)^2}\,\theta_2\left(0, \frac{\alpha d}{\pi}\right)\cdot \tag{21}$$

When αd is small, a fairly large number of terms in the summation in (18) are required; for a first approximation, the sum may be replaced by an integral. Thus

$$|\overline{W^2}| \cong \frac{(d/\lambda)P_g}{2(h/\lambda)^2}\int_0^\infty e^{-\alpha d g^2}dg = \frac{(d/\lambda)P_g}{4(h/\lambda)^2}\left(\frac{\pi}{\alpha d}\right)^{1/2}\cdot \tag{22}$$

This is valid for $\alpha d < 2$ (see Appendix). Making use of (19), this can be written in the compact form

$$|\overline{W^2}| \cong \left(\frac{d}{1.80\,h}\right)^{1/2}\left(\frac{1}{\sqrt{L} + \dfrac{1}{\sqrt{L}}}\right)^{1/2} \cdot \exp[-\sqrt{2G}\,d/h]. \tag{23}$$

Since

$$|\overline{E^2}| = |E_0|^2\,|\overline{W^2}|$$

and

$$|E_0{}^2| = \frac{\text{const.}}{d^2}$$

is it seen that distance dependence is described by

$$|\overline{E^2}| \propto \frac{1}{d^{3/2}}\exp[-\sqrt{2G}\,d/h]$$

$$\propto \frac{1}{d^{3/2}} \text{ for } G = 0 \ (i.e.\ \sigma_g \cong \infty). \tag{24}$$

To test the validity of this simple law, $|\overline{W^2}|$ given by (23) and $|W^2|$ given by (9) are plotted as a function of d/λ for the conditions, $h/\lambda = 5$, $L = 0.5$, $G = 0$. It is seen that over a range of d/λ from about 20 to 200, the straight line, which corresponds to $|\overline{W^2}|$, is a reasonable average of the undulating curve for $|W^2|$. It may also be demonstrated[1] that if h/λ is changed by 10 per cent or so, the computed mode sum for $|W^2|$ has maxima and minima that are shifted appreciably from the curve shown in Fig. 1. However, the same straight line is still a good measure of the averaged intensity for the distance range from 20 to 200 wavelengths.

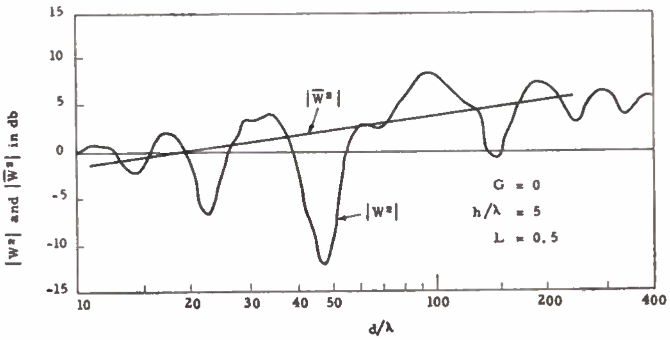

Fig. 1—A comparison of the intensity and the averaged intensity for a concrete case.

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[7] E. Jahnke, *et al.*, "Tables of Higher Functions," McGraw-Hill Book Co., Inc., New York, N. Y.; 1960.

When $\alpha d \gg 1$, the summation in (18) reduces essentially to one term. Thus,

$$|\overline{W^2}| \cong \frac{d/\lambda}{2(h/\lambda)^2} e^{-\alpha d/4} e^{-\sqrt{2G}(d/h)}. \qquad (25)$$

The average intensity is now exponentially damped. This corresponds to the region where only one waveguide mode is predominant. This particular case has been discussed at length in the literature.[1,8] It should be mentioned, however, that earth curvature increases the effective value of the attenuation of the dominant mode.[5] However, the flat-earth approximation is justified for the higher modes and thus averaged intensity in the intermediate distance range is not appreciably influenced by earth curvature.

CONCLUDING REMARKS

Eq. (23) for $|\overline{W^2}|$ allows for a very simple physical discussion of the averaged field intensity in the intermediate distance region. In the first place, it can be seen that $|\overline{E^2}|$ varies as $d^{-3/2}$ when the ground losses are negligible (such as over sea water). The ground conductivity results in additional attenuation per unit distance given by

$$(6.14/h)\sqrt{G} \times 10^3 \text{ db per 1000 km}$$

when h is expressed in kilometers.

For example, if $h = 80$ km, $f = 16$ kc, and $\sigma_g = 2$ millimhos/m, this amounts to 1.6 db/1000 km. This agrees quite well with the results obtained from a more rigorous numerical treatment of the modal equation.[1,4] It is interesting to note that this attenuation factor is proportional to $f^{1/2}$ and $\sigma_g^{-1/2}$. Again, it should be stressed that this is only true for the intermediate distance region. At very great distances (*i.e.*, $d > 3000$ km), the influence of earth curvature on the dominant mode leads to a more complicated dependence of ground losses on frequency and conductivity.[5]

Since the averaged field intensity is proportional to the factor

$$\left[\sqrt{L} + \frac{1}{\sqrt{L}}\right]^{-1/2} = \left[\sqrt{\frac{\omega}{\omega_r}} + \sqrt{\frac{\omega_r}{\omega}}\right]^{-1/2}$$

we see that a broad maximum occurs for a frequency $\omega = \omega_r$. Since ω_r is of the order of 10^5, this corresponds to a frequency of the order of 15 kc. Such a maximum is observed when the full mode sum is plotted as a function of frequency. Also, such a maximum has been deduced by Watt and Maxwell[9] on the basis of experimental data. It should be mentioned that the frequency dependence is a function of the ionospheric model. For a stratified or a *diffuse* model, the factor $\sqrt{L}+1/\sqrt{L}$ is replaced by a more complicated function. This is also true when the earth's magnetic field is included in the analysis.[5]

APPENDIX

From (8) and (21), it is seen that the averaged field intensity behaves in the manner

$$|\overline{E^2}| \propto \Omega$$

where

$$\Omega = \frac{1}{\alpha d}\theta_2\left(0, \frac{\alpha d}{\pi}\right)$$

$$= \frac{2}{\alpha d}\sum_{n=1}^{\infty} e^{-(n-1/2)^2 \alpha d}.$$

The function Ω is plotted in Fig. 2 as a function of αd. On the same graph, the integral approximation

$$\frac{2}{\alpha d}\int_0^\infty e^{-\alpha d g^2} dg = \frac{\pi^{1/2}}{(\alpha d)^{3/2}} = \Omega_0$$

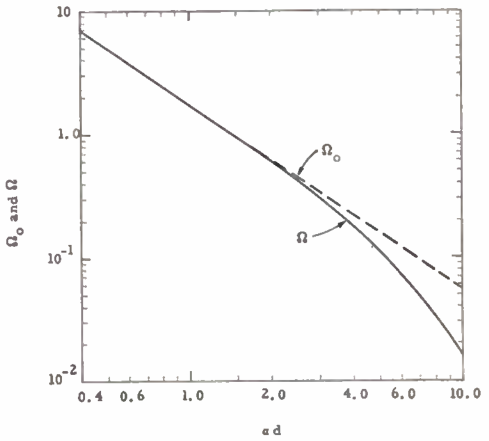

Fig. 2—Averaged decay laws.

is also shown for comparison. As can be seen, if $\alpha d < 2$, the summation is well approximated by the 3/2's power law. In the example shown in Fig. 1, this corresponds to $d/\lambda < 300$.

ACKNOWLEDGMENT

I would like to thank Mrs. A. M. Conda for computing the Theta function, and Mrs. E. Brackett for assistance in preparing the paper.

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[8] K. G. Budden, "The propagation of a radio atmospheric II," *Phil. Mag.*, vol. 43, pp. 1179–1200; November, 1952.

[9] A. D. Watt and E. L. Maxwell, "Characteristics of atmospheric noise from 1 to 100 kc/s," PROC. IRE, vol. 45, pp. 787–795; June, 1957.

Theory, Design and Performance of Maximum-Efficiency Variable-Reactance Frequency Multipliers*

T. UTSUNOMIYA†, MEMBER, IRE, AND S. YUAN‡

Summary—**Although the Manley and Rowe relations indicate that complete conversion of fundamental power to any specific harmonic with a nonlinear-reactance element is not impossible theoretically, many factors contribute to lowering the efficiency of real multipliers. Analysis, based on a Taylor series expansion of the nonlinear-reactance characteristic and Fourier expansion of the voltage and charge, with the introduction of suitable parameters and quiescent conditions, permits the exact determination of the various effects. Given the characteristics of a nonlinear element and the inevitable circuit losses, maximum obtainable efficiency can be predicted from the analysis. Significant increases in efficiency are obtained in various circuit configurations where selected intermediate harmonic currents are permitted to flow and where harmonics which cannot contribute to useful output at the desired harmonic frequency are suppressed. The characteristic of nonlinear reactance is shown to have a calculable effect on the multiplier efficiency, so that the most desirable characteristic can be specified.**

Experimental nonlinear capacitance multipliers designed according to the developed theory have provided an efficiency of 50 per cent for a quintupler and 70 per cent for a tripler, very close to the predicted values.

I. INTRODUCTION

FREQUENCY multipliers of the conventional vacuum-tube type suffer from noise generated in the interstage tubes, and low-noise requirements are growing severely with the developments of high-quality FM systems in various fields. Since variable-reactance devices, primarily as amplifiers, have proved their superiority especially as to low-noise characteristics, nonlinear-reactance frequency multipliers become important for realizing low-noise frequency multiplication. The high efficiency obtainable with these nonlinear-reactance multipliers provides the possibility of using them as multipliers of considerably high order, without introducing additional or excess noise from active elements.

In the developmental stage of such devices, various characteristics, such as noise characteristics, signal-handling capabilities, frequency limits, and the system design for such multiplier chains, should be investigated thoroughly. Above all, efficiency optimization is certainly of primary interest in any of the applications. This has led to an investigation of such devices to determine the maximum obtainable efficiency.

Although complete conversion of fundamental power to power at any specific harmonic is not impossible with an ideal nonlinear-reactance element,[1] many factors contribute to lowering the efficiency of real multipliers. Leeson and Weinreb[2] gave a detailed analysis which provided a basis for design and optimization of such frequency multipliers. Their results showed much improvement in efficiency as compared with that of a frequency multiplier employing an ideal rectifier (nonlinear resistance),[3] but only for low-order multiplication ($N \leq 5$). Obviously, high-order multipliers with high efficiency can be achieved by cascading highly efficient doublers and triplers, but with some inconvenience in the practical design of such multipliers. The purpose of this investigation then, was to improve the performance of higher-order frequency multipliers, *i.e.*, $N>2$, employing a fewer number of nonlinear-reactance elements.

A significant increase in efficiency is obtained in various circuit configurations where undesired harmonics are prevented from being generated or their power is returned to useful output at the desired harmonic frequency. A design theory has been developed showing the exact effect of various losses on efficiency, so that, given the characteristics of a nonlinear element and the inevitable circuit losses, maximum obtainable efficiency can be predicted. Performance characteristics are derived as functions of circuit variables. Measured efficiencies agree closely with the calculated efficiencies for a frequency doubler, tripler, quadrupler and quintupler. For each order of multiplication, a particular circuit was designed to maximize the efficiency by minimizing overall loss due to circuit elements, stray capacitances, undesired harmonics and so forth. Experimental nonlinear-capacitance multipliers designed according to the developed theory have provided an efficiency of 50 per cent for a quintupler and 70 per cent for a tripler.

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‡ Electronics Research Laboratories, School of Engrg., Columbia, University, New York, N. Y.

[1] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements—pt. I, general energy relations," PROC. IRE, vol. 44, pp. 904–913; July, 1956.

[2] D. B. Leeson and S. Weinreb, "Frequency multiplication with nonlinear capacitors—a circuit analysis," PROC. IRE, vol. 47, pp. 2076–2084; December, 1959.

[3] C. H. Page, "Harmonic generation with ideal rectifiers," PROC. IRE, vol. 46, pp. 1738–1740; October, 1958.

II. ANALYSIS

Although the following analysis is limited to using a variable capacitance as the nonlinear element, a similar analysis may also be carried out for the nonlinear-inductance multiplier. The variable-capacitance case is chosen because of the availability of variable capacitors and the simplicity in carrying out the experiment. The general case of a nonlinear capacitance and its external circuits series-tuned to a definite number of harmonic frequencies will be considered first for the analysis. Due to the complicated mathematical manipulation, only the general procedure of solving the problem will be outlined. A specific example will be given later to illustrate and clarify the procedure.

A. General Procedure

For frequency multipliers using a variable capacitance, the voltage-charge relationships of the nonlinear capacitor can be expressed in either of two general functional forms:

$$Q = Q(V) \tag{1}$$

or

$$V = V(Q). \tag{2}$$

The charge Q can be expressed

$$Q = Q_0(1 + q) = Q_c$$
$$= \text{charge across the variable capacitance} \tag{3}$$

where Q_0 is the quiescent operating point and q is the normalized variable component of the charge. Similarly, the voltage is expressed

$$V = V_0(1 + v) = V_c$$
$$= \text{voltage across the variable capacitance} \tag{4}$$

where V_0 is the quiescent operating point, and v is the normalized variable component of the voltage. It should be noted that the variable voltage components may also include a dc component due to the nonlinearity under the presence of ac charges. $Q(V)$ and $V(Q)$ may be approximated in the small signal case by Taylor series expansions about the operating point. As many terms in the series as are necessary may be included in the larger signal case. Thus,

$$Q(V) = Q_0(1 + \alpha_1 v + \alpha_2 v^2 + \cdots + \alpha_k v^k + \cdots) \tag{5}$$

and

$$V(Q) = V_0(1 + \beta_1 q + \beta_2 q^2 + \cdots + \beta_k q^k + \cdots) \tag{6}$$

where

$$\alpha_k = \frac{V_0^k}{Q_0 k!} \left. \frac{d^k Q}{dV^k} \right|_{V=V_0} \tag{7}$$

$$\beta_k = \frac{Q_0^k}{V_0 k!} \left. \frac{d^k V}{dQ^k} \right|_{Q=Q_0} \tag{8}$$

Comparison of (3) and (5) shows

$$q = \sum_{k=1}^{\infty} \alpha_k v^k \tag{9}$$

and (4) and (6) show

$$v = \sum_{k=1}^{\infty} \beta_k q^k. \tag{10}$$

Eqs. (9) and (10) are the general relations of normalized variable charge and voltage in the nonlinear element. Eq. (10) can be used for solving for the voltage in terms of charge; a dual analysis employs (9) to solve for the charge in terms of voltage.

The normalized variables are now expanded into complex Fourier series.

$$q = \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} \frac{q_n}{2} e^{j(n\omega t)} \tag{11}$$

$$i = \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} jn\omega \frac{q_n}{2} e^{jn\omega t} \tag{12}$$

$$v = \sum_{m=-\infty}^{\infty} \frac{v_m}{2} e^{jm\omega t} \tag{13}$$

where ω is the fundamental driving radian frequency. In these series $v_m = v_{-m}^* = |v_m| e^{j\phi_m}$ and $q_n = q_{-n}^* = |q_n| e^{j\phi_n}$. Substituting (11), and (13) into (10) yields

$$\sum_{m=-\infty}^{\infty} \frac{v_m}{2} e^{jm\omega t} = \sum_{k=1}^{\infty} \beta_k \left[\sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} \frac{q_n}{2} e^{j(n\omega t)} \right]^k. \tag{14}$$

For practical reasons to be justified later, (14) may be simplified to include only terms with $k \leq 2$. With this approximation, (14) is reduced to

$$\sum_{\substack{m=-\infty \\ m\neq 0}}^{\infty} \frac{v_m}{2} e^{jm\omega t}$$
$$= \beta_1 \left[\sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} \frac{q_n}{2} e^{j(n\omega t)} \right] + \beta_2 \left[\sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} \frac{q_n}{2} e^{jn\omega t} \right]^2 \tag{15}$$

where the existence of a charge component at any frequency is determined by the finite impedance properties of the external circuit at the corresponding harmonic frequency. Given a set of preselected desirable charge components, an equation is obtained for each harmonic by extracting the appropriate coefficients from (14). The simplified form of (15) thus provides the circuit equation relating the voltage for each frequency in terms of the preselected charge components. Then

$$\frac{v_1}{2} = \beta_1 \frac{q_1}{2} + \beta_2 \left[2 \frac{q_{-1} q_2}{4} + 2 \frac{q_{-2} q_3}{4} + \cdots \right.$$
$$\left. + 2 \frac{q_{-(n-1)} q_n}{4} + \cdots \right] \tag{16}$$

$$\frac{v_2}{2} = \beta_1 \frac{q_2}{2} + \beta_2 \left[\left(\frac{q_1}{2} \right)^2 + 2 \frac{q_{-1} q_3}{4} + \cdots + 2 \frac{q_{-(n-2)} q_n}{4} + \cdots \right] \qquad (17)$$

or more generally, for $m > 0$,

$$\frac{v_m}{2} = \beta_1 \frac{q_m}{2} + \beta_2 \left[\frac{1}{2} \sum_{n=1}^{(m-1)/2} q_n q_{m-n} + \frac{1}{2} \sum_{n=m+1}^{\infty} q_{-(n-m)} q_n \right] \quad m = \text{odd}$$

$$= \beta_1 \frac{q_m}{2} + \beta_2 \left[\left(\frac{q_{m/2}}{2} \right)^2 + \frac{1}{2} \sum_{n=1}^{m/2-1} q_n q_{m-n} + \frac{1}{2} \sum_{n=m+1}^{\infty} q_{-(n-m)} q_n \right] \quad m = \text{even} \qquad (18)$$

and, v_{-m} is just the complex conjugate of v_m. The terms in (18), together with the external circuit, provide the simultaneous circuit equations, and the basis for establishing an equivalent circuit for each frequency of interest. Eq. (18) will be greatly simplified when only a finite number of preselected charge components are allowed to exist in the nonlinear diode. The general circuit configuration for the nonlinear-capacitance frequency multiplier is then as shown in Fig. 1. The phase angle relations at these frequencies depend on the tuning conditions. The analysis of efficiency optimization can best be illustrated with a specific example since, in the general case, the notation becomes prohibitively complicated (see Section II-C).

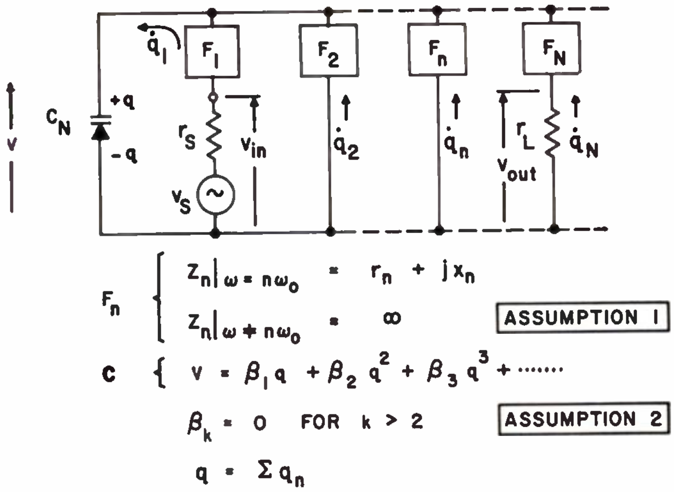


Fig. 1—General circuit for frequency multiplier with nonlinear capacitance.

E. *Specific Example*

The performance of a frequency quintupler employing a semiconductor nonlinear-capacitance diode will now be discussed in detail. First, the choice of the harmonics preferred to appear across the diode will be investigated.

1) 1–5 mode: This mode is the same as the scheme used by Leeson and Weinreb,[2] where currents of only the fundamental frequency and its fifth harmonic flow through the diode.

2) 1-4-5 mode: Employing one intermediate harmonic, the fourth, the desired harmonic output is obtained by the mixing of the fundamental frequency and the fourth harmonic, which is converted directly from the fundamental frequency.

3) 1-2-3-5 mode: This mode allows currents of the second and third harmonics to flow through the diode in addition to the fundamental frequency current and the desired harmonic output.

4) 1-2-4-5 mode: Currents of the second and fourth harmonics flow through the diode, in addition to the fundamental frequency and the desired harmonic output currents.

5) 1-2-3-4-5 mode: Currents of all harmonics below the desired output are allowed to flow.

6) 1-2-3-4-5 · · · n: Currents of all harmonics below the desired output flow, plus other harmonics higher than the desired output harmonic.

Case 1) was analyzed in detail by Leeson and Weinreb[2] and the mathematical expression for efficiency was

$$\frac{P_N}{P_{in}} = \frac{1}{\left(1 + \frac{R_a}{R_1}\right)\left(1 + \frac{R_b}{R_N}\right)}, \qquad (19)$$

where

R_1 = input resistance,
R_a = circuit loss of input circuit,
R_b = circuit loss of output circuit,
R_N = load resistance,
P_{in} = power input at fundamental frequency,
P_N = power output at Nth order harmonic.

R_1 is shown to be

$$R_1 = \frac{1}{\omega C_0} \frac{(1 - \Gamma)[(1 - \Gamma)\beta_N^2]^{1/(2N-1)}}{2^{(2N-2)/(2N-1)}} \cdot \frac{V_{in}}{V_0}^{(2N-2)/(2N-1)} \frac{1}{\omega C_0 R_N}^{1/(2N-1)} \qquad (20)$$

where

$$\beta_k = \frac{1}{k!} \left(\frac{1}{1 - \Gamma} \right) \left(\frac{1}{1 - \Gamma} - 1 \right) \left(\frac{1}{1 - \Gamma} - 2 \right) \cdots \cdot \left(\frac{1}{1 - \Gamma} - k + 1 \right). \qquad (21)$$

Γ defines the nonlinearity of the variable capacitance as expressed by

$$C(V) = \frac{C_1}{(V + \phi)^\Gamma} \qquad (22)$$

with

V = terminal voltage,
ϕ = contact potential of the junction,
C_1 = value of capacitance for $(V+\phi)=1$ volt,
$0<\Gamma<1$.

Eq. (20) shows that R_1 decreases rapidly as N increases, thus decreasing the efficiency. It is noted here that this is true only for those nonlinear capacitances that satisfy (22). By the same reasoning, case 2) should have better efficiency than case 1) but will still be less efficient than cases 3) and 4). Cases 3) and 4) require similar analysis and it has been determined both theoretically and experimentally that, of the two, case 4) gives a better efficiency. Cases 5) and 6) require lengthy mathematical manipulation.

For the purpose of demonstrating the efficiency calculation, the 1-2-4-5 mode will be discussed in detail. The conditions and approximations imposed on the analysis will be the subject of a slight digression.

1) For most conventional variable-capacitance diodes, Γ ranges from $\frac{1}{3}$ to $\frac{1}{2}$. For $\Gamma=\frac{1}{2}$, (21) indicates that $\beta_k=0$ for $k>2$. For $\Gamma\leq\frac{1}{2}$, it can be shown that an error in the calculated efficiency of less than 5 per cent will be introduced if β_k's are assumed to be zero for $k>2$. Moreover, the conditions for optimum efficiency require Γ to be as large as possible, as will be shown later. Therefore, only β_1 and β_2 are included in the calculation, and β_k's for $k>2$ are assumed to be zero. This is the justification for the use of (15) in place of (14).

2) Since the only frequencies of interest are the first, second, fourth and fifth harmonics, the other frequency component currents are assumed to be nonexistent in the circuit.

3) Loss resistance in each harmonic circuit is assumed to be small compared with the total conversion resistance in the respective harmonic circuit.

4) Perfect tuning at each frequency is assumed.

5) The diode is prevented from conducting in the forward region, or in the region of reverse breakdown.

6) All terms in the analysis are normalized.

Now, the general procedure outlined above will be particularized. First, the normalizing factors are as follows:

V_0: the bias voltage, for voltages,
Q_0: the charge on the nonlinear capacitor at V_0, for charges,
ωQ_0: for currents,
$V_0/\omega Q_0$: for impedances,
Q_0/V_0: for capacitances.

For the 1-2-4-5 mode, (11) becomes

$$q=\frac{q_1}{2}e^{j\omega t}+\frac{q_{-1}}{2}e^{-j\omega t}+\frac{q_2}{2}e^{j2\omega t}+\frac{q_{-2}}{2}e^{-j2\omega t}+\frac{q_4}{2}e^{j4\omega t}$$
$$+\frac{q_{-4}}{2}e^{-j4\omega t}+\frac{q_5}{2}e^{j5\omega t}+\frac{q_{-5}}{2}e^{-j5\omega t}. \tag{23}$$

Eq. (18) produces for the frequencies of interest, namely, the fundamental, the second harmonic, the fourth harmonic and the fifth harmonic the relations

$$v_1=\beta_1|q_1|e^{j\theta_1}$$
$$+\beta_2\left[|q_1|\,|q_2|\,e^{j(\theta_2-\theta_1)}+|q_4|\,|q_5|\,e^{j(\theta_5-\theta_4)}\right] \tag{24}$$
$$v_2=\beta_1|q_2|e^{j\theta_2}$$
$$+\beta_2\left[\frac{|q_1|^2}{2}e^{j2\theta_1}+|q_2|\,|q_4|\,e^{j(\theta_4-\theta_2)}\right] \tag{25}$$
$$v_4=\beta_1|q_4|e^{j\theta_4}$$
$$+\beta_2\left[\frac{|q_2|^2}{2}e^{j2\theta_2}+|q_1|\,|q_5|\,e^{j(\theta_5-\theta_1)}\right] \tag{26}$$
$$v_5=\beta_1|q_5|\,e^{j\theta_5}+\beta_2|q_1|\,|q_4|\,e^{j(\theta_1+\theta_4)}. \tag{27}$$

The parameters of the external circuits are now introduced. By the simple application of Kirchhoff's laws to Figs. 2–5, (24)–(27) result in the following circuit equations. For convenience in notation, we define

$$q_n=|q_n|, \tag{28}$$

$$v_1=V_{\text{in}}e^{j\phi_{\text{in}}}-q_1(r_1+jx_1)e^{j[\theta_1+(\pi/2)]}$$
$$=\beta_1q_1e^{j\theta_1}+\beta_2\left[q_1q_2e^{j(\theta_2-\theta_1)}+q_4q_5e^{j(\theta_5-\theta_4)}\right], \tag{29}$$
$$v_2=-2q_2(r_2+jx_2)e^{j[\theta_2+(\pi/2)]}$$
$$=\beta_1q_2e^{j\theta_2}+\beta_2\left[\frac{q_1^2}{2}e^{j2\theta_1}+q_2q_4e^{j(\theta_4-\theta_2)}\right], \tag{30}$$
$$v_4=-4q_4(r_4+jx_4)e^{j[\theta_4+(\pi/2)]}$$
$$=\beta_1q_4e^{j\theta_4}+\beta_2\left[\frac{q_2^2}{2}e^{j2\theta_2}+q_1q_5e^{j(\theta_5-\theta_1)}\right], \tag{31}$$
$$v_5=-5q_5(r_L+r_5+jx_5)e^{j[\theta_5+(\pi/2)]}$$
$$=\beta_1q_5e^{j\theta_5}+\beta_2\left[q_1q_4e^{j(\theta_1+\theta_4)}\right]. \tag{32}$$

Equivalent circuits for this device at ω, 2ω, 4ω and 5ω are, respectively, Figs. 2, 3, 4, and 5. A close examination of the equivalent circuits leads to the following conclusions:

1) Both multiplying and mixing processes are involved in such devices, that is, the fundamental frequency is multiplied to the second harmonic and then to the fourth harmonic, which is mixed with the fundamental to give the fifth harmonic, the desired output frequency.

2) Currents of various harmonic frequencies are closely related because of the power-transferring mechanism. They are also functions of the input voltage, load resistance and the losses of circuit elements.

3) The over-all efficiency will be increased by maximizing the efficiency of individual loops, since each loop contributes to the output power. The over-all efficiency, however, is not the product of the efficiencies of the individual loops because of the mixing process.

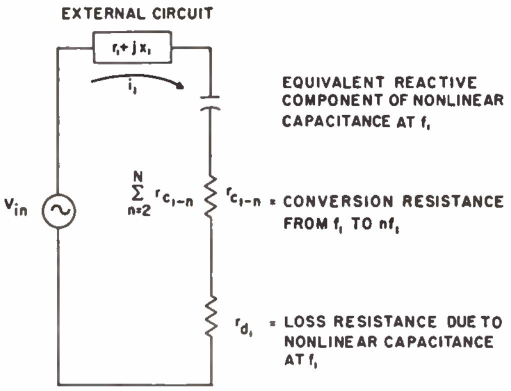


Fig. 2—Equivalent circuit at fundamental frequency f_1.

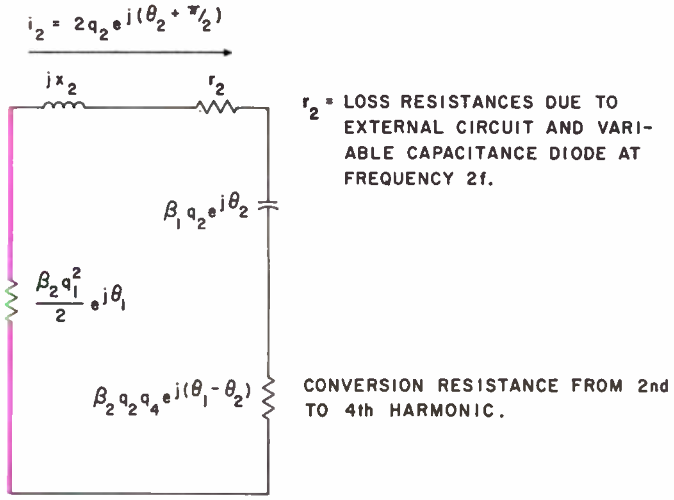


Fig. 3—Equivalent circuit of the 1-2-4-5 mode frequency quintupler at frequency $2f$.

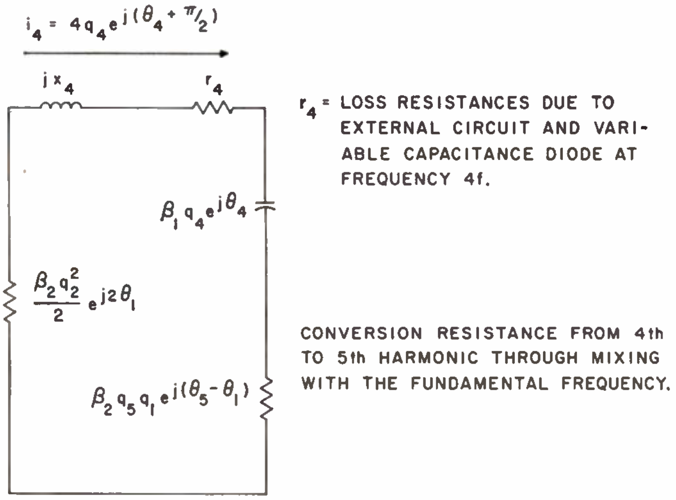


Fig. 4—Equivalent circuit of the 1-2-4-5 mode frequency quintupler at frequency $4f$.

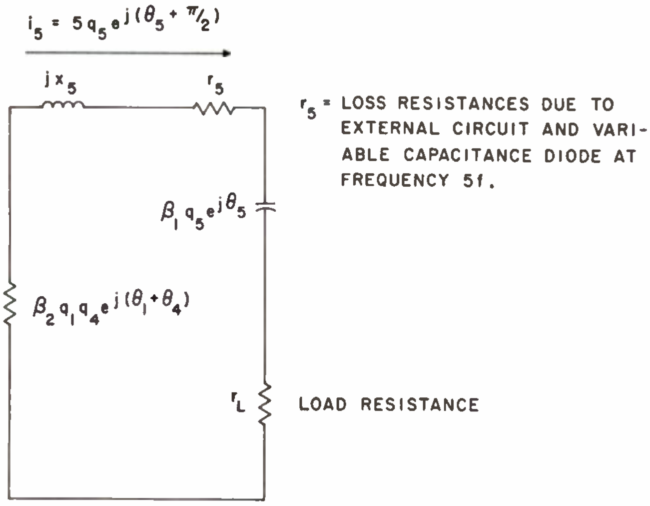


Fig. 5—Equivalent circuit of the 1-2-4-5 mode frequency quintupler at frequency $5f$, the output frequency.

C. Efficiency Calculation

There are many ways that can be used to determine efficiency. The easiest and most direct way is to calculate the input power and output power by means of input current, output current, input resistance and output resistance.

A more general approach is to express the over-all efficiency in terms of the efficiencies of the individual loops. For a frequency quintupler using the 1-2-4-5 mode, a flow graph of power can be drawn, as in Fig. 6, where α denotes the amount of fundamental power being converted to the second harmonic and $(1-\alpha)$ denotes the rest of the fundamental power being mixed with the fourth harmonic to give the fifth harmonic. Associated with each frequency is the efficiency of the individual loop. From Fig. 6, the output power is

$$P_0 = P_{\text{in}}[\alpha\eta_1\eta_2\eta_4 + (1-\alpha)\eta_1]\eta_5, \tag{33}$$

where

$P_{\text{in}}(1-\alpha)\eta_1 = P_{15}$ = Power converted from fundamental frequency to output frequency due to mixing with power at fourth harmonic,

$P_{\text{in}}(\alpha\eta_1\eta_2\eta_4) = P_{45}$ = Power converted from fourth harmonic to output frequency due to mixing with power at the fundamental frequency.

From (24) and (26), the relations $P_{15} = \beta_2 q_4 q_5 q_1$, $P_{45} = 4\beta_2 q_4 q_5 q_1$ are obtained. P_{15} and P_{45} are related by the Manley and Rowe[1] relations, $4P_{15} = P_{45}$, from which

$$P_{\text{in}}\alpha\eta_1\eta_2\eta_4 = 4(1-\alpha)\eta_1 P_{\text{in}} \tag{34}$$

and

$$\alpha = \frac{4\eta_1}{\eta_1\eta_2\eta_4 + 4\eta_1}. \tag{35}$$

Substitution of (35) into (33) gives

$$\frac{P_0}{P_{\text{in}}} = \eta = \frac{5\eta_1\eta_2\eta_4\eta_5}{4 + \eta_2\eta_4}. \tag{36}$$

This is the relation for over-all efficiency in terms of the efficiencies of each circuit. These latter are difficult to compute, so that the more direct calculation of the ratio of output to input powers will be undertaken.

The calculation of over-all efficiency can be greatly simplified if perfect tuning conditions are assumed. The phase angle relations for perfect tuning at each frequency derived from (29)–(32) are

$$2\theta_1 - \theta_2 + \frac{\pi}{2} = 0 \tag{37}$$

$$2\theta_2 - \theta_4 + \frac{\pi}{2} = 0 \tag{38}$$

$$\theta_1 + \theta_4 - \theta_5 + \frac{\pi}{2} = 0. \tag{39}$$

An over-all simplified equivalent circuit, corresponding to Fig. 6, is presented in Fig. 7. Eqs. (29)–(32), together with (37)–(39), give

$$\frac{V_{\text{in}}}{\beta_2 q_1} = \frac{r_{\text{in}}}{\beta_2} = \frac{r_1}{\beta_2} + q_2 + \frac{q_5 q_4}{q_1}, \tag{40}$$

$$\frac{r_2}{\beta_2} + \frac{q_4}{2} = q_1^2/4q_2, \tag{41}$$

$$\frac{r_4}{\beta_2} + \frac{q_5 q_1}{4q_4} = q_2^2/8q_4, \tag{42}$$

$$(r_5 + r_L)/\beta_2 = \frac{q_1 q_4}{5q_5}. \tag{43}$$

A method for obtaining the performance characteristics from these relations is illustrated as follows. A parameter

$$\rho \triangleq q_2/q_1, \tag{44}$$

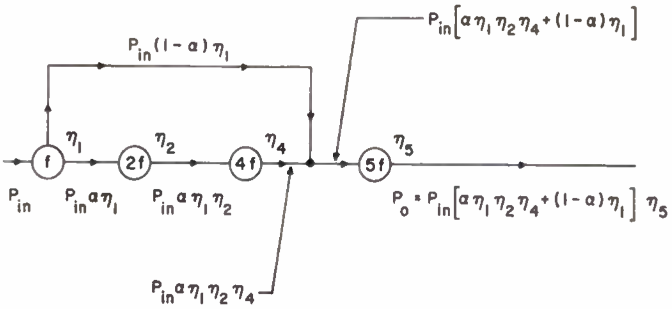
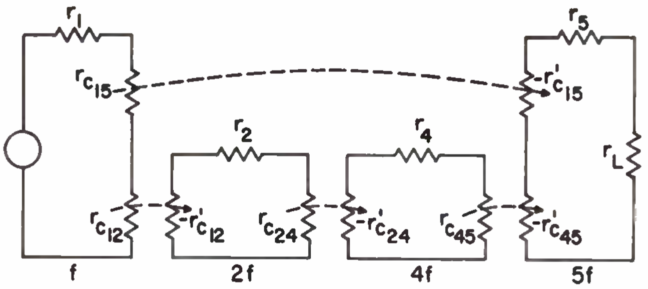


Fig. 6—Power-flow graph of the 1-2-4-5 mode frequency quintupler.



r_n = TOTAL CIRCUIT LOSS RESISTANCE AT FREQUENCY nf

$r_{C_{mn}}$ = CONVERSION RESISTANCE FROM FREQUENCY mf TO FREQUENCY nf

$-r'_{C_{mn}}$ = EQUIVALENT NEGATIVE RESISTANCE AT FREQUENCY nf DUE TO POWER CONVERSION FROM FREQUENCY mf

Fig. 7—Simplified equivalent circuit of the 1-2-4-5 mode frequency quintupler.

which is a function of load resistance, is introduced. Effective normalized loss factors are defined:

$$\Delta_n = nr_n/\beta_2 = nR_n\omega_0C_0\beta_1/\beta_2 = \beta_1/\beta_2Q_n, \qquad (45)$$

where r_n includes both the external circuit losses and the loss due to the variable-capacitance diode at frequency nf. Note that the normalizing factor for impedance is now $n/\beta_1\omega c_0$ instead of $1/\omega c_0$ as defined previously. This is because the presence of variable charge components changes the apparent capacitance of the diode at nf by this amount. In order to have the maximum voltage driving the diode and yet prevent the diode from conducting in the forward region,

$$q_1 + q_2 + q_4 + q_5 = q_T = 1 \qquad (46)$$

under the restrictive assumption that all of the charges can add in phase.

With the use of (44) and (46), (41) becomes

$$q_4 = \frac{q_1}{2\rho} - \Delta_2. \qquad (47)$$

Substitution of (47) into (42), neglecting the terms with $(\Delta_4)^2$, gives

$$q_5 = \rho^2\frac{q_1}{2} - \frac{\Delta_4}{2\rho}. \qquad (48)$$

Finally, (44) and (46)–(48) are combined to give

$$q_1 = \frac{1 + \Delta_2 + \dfrac{\Delta_4}{2\rho}}{1 - \dfrac{1}{\rho} + \dfrac{1}{2\rho} + \dfrac{\rho^2}{2}}, \qquad (49)$$

where the terms containing $\Delta_2\Delta_4$ are neglected. All charges are now determined as functions of ρ and the loss factors. The corresponding normalized input and load resistances (r_{in} and r_L) can also be determined from (40) and (43), respectively. The over-all efficiency is then

$$\eta = \frac{i_1{}^2r_{\text{in}}}{i_5{}^2r_L}, \qquad (50)$$

where each term is expressible, with the use of the preceding relations, in terms of only the loss factors and the parameter. The procedure, then, is to calculate q_1, from (49), and use the result to obtain q_4 and q_5 from (47) and (48). The input and load resistances are then found from (40) and (43). Finally, the efficiency is obtained from (50).

The above analysis dealt with efficiency calculation, given the circuit losses and the diode characteristic. Efficiency optimization is possible by varying the loss factors and the diode characteristic. It is obvious that better efficiency can always be obtained with circuit elements with smaller loss. However, it is not immediately clear how the diode characteristics affect the efficiency. Efficiencies, as functions of r_L and Γ, the nonlinearity of the variable capacitance, calculated with the above procedure, using measured circuit loss, are shown in Fig. 8. At each frequency, the losses were measured with the variable capacitance replaced by a fixed capacitance. It is now seen that higher efficiency is obtained with larger values of Γ. This is because the second harmonic is used as one of the intermediate harmonics. In general, for any order of multiplication the second harmonic should always be used as an intermediate harmonic since it provides the closest coupling with the fundamental. For conventional diodes, where the range of Γ is restricted, $\Gamma = 0.5$ will give the best efficiency. This, again, justifies the assumption that β_k's$=0$ for $k > 2$, since Γ should be made to approach 0.5 for best efficiency.

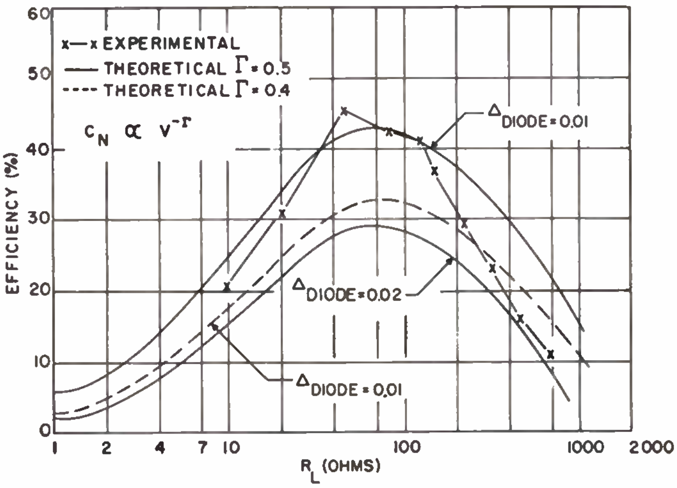


Fig. 8—Efficiency vs load resistance for 1-2-4-5 mode frequency quintupler.

In practical circuits, one of the important factors that lowers the efficiency is stray and case capacitances, since any stray capacitance across the variable capacitance will decrease the effective nonlinearity, Γ', of the diode. Without stray capacitance, Γ is defined by (23) or

$$\Gamma = \frac{\log\left(\dfrac{C_1}{C}\right)}{\log(V + \phi)}, \qquad (51)$$

With stray capacitance C'.

$$\Gamma' = \frac{\log\dfrac{C_1 + C'}{C + C'}}{\log(V + \phi)}. \qquad (52)$$

Since

$$\log\frac{C_1 + C'}{C + C'} < \log\frac{C_1}{C}, \qquad \Gamma' < \Gamma.$$

The decrease in efficiency due to stray capacitance can then be determined exactly.

III. EXPERIMENTAL VERIFICATION

In carrying out the experiments, consideration has to be given to all the points emphasized in the previous section. In order to restrict the current components of any particular harmonic to a single loop, a minimum number of lossy elements must be used. For any particular mode of operation, there are many circuit configurations possible to produce the desired harmonic. A balanced form of the 1-2-4-5 mode was chosen since it eliminates the need for filters at even harmonics otherwise required to prevent the power at these frequencies from being dissipated by the input source resistance and the output load resistance. Careful layout of circuit elements was used to minimize the stray capacitances, thus making the effective Γ very close to the actual Γ of the nonlinear capacitor. The choice of the relative magnitudes of circuit elements was of primary importance, since it determined 1) the effectiveness of localizing the power of the various frequency components to their respective portions of the circuit, 2) the impedance level, 3) the loss in each loop, and 4) the relative losses of each loop with respect to other loops. The circuit used for this 1-2-4-5 mode frequency quintupler is shown in Fig. 9. The variable capacitances used were Hughes HC7001 diodes, which have a Γ of 0.47. Air-core coils having Q values of approximately 130 were used for most of the inductances to minimize the losses. The experiment was carried out at an input frequency of 2.8 Mc and output frequency of 14 Mc. The diodes were back-biased at 10 vdc to provide an adequate working power level. The particular frequencies were chosen because the capacitance of the diode at 10 vdc bias limited the operating frequency to a range of 0.5–50 Mc with reasonable values of inductances and with suitable impedance levels. The circuit was carefully laid out to minimize lead length and stray capacitances. All loops were separately tuned and their losses were measured at their respective frequencies with the diode replaced by a fixed capacitor. To increase the conversion resistances, the input voltage of the frequency multiplier in operation should be as high as possible, without causing the diode to conduct in the forward region. The experimental data obtained with the frequency quintupler is presented in Fig. 8. The series resistance of the diode at these frequencies was measured and found to be constant with frequency. The normalized loss of the diode at fundamental frequency is approximately 0.016. As a function of load resistance, a maximum efficiency of 46 per cent was obtained. The input resistance as a function of signal level is shown in Fig. 10. The input resistance is seen to be directly proportional to the signal level. This also indicates that efficiency decreases as the signal level is decreased, since circuit loss resistances are independent of signal level. Other order multipliers have been built and tested using a similar procedure. A summary of the experimental data is shown in Fig. 11 for multipliers of the order 2 to 5. Also shown is a pre-

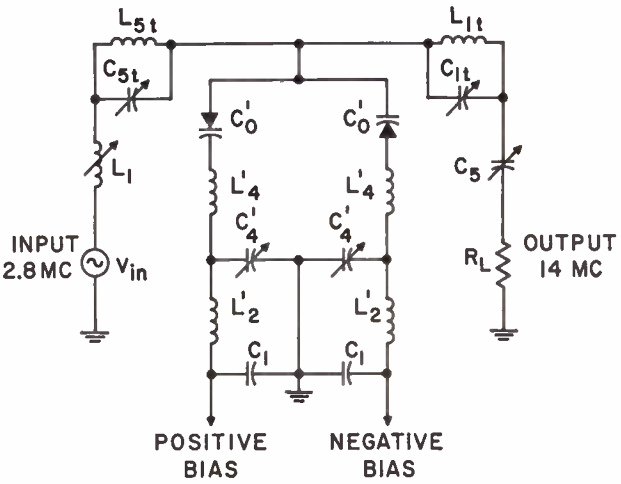


Fig. 9—Experimental circuit for frequency quintupler.

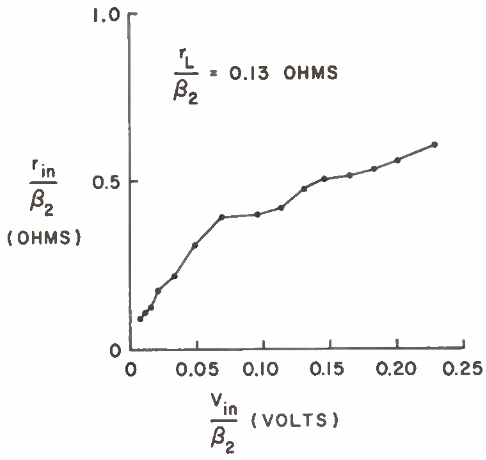


Fig. 10—Input resistance vs input voltage.

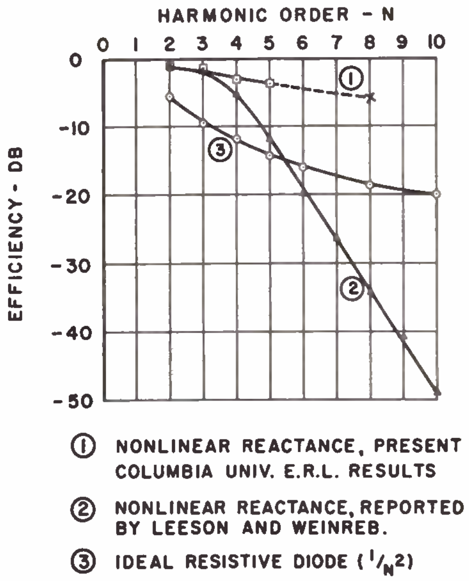


Fig. 11—Efficiency in db vs harmonic number.

dicted efficiency for an eight times multiplier calculated on the basis of using the same diode, and circuit elements similar to those used in the lower-order multipliers. The data is compared with the results obtained by Leeson and Weinreb[2] and Page.[3]

IV. Conclusion

The general procedure outlined in the preceding sections is applicable to frequency multipliers of any order, at any frequency and impedance level since, throughout the analysis, no limitations have been imposed on these factors. The difficulty in obtaining efficient multipliers in the microwave frequency range would be in the excessive loss of the diode. External circuit losses, however, could be reduced since low loss circuits of coaxial or waveguide type may be employed. At low frequencies, *i.e.*, $f < 1$ Mc, the problem is just the reverse, since the diode loss will be small compared with the loss of the coils needed for tuning and effective filtering.

The undesirability of the existence of unwanted harmonics has been repeatedly emphasized. This is somewhat contradictory to the statement that better efficiency can be expected if more currents of different harmonics be allowed to flow through the nonlinear element. This can be explained by a simple example. The 1-2-4-8 mode is used for comparison for an eight-times multiplier. By allowing power at the fifth harmonic to flow through the nonlinear element, the over-all efficiency for this 1-2-4-5-8 mode of operation will be less than that of the 1-2-4-8 mode since power at the fifth harmonic cannot be returned to useful output. If, however, the power of both the third and fifth harmonics are allowed to flow simultaneously, the over-all efficiency of this 1-2-3-4-5-8 mode of operation might be increased due to mixing of the third and fifth harmonics. Hence, the choice of harmonic frequencies is most important in maximizing the efficiency. Having chosen the desired harmonics to improve the performance of the multiplier, the current of each harmonic has to be restricted to flow only in selected portions of the circuit in order to assure maximum usage of its power. To achieve this, high impedance isolation filters have to be used. The use of high impedance filters, however, will generally increase the circuit losses in a practical circuit. Thus, working impedance levels must be determined so that effective filtering can be obtained with reasonable circuit losses.

The nonlinearity of the element used plays an important role in increasing the efficiency of such devices. For a variable-capacitance diode that satisfies the relation of (22), higher efficiency can always be expected with a higher Γ, as witnessed in Fig. 8. The efficiency of the multiplier is a definite function of the Q of the diode. For high-frequency operation, the loss of the diode will be much greater than other circuit losses. A diode with high Q is therefore essential if the frequency multiplier is to be used as power source at microwave frequency.

It is assumed in the calculation that currents of the harmonics of interest have certain phase angle relations when all circuits are perfectly tuned. In the experiment, however, the circuit is tuned for maximum efficiency, which actually corresponds to a slightly detuned condition. Theoretically, higher efficiency is generally obtainable with the circuit slightly detuned. This is because the relative current magnitudes, not the absolute current magnitudes, are very critical in the determination of efficiency. For a frequency tripler, the detuned condition showed much better efficiency than the perfectly tuned condition. For the frequency quintupler, the effect of detuning was much less.

The power handling capability of these devices again depends on the characteristics of the nonlinear element used. For a variable-capacitance diode, the power input is limited by the reverse breakdown voltage of the diode. When the diode is biased at a high negative voltage, the input voltage can be increased proportionally without the diode conducting in the forward region. The negative-bias voltage is nevertheless limited by the condition that the negative peak of the variable component plus the negative-bias voltage should not exceed the breakdown voltage of the diode. This limitation has been imposed because the noise due to reverse current is much greater than that of the forward conduction current. For the experimental frequency quintupler, the input power is approximately 5 mw and the output power 2.3 mw. This is not the maximum power the quintupler can handle. These power levels were chosen because they fitted best to the range of test equipment available.

From a preliminary analysis, for a frequency chain using a nonlinear capacitor diode, an improvement in signal-to-noise ratio over the conventional vacuum tube multipliers of about 10 to 20 db is indicated. Experimental work is currently being carried out, with no figures yet available. The problem of cascading two multipliers efficiently has also been considered.[4]

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[4] T. Utsunomiya and S. Yuan, "High efficiency variable-reactance frequency multiplier," Proc. IRE (Correspondence), vol. 49, p. 965; May, 1961.

The Study of Large-Signal High-Frequency Effects in Junction Transistors Using Analog Techniques*

J. R. A. BEALE† AND A. F. BEER†

Summary—An analog has been designed and constructed in which the base-emitter junction of a transistor is represented by a transmission line lumped into several elements representing distributed base resistance and base-emitter capacitance. The base-emitter capacitance varies with current and therefore each element of the junction is represented by a special circuit which presents the required impedance at any current. The distributed voltage-dependent collector-base capacitance and the resistance of the collector body are also represented.

The analog is used to indicate the magnitude of several effects that arise at large signal levels and high frequencies. The results are generalized to obtain expressions for the variation of base resistance and collector series resistance with transistor parameters and operating conditions.

I. Introduction

In all transistors the base resistance, collector capacitance and emitter capacitance are distributed quantities. Various forms of lumping have been used to derive convenient *equivalent* circuits which are very useful over a limited range of small signal operating conditions. However, the current or voltage dependence of the distributed quantities generally makes any simple form of lumping invalid for large signal applications. To make a detailed analysis of the large signal behavior of a transistor it is, in general, necessary to use at least a partially distributed equivalent circuit. However, this approach leads to some major problems of analysis, and before considering how to tackle them one should decide what effects one can expect to observe.

The flow of lateral current in the base causes the emitter-to-base junction to be nonuniformly biased. The low-frequency effect of the junction conductance has been analyzed by Fletcher [1]. However, at frequencies above the common emitter cutoff frequency f_β the admittances of the elements of emitter capacitance $\omega\delta C_{b'e}$ are greater than the corresponding junction conductances. Therefore, at large signal levels and high frequencies there will be a lateral biasing effect which is greater than the low-frequency effect. The larger the emitter current and the higher the frequency, the more nonuniformly will the emitter inject into the base.

The effects that can be expected from this nonuniform injection are:

1) The part of the emitter which is nearest the base contact injects more than the rest of the emitter and therefore the effective base resistance will be lowered.
2) In transistors which have a collector region of high resistivity, such as diffused-base and grown-junction transistors, the nonuniform injection will cause an increase in the effective value of collector series resistance.
3) At high enough currents the injected carrier concentration will cause significant local modulation of base conductivity. This will entail the setting up of a drift field in this region which is different from that at low injection levels.

It is possible, by making some rather crude approximations, to estimate the order of magnitude of these effects in any particular design of transistor. The authors were required to design a medium power low-voltage transistor to operate at 100 Mc and it appeared that the effects would certainly be significant in such a transistor. It was, therefore, important to find a method of analyzing a partially distributed equivalent circuit. It was found that this could be done quite neatly by analog techniques.

In this paper the theory and construction of such an analog are discussed, and some results are given on the analog of the 100 Mc power transistor. These results are generalized to show how base resistence and collects series resistance depend on transistor parameters and operating conditions.

II. The Equivalent Circuit

Ideally a transistor should be represented by a three-dimensional equivalent circuit in which all the elements are distributed. However, for most transistor structures it is possible to make certain reasonable assumptions about the current flow in order to derive a two-dimensional equivalent circuit. In this paper a diffused base transistor with long parallel base and emitter stripes is considered, and such a transistor, with idealized geometry, is shown in Fig. 1. (A concentric configuration can be similarly treated.)

It will be assumed that:

1) The transistor is to be operated in a common emitter configuration.
2) The injected carrier concentration is everywhere small compared with the equilibrium majority

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carrier concentration (the validity of this assumption can be checked with the analog when it is constructed).

3) The operating frequency is substantially greater than f_β and substantially less than f_1 (say $2f_\beta < f < f_1/2$), where f_1 is the frequency at which the common emitter current amplification $|\beta|$ is equal to one.

4) The emitter depletion capacitance can be regarded as being independent of emitter current.

5) The Early effect can be neglected over the frequency range being studied, since movement of the depletion layer with voltage occurs largely in the relatively lightly doped collector region, and the output impedance is determined mainly by feedback through the collector capacitance.

6) The emitter and base contacts are sufficiently long that current flow in the z direction (Fig. 1) can be neglected. This allows a two-dimensional circuit to be developed.

7) The base layer is thin and can be regarded as a sheet in the y-z plane (Fig. 1) of sheet resistance ρ_b. In the two-dimensional circuit this sheet can be represented by a one-dimensional distributed resistance ρ_b/Z per unit length. This resistance is short-circuited under the base contact.

8) Based on assumption 6), the collector body can be represented by a two-dimensional sheet of sheet resistance ρ_c/Z, where ρ_c is the resistivity of the collector body.

9) The base region can be considered as divided into a finite number of elements of area $Z\delta y$. Provided that the variation of any parameter over any element is insignificant this partial lumping of the distributed quantities introduces no error. (The determination of the minimum number of such elements is discussed in Section V-B.)

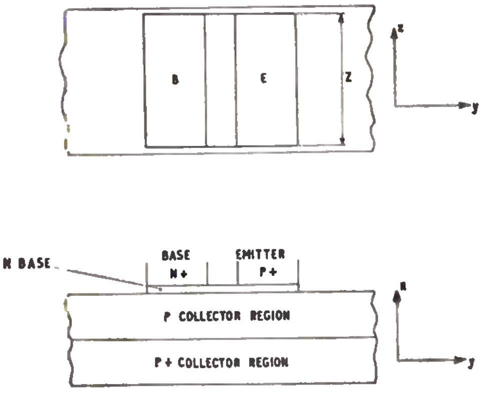

Fig. 1—Idealized transistor structure.

An equivalent circuit based on the same concepts as the hybrid-π circuit (Fig. 2) [2] can now be developed and this is shown in Fig. 3. The effect on this circuit of a drift field in the base region may be neglected in grounded emitter but substantial corrections are necessary in grounded base [3].

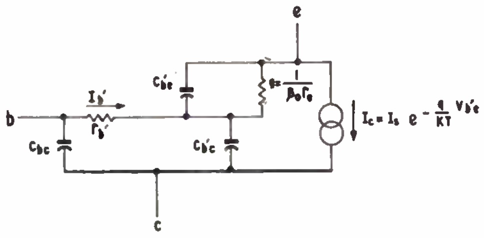

Fig. 2—Hybrid-π equivalent circuit. (Collector series resistance has been omitted.)

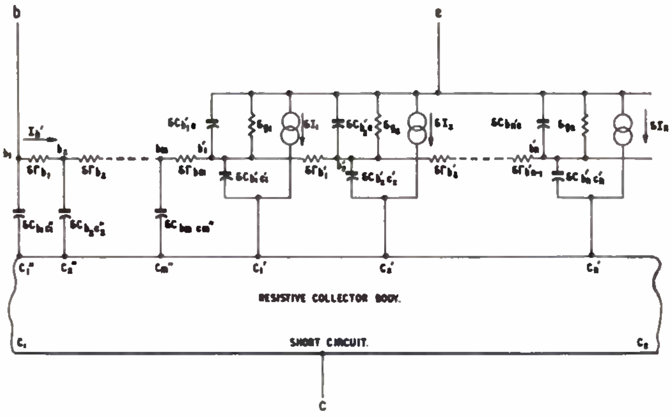

Fig. 3—Partially distributed hybrid-π equivalent circuit.

III. The Basic Operation of the Analog

A. Basic Equations Describing the Transistor To Be Simulated

As the equivalent circuits of Figs. 2 and 3 are basically similar, the equations describing the elements are the same except that in the partially distributed circuit the parameters have values appropriate to elements of the transistor. These equations are [2]

$$\delta I = k_1 Z \delta y e^{-(q/kT)V_{b'e}} \tag{1}$$

$$\begin{aligned}\delta C_{b'e} &= \delta C_{te}\left(1 + \frac{\delta C_{se}}{\delta C_{te}}\right)\\ &= \delta C_{te}\left(1 + \frac{q}{kT}\cdot\frac{\delta I}{\omega_I \delta C_{te}}\right)^{1}\end{aligned} \tag{2}$$

---

[1] Eq. (2) has been expanded using the relations

$$\omega_I = \frac{1}{r_e C_{se}} [4] \quad \text{and} \quad r_e = \frac{kT}{q}\frac{1}{\delta I}.$$

where

$Z\delta y$ is the area of the element, and k_1 is a constant,
ω_I is the reciprocal of the base transit time,
δC_{te} is the depletion capacitance of the junction,
δC_{se} is the diffusion capacitance.

$$\delta r_{b'} = \frac{\rho_b}{Z}\cdot\delta y, \tag{3a}$$

$$\delta r_b = \frac{\rho_b}{Z}\cdot\delta y, \tag{3b}$$

$$\delta C_{b'c'} = k_2 Z\delta y\cdot[f(V_{c'} - V_{b'})], \tag{4a}$$

$$\delta C_{bc''} = k_2 Z\delta y\cdot[f(V_{c''} - V_b)], \tag{4b}$$

$$\delta g = \frac{q}{kT}\cdot\frac{\delta I}{\beta_0}, \tag{5}$$

where β_0 is the low-frequency value of β.

In the frequency range considered, δg is negligible compared with the admittance of $\delta C_{b'e}$.

The resistance in the collector body is considered separately in Section III-C-2). It is seen that (1) and (2) are interdependent and that they, in conjunction with (3) completely describe the emitter parameters of the transistor. Eqs. (1) and (4) describe the collector parameters.

B. Simulation of Emitter Parameters

Consider the element shown in Fig. 4. The capacitance C_M is a Miller capacitance [5] and the charge on it is proportional to the difference between the collector and base voltages. If the base voltage is changed at a sufficiently low frequency, the change of voltage at the collector is opposite in sign. It is shown in Appendix I that the current flowing into C_M is the same as that which would flow in an equivalent capacitance C_{eq} between base and emitter:

$$C_{eq} = C_M\left(1 + \frac{q}{kT}\cdot R_{LA}\cdot I_{LA}\right), \tag{6}$$

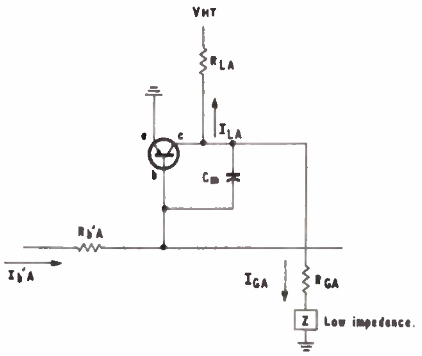

Fig. 4—Miller element to simulate R_b', C_e and δI.

where

$$I_{LA} = k_1 e^{-qV_{be}/kT}$$

It can be seen from Fig. 4 that

$$I_{GA} = I_K + I_{LA}\cdot\frac{R_{LA}}{R_{GA}} = I_K + \frac{R_{LA}}{R_{GA}}\,k_1 e^{-qV_{be}/kT}, \tag{7}$$

where I_K is a constant.

Eqs. (7) and (6) are of the same form as (1) and (2), respectively, (except for the constant current I_K). The emitter-to-base junction characteristics of each element of the transistor can therefore be simulated by such a Miller element. The bases of these elements are connected together through resistances $R_{b'A}$ [corresponding to (3)] to build up an analog of the base and emitter regions of Fig. 3.

The matching of the analog to the transistor is discussed in Section IV.

C. Simulation of Collector Parameters

1) *Collector Capacitance:* In order to simulate directly the effect of the distributed collector capacitance $C_{b'c'}$ it would be necessary to connect capacitors between the b' points and the c' points in the analog (Fig. 3). (For brevity the identical discussion regarding the $C_{bc''}$ elements is omitted.) However, the collector junction capacitance is dependent on voltage and for a large signal analog it would be necessary to use reverse biased junction diodes. Furthermore, if the analog frequency is to be kept conveniently low, the junctions of these diodes would need to be very large and it would be inconvenient to change them in order to simulate different transistors.

For these reasons a more indirect technique has been chosen. The current from the current generators is fed into a resistive sheet which represents the collector body and from there into a common load. The voltage developed across this load is amplified to correspond with the collector voltage on the transistor and is applied to a diode in which the junction has been formed in the same manner as the collector-base junction of the transistor. The capacitive current through this diode, after suitable amplification, is fed back to the b' points and also into the collector circuit to simulate the current flowing from collector to base through $C_{b'c'}$. Different collector capacitances can then be simulated by varying the gain of the amplifiers. The circuit is shown schematically in Fig. 5.

In this method the changes of voltage in the base, and voltage drops in the resistive collector body, are neglected in comparison with changes in voltage across the collector load. Measurements on the analog of voltages in the base and collector body indicate that these assumptions give a self-consistent system for reasonably designed transistors operating within their normal range. However, some correction could be made for the effect of changes in base voltage by inserting suitable capacitances from base to earth, and for voltage drops in the collector body by adjustment of the amplifier gains.

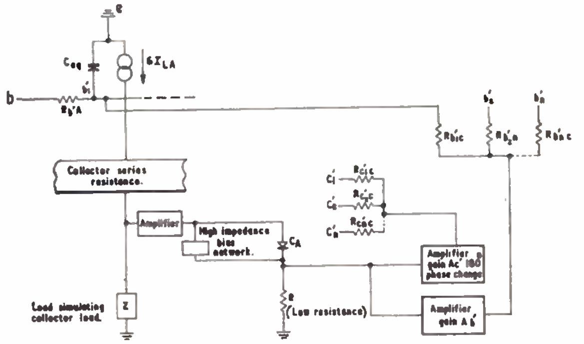

Fig. 5—Diagrammatic layout of circuit for the simulation of feedback through the collector capacitance.

2) *Resistive Collector Body:* In grown-junction and diffused-base transistors, the collector region is of fairly high resistivity and has a resistance which cannot in general be neglected. For large signal operation the resistance is particularly important in that it determines the minimum operating voltage for a given collector current. One may therefore define an effective collector series resistance $r_{cc'}$ as

$$r_{cc'} = \frac{\text{Maximum peak voltage drop across the collector body}}{\text{Peak current through the collector body}}.$$

However, this collector series resistance is not constant. If the base is sufficiently thin it may be assumed that the distribution of the current entering the collector is the same as that of the current injected by the emitter into the base. Since the lateral bias in the base controls this distribution $r_{cc'}$ varies with frequency and current. (It should be noted that since the majority of the collector current I_c flows through a smaller effective resistance than $r_{cc'}$, the instantaneous power loss is less than $I_c^2 r_{cc'}$, and for other calculations, such as power amplification, a lower value should be used.)

The voltage drops in the collector body could be simulated and measured using an electrolytic tank, but in accordance with assumption 6) voltage drops along the length of the base contact (*i.e.*, the z direction of Fig. 2) can be neglected. The resistive body can therefore be simulated by a resistive sheet such as that obtained by coating an insulating substrate with graphite.

VI. MATCHING ANALOG TO TRANSISTOR

A. Base Circuit and Current Generators

It can be seen from (1) and (7) that the current from the current generators in both the analog and the transistor are related to the base voltages by the same exponential factor. If voltage changes in the base circuit of the analog and the transistor are equal, the ratio of changes in I_{GA} to δI will be constant. However, (2) and (6) show that the ratio of C_{eq} to $\delta C_{b'e}$ can be made independent of current only if the right-hand terms in the brackets are equal. This fixes the ratio of the current generators, *i.e.*,

$$\frac{\delta I}{(I_{GA} - I_K)} = R_{GA}\omega_I \delta C_{te}. \tag{8}$$

Since changes in base voltage in the analog are equal to those in the transistor, the product of the base current and resistance must be made the same, *i.e.*,

$$\frac{I_{b'}}{I_{b'A}} = \frac{R_{b'A}}{\delta r_{b'}}. \tag{9}$$

Apart from current through $C_{b'c'}$ which is simulated independently, the current distribution in the base will be correct if the ratio of the reactance of C_{eq} of each element to the reactance of the corresponding element in the transistor is also $R_{b'A}/\delta r_{b'}$.

$$\therefore \frac{\omega \delta C_{te}}{\omega_A C_M} = \frac{\omega}{\omega_A} \cdot \frac{\delta C_{b'e}}{C_{eq}} = \frac{R_{b'A}}{\delta r_{b'}}. \tag{10}$$

B. Collector Circuit

1) *Collector Series Resistance:* It was assumed in the introduction that the collector body behaves as a conducting sheet (with the x and y dimensions of the original body) into which a distributed current is fed by the current generators. In the partially lumped analog it is necessary to feed the currents into a sheet (of sheet resistance R_A) of the same effective shape. The problems associated with this are discussed in Section V-C.

Voltage drops $V_{cc'A}$ in the conducting sheet are related to those in the collector body $V_{cc'}$ by the equation

$$\frac{V_{cc'A}}{V_{cc'}} = \frac{(I_{GA} - I_K)}{\delta I} \cdot \frac{Z}{\rho_c} \cdot R_A. \tag{11}$$

2) *Collector Capacitance:* On the assumptions made in Section III-C the current flowing through the collector capacitance is given by the expression

$$I_{b'c'} = \frac{d}{dt} f(V_c, C_{b'c'}).$$

In the analog the capacitance C_A (Fig. 5) representing the collector capacitance is the same function of voltage as $C_{b'c'}$ and the voltage across it has the same wave form and amplitude as V_c although the frequency is different. It is intuitively obvious (a proof is given in Appendix II) that the waveforms of the currents through the two capacitances are the same and that the ratio of the amplitudes is given by

$$\frac{I_{cA}}{I_{b'c'}} = \frac{C_A}{C_{b'c'}} \cdot \frac{\omega_A}{\omega}, \tag{12}$$

where $C_A/C_{b'c'}$ is the ratio of the capacitances when the same voltage V_L is applied to both.

If the analog capacitance C_A is placed in the circuit shown diagrammatically in Fig. 5, the ratio of currents is given by

$$\frac{I_{b'c'A}}{I_{b'c'}} = \frac{\sum\limits_n I_{b_n'c}}{I_{b'c'}} = \frac{C_A}{C_{b'c'}} \cdot \frac{\omega_A}{\omega} \cdot \frac{R}{\sum\limits_n R_{b_n'c'}} A_{b'}, \tag{13}$$

where $\sum_n$ means the parallel combination of the resistances. The gain of the amplifier ($A_{b'}$) must be adjusted so that the ratio of the transistor to analog currents in (13) is the same as that in (9). Therefore,

$$A_{b'} = \frac{C_{b'c'}}{C_A} \cdot \frac{\omega}{\omega_A} \cdot \frac{\sum\limits_n R_{b_n'c}}{R} \cdot \frac{\delta r_{b'}}{R_{b'A}} \cdot \tag{14}$$

Corresponding currents must be fed into the collector circuit and the required amplifier gain can similarly be determined from (8).

$$A_{c'} = \frac{C_{b'c}\omega}{C_A\omega_A} \cdot \frac{\sum\limits_n R_{c_n'c}}{R} \cdot \frac{1}{\omega_I \delta C_{te} R_{GA}} \cdot \tag{15}$$

The currents fed into the base and collector circuits must be 180° out of phase and may be conveniently derived from a phase-splitting amplifier.

V. DETAILS OF ANALOG CONSTRUCTION

A. Parameters of Miller Element

As mentioned in Section III-B, there are conditions on the parameters C_M, R_{LA} and I_{LA} which must be met if (6) is to be accurate. The current in the base line (Fig. 4) must be capacitive and therefore the reactance of C_{eq} must be very small compared to the input resistance $\beta_0 r_e$ of the analog transistor. In order to have a 180° phase shift in voltage between base and collector, the analog frequency f_A must be much less than the cutoff frequency f_β of the analog transistor and the collector load must be mainly resistive (*i.e.*, $1/\omega_A C_M \gg R_{LA}$) (Appendix I). Finally, the injection levels in the transistor must always be low enough for (7) to be reasonably accurate.

1) Range of C_{eq}: In large signal class B operation the analog transistor currents will swing from zero to some maximum at which $R_{LA} \cdot I_{LA}$ approaches the *HT* voltage V_{HT}. Thus to obtain the maximum range of C_{eq} [see (6)], V_{HT} must be as high as possible, but it is limited by the voltage which can safely be applied to the transistors in the cutoff state.

It is possible that the maximum range of capacitance that can be obtained (approximately 400:1 for $V_{HT} = 10$ volts) is not sufficient particularly for the first elements in the base line. However, in this case, the effect of the depletion capacitance of these elements on the total base current will generally be very small. Therefore, C_M can be increased from the value corresponding with δC_{te} (and R_{LA} decreased by the same factor) in order to make C_{eq} correspond with $\delta C_{b'e}$ in the conducting state without introducing a significant error. Since I_{GA} must remain unaltered it can be seen from (7) that R_{GA} must also be decreased by the same factor. However, for simplicity it is assumed in the rest of this paper that C_M corresponds with δC_{te}.

B. Number of Elements

The number of elements required to represent the transistor with sufficient accuracy depends on the operating conditions. Owing to the ac lateral biasing effect, the current distribution will become less uniform the higher the frequency and the larger the current swing. To find the required number of elements the following procedure may be used. The analog is constructed with a few elements representing equal junction areas in the transistor, and is set up to simulate the extreme condition in which the transistor is required to operate. The peak currents through the collector loads R_{LA} of the elements are measured and are plotted as a function of distance in the y direction. A stepped line is obtained which represents the current density as a function of distance. A smooth curve is constructed such that for each element the areas under the stepped line and the curve are equal.

It will be found that differences in the current density in the elements near the base contact are greater than those between the other elements, and the first elements should be subdivided. If desired, the lumping of the later elements can usually be made more coarse. Another stepped line is then plotted and the corresponding smooth curve is drawn. This process is repeated until there is no significant difference between successive smooth curves. The resultant curve represents the actual current distribution in the transistor (Fig. 6).

C. Current Generators and Collector Series Resistance

The current through the resistances R_{GA} must be fed into the resistive body at points which correspond to the points 1, 2, $\cdots$, n, in Fig. 6. To avoid errors due to the contacts with the resistive body it is desirable to allow the currents to spread in the conducting sheet before they enter the part of the sheet that represents the collector body. The arrangement shown in Fig. 7 appears to be very satisfactory. The line $c'c''$ represents the collector edge of the collector depletion layer, movements in this edge with collector voltage being neglected. The line c_1c_2 represents the edge of the p^+ region and at this line the conducting sheet must be short circuited.

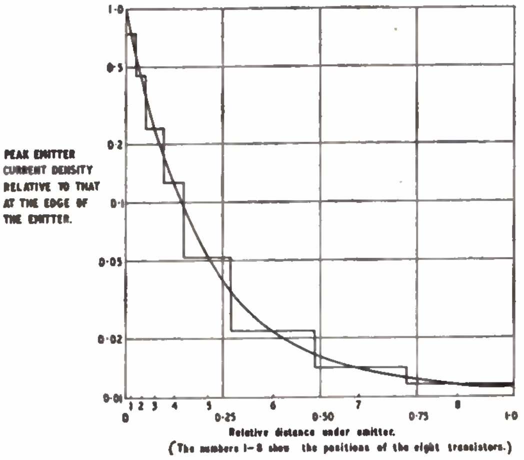

Fig. 6—Plot of peak current density (100 ma total current) vs distance under emitter showing final number and spacing of elements.

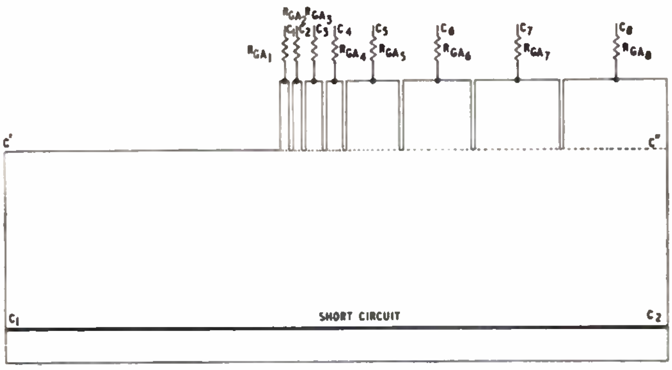

Fig. 7—Diagram of conducting sheet representing the resistive collector body. (The region between the lines $C'C''$ and C_1C_2 simulates the collector body of the transistor.)

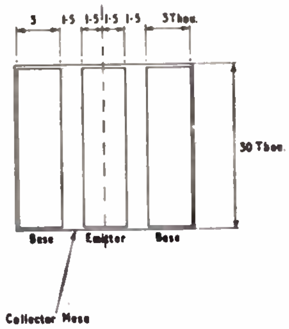

Fig. 8—Plan view of the 100-Mc 0.5-watt diffused base transistor.

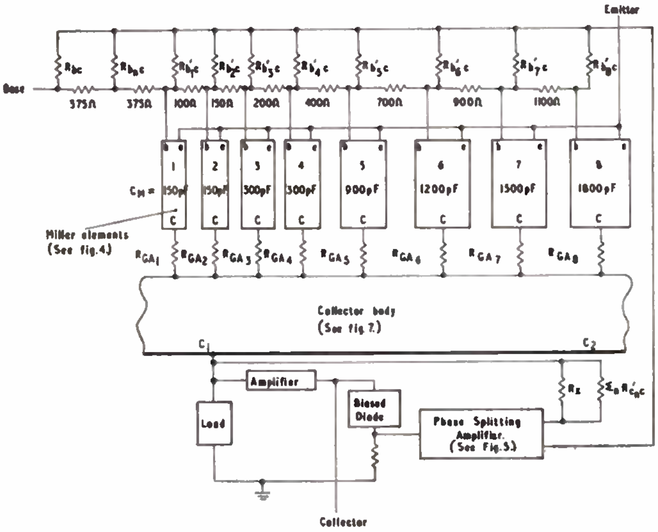

Fig. 9—Diagram of analog. (Variable resistors were used for the elements of base resistance, the values shown corresponding to the particular transistor studied.)

VI. Analog of 100-Mc Medium Power Transistor

The preliminary design of the 100-Mc medium power transistor is shown in Fig. 8. Owing to the symmetry there will be no current across a plane through the center line of the emitter, and it was decided to build an analog of one half of the transistor. All quantities discussed in this paper are those of half the transistor.

The doping of the base layer was such that the resistance between the base and the edge of the emitter was about 7 Ω and the resistance of the layer under the emitter was about 40 Ω and was taken, in fact, as $6\times 7=42$ Ω. The emitter depletion capacitance when the transistor was conducting was taken as $9\times 7=63$ pf, and the intrinsic f_1 $(\omega_l/2\pi)$ was about 300 Mc.

A. Number of Elements

The half transistor was required to deliver up to 100-ma peak current in class B at 100 Mc. This operating condition was used to find the minimum permissible number of elements (Section V-B). Six elements representing equal areas were tried initially, and eventually eight elements, representing different areas, were found to be satisfactory. The area represented by each element can most conveniently be expressed in terms of a unit of 1/42 of the emitter area. Starting from the base side of the emitter the 8 elements corresponded to 1, 1, 2, 2, 6, 8, 10, 12 units, respectively. A block diagram of the analog is shown in Fig. 9.

B. Base Resistance

The resistance between the base contact and the center of the first element is $7\frac{1}{2}$ Ω and between the centers of the other elements the resistances are 1 $1\frac{1}{2}$, 2, 4, 7, 9, 11 Ω. It was found convenient to use 100 times these resistances in the base line of the analog, and this fixed the ratio of the base currents [see (9)]. The appropriate resistances are shown in Fig. 9.

C. Emitter Capacitance

10 kc was chosen as a convenient analog frequency to represent 100 Mc. Since the emitter depletion capacitance of 63 pf is to be simulated in 42 elements, the depletion capacitance per element is 1.5 pf. Eq. (10) therefore defines C_M as 150 pf per unit. The appropriate capacitance value for each element is shown in Fig. 9.

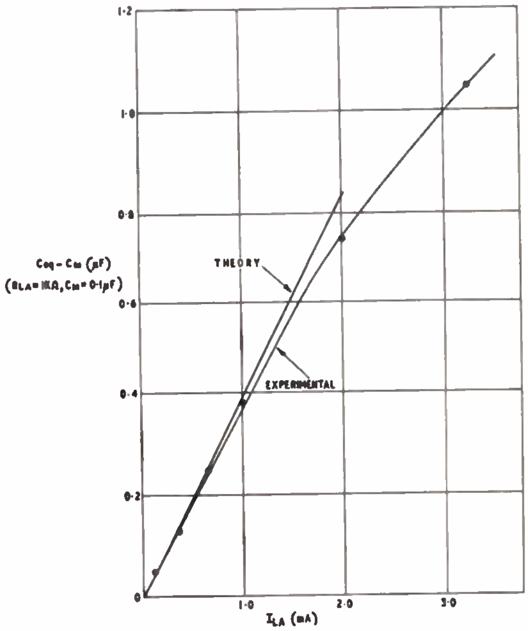

Fig. 10—Experimental check of the relationship between the equivalent capacitance C_{eq} and the collector current I_{LA} in an OC44 transistor at room temperature (Fig. 4).

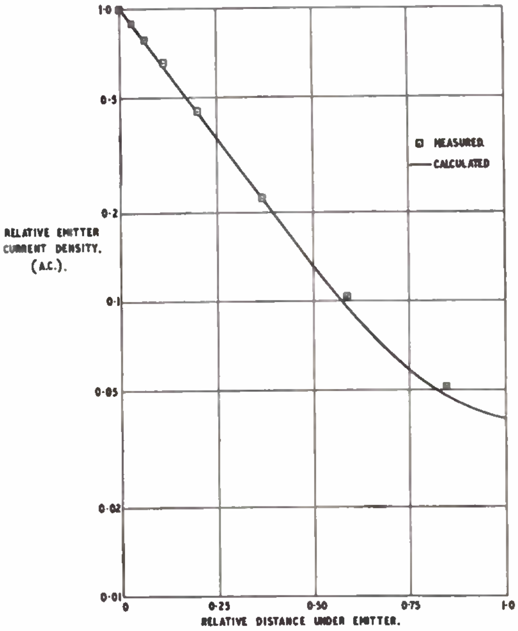

Fig. 11—Experimental check of the small signal emitter current density as a function of distance under emitter (50 ma dc).

Fig. 10 shows the measured relationship between C_{eq} and emitter current of an OC44 which was the type chosen for the Miller transistors. This shows that (6) is closely obeyed up to about 2 ma. At higher currents the high injection level effects reduce the capacitance. It is therefore desirable to limit the maximum current in each transistor to 2 ma.

The maximum ratio of C_{eq} to C_M is obtained when $I_{LA}R_{LA} = V_{HT}$. Choosing $V_{HT} = 10$ v, a desirable minimum value of R_{LA} is 5 kΩ.

However, the last transistor in the line must have a C_M of $12 \times 150 = 1800$ pf which at 10 kc has an impedance of about 10 kΩ. This impedance should ideally be much more than R_{LA} (Appendix I) and a desirable minimum value of R_{LA} is 1 kΩ.

A compromise between these two desirable limits must be made and R_{LA} was chosen to be 1.8 kΩ. The technique discussed in Section V-A-1) was in fact used to limit the maximum current in all elements to less than about 2 ma, but this complication produced no substantial change in the results.

The accuracy of the analog *RC* transmission line in the base was checked by comparing the measured small-signal current distribution with the theoretical distribution for an open-circuited line. The result, shown in Fig. 11, was satisfactory.

D. Emitter to Collector Current Generators

In order to simulate accurately the infinite internal impedance of the current generators, the R_{GA} resistances must be large compared with the common loads in the collector circuit, including resistances in the graphite sheet which simulates $r_{cc'}$. The ratio of $\delta I/I_{GA}$ of 2000 was chosen and (8) defines R_{GA} as 705 kΩ for one unit. Eq. (8) also shows that the R_{GA} resistance for each element must be inversely proportional to the area it represents.

E. Collector Series Resistance

In the design of the transistor it was assumed that the distance between the base-collector junction and the p^+ contact could readily be made about half the emitter width, but that a smaller value would be somewhat difficult to reproduce with alloying techniques.

The shape of the conducting sheet in the analog was therefore chosen to simulate this condition. The currents from the R_{GA} resistors were fed into a graphite-coated sheet, over a front of 5.2 inches (Fig. 13). At a distance of 2.6 inches a line of silver paint was painted on the graphite to simulate the p^+ contact.

To avoid distorting the potential distribution within the simulated collector body, the contacts from the R_{GA} resistances were made remote from the simulated collector edge of the depletion layer ($c'c''$ of Fig. 13). To give a slightly better representation of the smoothly varying current density across the collector to base junction, the I_{GA} currents from the first two elements were each divided into two components which were fed into the collector body separately.

F. Collector Capacitance

The $R_{b'c}$ resistances must act as current generators and must therefore be large compared with the base impedances. Their parallel combination resistance $\sum_n R_{b_n'c}$ determines the required amplifier gain $A_{b'}$ [see (14)]. The $R_{b'c}$ for one unit was chosen as 420 kΩ. The value for each element must be inversely proportional to the area it represents. Eq. (15) defines the relationship between the amplifier gain and $\sum_n R_{c_n'c}$ in order to simulate the collector current that flows into the collector side of $C_{b'c}$. $\sum_n R_{c_n'c}$ was chosen as 200 kΩ as shown in Fig. 9 in order that $A_{b'}$ and $A_{c'}$ should be equal (a lumped feed to the collector terminal was used because the result differed little from that obtained with a distributed feed to the c' points).

G. *Notes on the Complete Analog*

Most of the analog has been discussed in the previous sections, but a few further remarks are necessary to explain all the components shown in Fig. 9.

1) The 7.5 Ω between base contact and the first element of the emitter has been simulated by two resistors of 375 Ω. To the connection between them is connected a resistor R_{b_nc} which simulates the collector capacitance in the region between base and emitter contacts. The feedback through the capacitance under the base contact is simulated by R_{bc}. The magnitudes of R_{b_nc} and R_{bc} are determined by the relationship

$$\frac{1}{R_{bc}} : \frac{1}{R_{b_nc}} : \frac{1}{\sum_n R_{b_n'c}} = A_b : A_{be} : A_e, \tag{16}$$

where A_b, A_{be} and A_e are the areas of the base, the region between base and emitter, and the emitter, respectively.

The collector current flowing into the collector side of these portions of the junction is simulated by R_x where

$$\frac{R_x}{\sum_n R_{c_n'c}} = \frac{A_e}{A_b + A_{be}}. \tag{17}$$

2) An amplifier is used between the load and the biased diode. This is necessary because the load must have a low impedance in order that the R_{GA} resistors act as current generators. The load was chosen to simulate the characteristics of the transistor load under consideration, but its dynamic impedance was always made less than 1 kΩ. The amplifier is required to obtain the same amplitude of collector voltage as in the transistor.

H. *Results for a 100-Mc Medium Power Transistor*

The principal purpose of the analog was to examine large signal bias effects in the base line. This effect is shown in Fig. 12 for a class B sine wave of base voltage. The peak emitter current densities have been normalized to the density at the edge of the emitter. It can be seen that the emitter current is highly localized to the region nearest the base contact, particularly at high currents. This result is generalized in Section VII.

One of the effects of this current crowding is that on the average the base current flows a smaller distance under the emitter. Qualitatively, this clearly results in a lower effective base resistance, but the distortion of the current waveform due to the nonlinear capacitances makes it difficult to measure this resistance. However, the total power dissipated in the base layer can be determined by measuring across the elements of base resistance with a wave analyzer to find the power at all harmonic frequencies. (In practice a good approximation can be obtained from the fundamental only, or from peak measurements on an oscilloscope.) The effective value of $r_{b'}$ is that in which the total base current would dissipate the same power. At 100-Mc with a class B sine wave of base voltage this resistance was determined as the base resistance between base and emitter contact plus 2.2 Ω. This resistance under the emitter is about 6 times less than that at low currents and small signal levels.

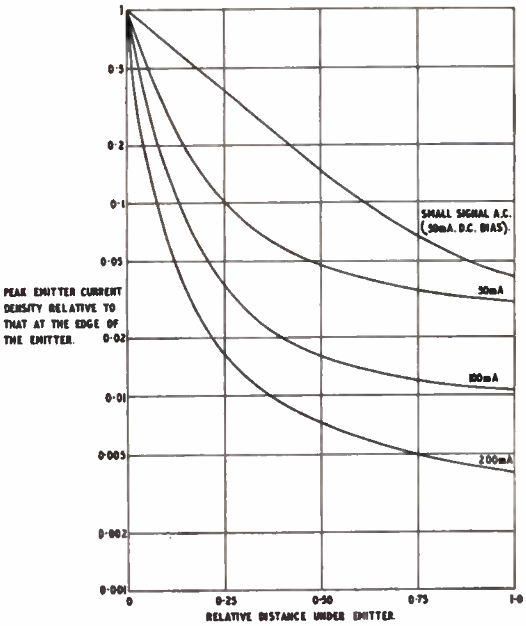

Fig. 12—Normalized peak emitter current density as a function of distance under emitter for various peak currents.

The effect of the nonuniform injection on the collector series resistance is shown in Figs. 13 and 14. In Fig. 13 lines of equal peak-voltage drop across the collector body are shown for a 100 ma peak class B waveform at 100 Mc. These have been normalized to the maximum measured voltage drop. Fig. 14 shows the variation of $r_{cc'}(Z/\rho_c)$ with collector current and frequency.

The circuit simulating collector capacitance can be used in a variety of ways: for example, to demonstrate effects of complete neutralization, partial neutralization or neutralization with a fixed capacitance. However, in this paper we have only included the effect of feedback through that part of the collector capacitance that is under the emitter $C_{b'c'}$. The conditions simulated were that the collector load was a tuned circuit of high Q and that the collector voltage was a sine wave of 10 volts amplitude superimposed on a -12 volt bias. $C_{b'c'}$ was taken as 2 pf at -12 volts bias. The results are shown in Fig. 15 where the waveform and amplitude of the emitter current density at various points are shown

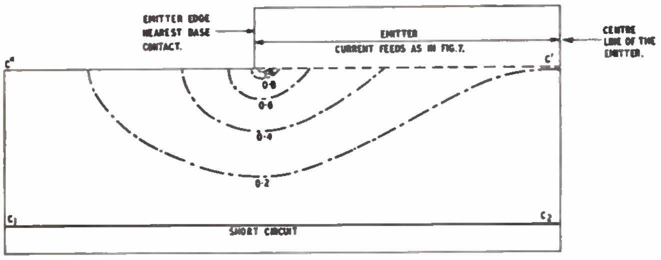

Fig. 13—Lines of peak potential in the collector body relative to the P^+ contact for a 100 ma peak class B waveform.

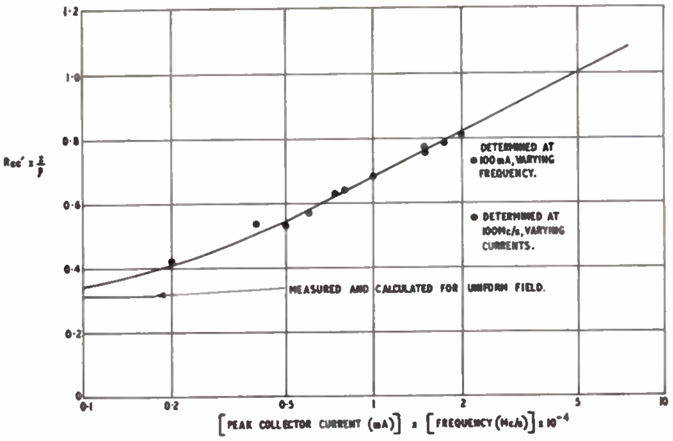

Fig. 14—Increase of collector series resistance with collector current and frequency.

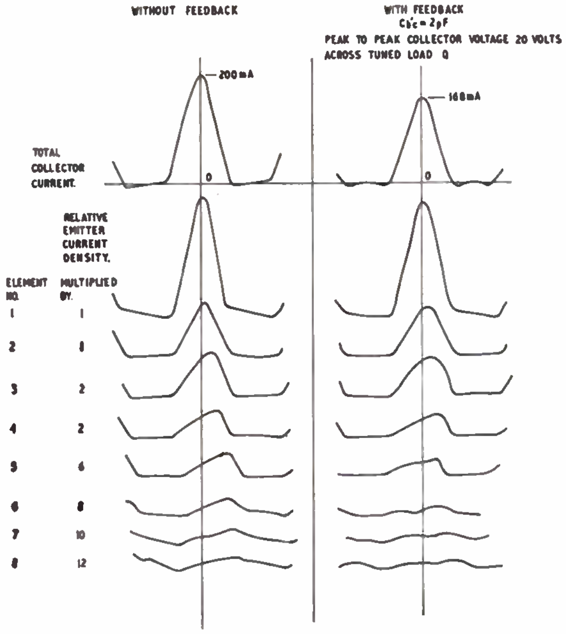

Fig. 15—Emitter current waveforms in the elements. The positions of the elements shown in Fig. 6.

for the same input drive. As can be seen the attenuation along the line is increased and the waveforms are further distorted by the feedback.

VII. GENERALIZATION OF RESULTS

Although the analog can solve a wide range of specific problems it is useful to be able to generalize the results obtained from it. This can be done by finding the dependence of the emitter current distribution on the resistance of the base layer, the f_1 of the transistor, and on the amplitude and frequency of the collector current. From this dependence the variation of base resistance and collector series resistance can be deduced.

Two assumptions are required in addition to those in Section II:

1) The emitter and collector depletion capacitances can be neglected.
2) There is negligible emitter current in the region furthest from the base contact.

It is shown in Appendix III that for a given waveform of base current, the emitter current per unit length in the y direction can be found from a normalized distribution by

a) Multiplying the current density by

$$\left(I_{cpk}{}^2 \frac{\omega}{\omega_1} \frac{\rho_b}{Z}\right).$$

b) Cramping the scale in the y direction by a factor

$$\left(I_{cpk} \frac{\omega}{\omega_1} \frac{\rho_b}{Z}\right),$$

where I_{cpk} is the peak collector current.

Case a) is relevant when considering the possibility of high injection levels. This effect would cause local swamping of the drift field and hence variation in the f_1 and reduction in emitter efficiency. Case b) determines the dependence of the effective width of the band of emitter current on the various parameters.

A. Base Resistance

The effective base resistance is the sum of the resistance from the base contact to the nearest edge of the emitter r_{b0} and the resistance of the effective base layer under the emitter. The width of this layer is inversely proportional to the cramping factor and therefore

$$r_{b'} = r_{b0} + \frac{k_1}{I_{cpk} \cdot \dfrac{\omega}{\omega_1}}, \tag{18}$$

where k_1 is a constant for a given base current waveform. (It is interesting that the resistance per unit length of the base layer under the emitter cancels out.) Measurements on the analog were made with class B sine waves of base voltage. This is not sufficient to deine the base current waveform which is also affected by the relative magnitudes of the two terms in the expression for $r_{b'}$ but this was probably reasonably typical of class B operation. r_{b0} was 7 Ω and at $\omega_1/\omega = 3$, $I_{cpk} = 100$ ma the second term in (18) was about 2 Ω as calculated from peak power dissipation. Therefore, for this class B operation

$$k_1 \doteq 67\ \Omega\ \text{ma}. \tag{18a}$$

B. *Collector Series Resistance*

If the band is narrow compared with the width of the collector body t the current spreads out in the body nearly cylindrically (Fig. 13).

$$\therefore\ r_{cc'} = k_2 \frac{\rho_c}{Z} \ln\left(\frac{k_3 l}{\text{bandwidth}}\right)$$

$$= \frac{k_2 \rho_c}{Z} \ln\left(k_3 l I_{cpk} \frac{\omega}{\omega_1} \frac{\rho_b}{Z}\right). \tag{19}$$

Fig. 14 shows the verification of this formula with the analog. For the same class B waveform as for (18a), Fig. 14 can be used to determine that

$$k_2 \doteq 0.20, \tag{19a}$$

$$k_3 \doteq \frac{1}{26} \cdot (\Omega\ \text{ma})^{-1}. \tag{19b}$$

VIII. CONCLUSIONS

An analog has been designed and constructed which indicates the behavior of transistors under large signal high-frequency operating conditions. The analog is capable of determining the magnitude of many different effects that arise. In particular the ac lateral biasing in the base and its effect on such things as bottoming voltage can be measured.

In the particular transistor studied the analog showed that when delivering 0.5 watt at 100 Mc in class B, the effective base resistance would be reduced by a factor of about 2.3 and the effective collector series resistance increased by a factor of about 2.0 compared with the values for uniform current density.

These results have been generalized by making two additional assumptions in order to find the variation of these factors with transistor parameters and operating conditions. General expressions for base resistance and collector series resistance have been derived, and the value of the constants for typical class B operation have been determined. The analog could be used to find the constants for other modes of operation.

APPENDIX I

ANALYSIS OF THE MILLER ELEMENT (FIG. 4)

The impedance between base and earth is the sum of the input impedance of the transistor Z_{in} and the effective impedance of C_M.

If the analog transistor is operating well below its common emitter cutoff frequency

$$Z_{\text{in}} = \beta r_e + r_{b'}. \tag{20}$$

At low injection levels

$$r_e = \frac{kT}{q} \frac{1}{I_e}.$$

Therefore, providing I_e is not excessive, $\beta r_e \gg r_{b'}$ and most of the applied voltage appears across the emitter junction

$$\therefore\ I_e \doteq k_1 e^{-qV_{be}/kT},$$

and providing the transistor has a constant α,

$$I_{LA} \doteq k_1 e^{-qV_{be}/kT}. \tag{21}$$

If the analog frequency $(\omega_A/2\pi)$ is low enough, *i.e.*,

$$R_{LA} \ll \frac{1}{\omega_A C_M} \quad \text{and if} \quad R_{GA} \gg R_{LA},$$

$$V_c \doteq R_{LA} k_1 e^{-qV_{be}/kT}.$$

Differentiating

$$dV_c \doteq -\frac{q}{kT} R_{LA} k_1 e^{-qV_{be}/kT} dV_{be}.$$

From (21)

$$dV_c = -\frac{q}{kT} R_{LA} I_{LA} dV_{be}.$$

The equivalent capacitance C_{eq} between base and earth that would carry the same current as C_M is

$$C_{eq} = C_M \left[\frac{dV_{be} - dV_c}{dV_{be}}\right]$$

$$= C_M \left[1 + \frac{q}{kT} \cdot R_{LA} \cdot I_{LA}\right]. \tag{22}$$

Therefore providing $\beta r_e \gg 1/\omega_A C_{eq}$ the impedance between base and earth is determined by (22) where I_{LA} is given by (21).

The assumptions made to derive this relation are listed below for convenience.

1) Analog transistor operating well below its common emitter cutoff frequency.
2) Low injection levels.
3) $\beta r_e \gg r_{b'}$.
4) constant α.
5) $R_{LA} \ll 1/\omega_A C_M$.
6) $\beta r_e \gg 1/\omega_A C_{eq}$.

APPENDIX II

The relation between the current through the collector to base-junction capacitance $I_{b'c'}$ and that through the analog-diode capacitance I_{cA} is considered here.

Assume the two capacitances are the same function of voltage, and voltages of the same waveform are applied to each.

$$\therefore\ C_{b'c'} = f_1(V) \quad\text{and}\quad C_A = \frac{C_0}{C_T} f_1(V),$$

where C_0 and C_T are the capacitances of the diode and transistor junctions with a particular dc bias applied to both.

$$V_{b'c} = f_2(t) \quad\text{and}\quad V_A = f_2\left(t\frac{\omega_A}{\omega}\right).$$

The current through both junctions is

$$I = \frac{d}{dt}(C\cdot V) = C\frac{dV}{dt} + V\frac{dC}{dt}\cdot$$

Writing $f_1(f_2(t))$ as $f_3(t)$

$$I_{b'c'} = f_3(t)\frac{d}{dt}(f_2(t)) + f_2(t)\frac{d}{dt}(f_3(t)), \tag{23}$$

$$\begin{aligned} I_{cA} &= \frac{C_0}{C_T} f_3\left(t\frac{\omega_A}{\omega}\right)\frac{d}{dt}\left(f_2\left(t\frac{\omega_A}{\omega}\right)\right) \\ &\quad + \frac{C_0}{C_T} f_2\left(t\frac{\omega_A}{\omega}\right)\frac{d}{dt}\left(f_3\left(t\frac{\omega_A}{\omega}\right)\right) \\ &= \frac{C_0\omega_A}{C_T\omega}\left[f_3(z)\frac{d}{dz}(f_2(z)) + f_2(z)\frac{d}{dz}(f_3(z))\right], \end{aligned} \tag{24}$$

where

$$z = \frac{\omega_A}{\omega} t.$$

Eqs. (23) and (24) show that at any corresponding points on the analog and transistor-voltage waveforms the ratio of currents is given by

$$\frac{I_{cA}}{I_{b'c'}} = \frac{C_0\omega_A}{C_T\omega} = \frac{C_A\omega_A}{C_{b'c'}\omega},$$

providing C_A and $C_{b'c'}$ are the values of the two capacitances when the same voltage is applied to each.

APPENDIX III

Suppose that transistor A has a certain distributed base resistance under the emitter ρ_b/Z per unit length in the y direction [(assumption 7), Section I]. Suppose that the base current has some arbitrary waveform

$$I_{b(A)} = I_0(\omega t), \tag{25a}$$

and that the corresponding emitter current per unit length in the y direction is

$$i_{e(A)} = i_0 f(y) g(y, \omega t), \tag{25b}$$

where y is the distance from the edge of the emitter nearest the base contact.

Consider transistor B whose distributed base resistance per unit length in the y direction is p times, and whose f_1 is l times that of A. Suppose that it is required to pass a collector current a times that of A at m times the frequency. In the frequency range of assumption 3) $I_c \propto I_b(\omega_1/\omega)$,

$$\therefore\ I_{b(B)} = \frac{am}{l} I_0(m\omega t) \tag{26a}$$

It will be shown by induction that the emitter current per unit length is similar to that in A the distance scale being compressed by a factor amp/l and the amplitude increased by a^2mp/l, *i.e.*,

$$i_{e(B)} = \frac{a^2mp}{l} i_0 f\left(\frac{amp}{l}\cdot y\right) g\left(\frac{amp}{l}\cdot y, m\omega t\right). \tag{26b}$$

Consider the emitter to base junctions of the transistor to be divided into strips, those of transistor A having a width δy and those of B a width $(q/amp)\delta y$. Considering the rth elements in each case,

$$\frac{i_{e(B)r}}{i_{e(A)r}} = \frac{a^2mp}{l}\,\frac{g(r\delta y, m\omega t)}{g(r\delta y, \omega t)}\cdot$$

However, the reactance X of the storage capacitance of an element is given by

$$X = \frac{\omega_1}{\omega} r_e = \frac{\omega_1 kT}{\omega q i_e \delta y}\cdot$$

Therefore, considering corresponding points in the cycle in each case the ratio of the reactances of the rth elements is

$$\frac{X_{(B)r}}{X_{(A)r}} = \frac{l}{m}\cdot\frac{l}{a^2mp}\,\frac{amp}{l} = \frac{l}{am}\cdot$$

However the ratio of the resistances R of the elements is proportional to their width and the resistance per unit length.

$$\therefore\ \frac{R_{(B)r}}{R_{(A)r}} = \frac{l}{amp}\cdot p = \frac{l}{am},$$

$$\therefore\ \frac{X_{(B)r}}{R_{(B)r}} = \frac{X_{(A)r}}{R_{(A)r}}\cdot$$

Therefore all corresponding elements behave in the same way until the end of the line is reached at which the currents are negligible [assumption 2), Section VII]. Therefore, (*25*) follows from (*26*).

This means that the current distribution in all transistors can be found from a normalized distribution by

1) Multiplying the density by

$$\left(I_{cpk}{}^2\frac{\omega}{\omega_1}\,\frac{\rho_b}{Z}\right).$$

2) Cramping the scale in the y direction by a factor

$$\left(I_{cpk}\frac{\omega}{\omega_1}\,\frac{\rho_b}{Z}\right),$$

where I_{cpk} is the peak collector current.

APPENDIX IV

LIST OF SYMBOLS

β = common emitter current amplification.

β_0 = low-frequency value of β.

$\delta C_{b'e}$ = elemental emitter base capacitance in the transistor.

C_{eq} = analog capacitance representing $\delta C_{b'e}$.

δC_{se} = elemental diffusion capacitance in the transistor.

δC_{te} = elemental emitter depletion capacitance in the transistor.

C_M = Miller capacitance (Fig. 4) in the analog, representing δC_{te}.

$\delta C_{b'c'}$ = elemental capacitance of collector depletion layer under the emitter contact. $C_{b'c'}$ is the sum of all the $\delta C_{b'c'}$.

$\delta C_{bc''}$ = elemental capacitance of collector depletion layer not under the emitter contact. $C_{bc''}$ is the sum of all the $\delta C_{bc''}$.

C_A = the capacitance of the reverse biased diode used to represent $C_{b'c'}$ and $C_{bc''}$.

f_β = frequency at which $|\beta|$ is 3 db down on β_0.

f_1 = frequency at which $|\beta| = 1$.

δI = current from the elemental current generators in the common emitter hybrid π circuit.

I_{GA} = current through the resistance R_{GA} representing δI.

$I_{b'c'}$ = current through the collector depletion capacitance $C_{b'c'}$ in the transistor.

$I_{b'c'A}$ = current fed to the base of the analog, representing $I_{b'c'}$.

I_{LA} = current flowing in the resistance R_{LA} in the analog.

ρ_b = sheet resistance of the base layer of the transistor.

ρ_c = resistivity of the collector body.

R_A = sheet resistance of the graphite coated sheet in the analog representing the collector body.

δr_b = elemental resistance of the base layer between the emitter and base contacts.

$\delta r_{b'}$ = elemental resistance of the base layer under the emitter contact.

$R_{b'A}$ = resistance elements in the analog representing δr_b and $\delta r_{b'}$.

$\sum_n R_{b_n'c}$ = parallel combination of resistances through which $I_{b'c'A}$ flows to the base in the analog.

$\sum_n R_{c_n'c}$ = parallel combination of the resistances through which the current corresponding to $I_{b'c'}$ flows to the collector in the analog.

R_{LA} = load resistance of the Miller elements (Fig. 4).

$r_{cc'}$ = collector series resistance as defined in Section III-C-2).

ω_l = reciprocal of the base transit time.

ω = frequency at which the transitor operates.

ω_A = frequency at which the analog operates.

Z = length of the transistor in the z direction (Fig. 1).

Only those symbols which are used frequently are defined in this list.

ACKNOWLEDGMENT

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REFERENCES

[1] N. H. Fletcher, "Self-bias cutoff effect in power transistors," PROC. IRE (Correspondence), vol. 43, p. 1669; November, 1955.

[2] E. Wolfendale, Ed., "The Junction Transistor and Its Applications," Heywood and Co., Ltd., London, Eng.; 1958.

[3] J. Te Winkel, "Drift transistor," *Electronics and Radio Engineer*, vol. 36 (New Series), pp. 280–288; August, 1959.

[4] L. G. Cripps, "Transistor HF parameter f_1," *Electronics and Radio Engineer*, vol. 36 (New Series), pp. 341–346; September, 1959.

[5] J. M. Miller, "The dependence of the amplification constant and internal plate circuit resistance of a three-electrode vacuum tube upon the structural dimensions," PROC. IRE, vol. 8, pp. 64–73; February, 1920.

Parametric Behavior of an Ideal Two-Frequency Varactor*

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***Summary*—The Manley-Rowe equations for a lossless varactor network yield little information for the two-frequency case. With the assumption of a linear signal termination, it is shown that the ratio of signal and pump frequencies must be $M/2$, where M is an integer. For M odd, a regenerative network is possible at the signal frequency. For M even, the network is possibly regenerative or nonregenerative. Other frequency ratios are not possible; frequency division is not possible except for regenerative division by 2. The effects of strong pumping are predictable if the varactor characteristic is given.**

INTRODUCTION

THE simplest varactor networks exchange power at only two frequencies. The first is the frequency of the energy source, or pump. The second is the frequency of the useful output, or signal. Two-frequency circuits are used as parametric oscillators, amplifiers, or harmonic generators, and the literature is abundant in applications. The theory, however, is incomplete. The usual analysis assumes a small signal and neglects high-order nonlinearity of the varactor characteristic.[1] As a result, possible operating modes have been overlooked; the impossibility of certain modes is not clearly appreciated; and the effects of strong pumping are poorly understood. The object of this paper is a two-frequency theory of the varactor, ideal in the assumption that energy is dissipated only in the terminations of the varactor network. The theory accommodates a pump and signal of any amplitude, makes use of the complete varactor characteristic, and predicts all possible power conversions for a linear signal termination.

THE PROBLEM

A varactor is an element whose displacement q (charge, flux linkage) is a nonlinear function of the applied force v (voltage, current), so that

$$q = F(v). \tag{1}$$

The varactor is ideal if the function $F(v)$ is single-valued. If the remaining properties of $F(v)$ permit series expansion, then

$$q = \sum_{n=1}^{\infty} C_n v^n, \tag{2}$$

where the constants are

$$C_n = \frac{1}{n!}\cdot\frac{d^nF}{dv^n}, \qquad v = 0. \tag{3}$$

Let v_p be the pumping force and v_s the signal force,[2] so that

$$v = v_p + v_s = E_p \cos pt + E_s \cos (st + \phi). \tag{4}$$

Let $W_p = \langle w_p \rangle$ be the average power absorbed from the pump, and $W_s = \langle w_s \rangle$ the average power absorbed from the signal. Because of the ideal assumption, the average power is conserved, and $W_p + W_s = 0$. Finally,

$$w_s = v_s(dq/dt) = \sum_{n=1}^{\infty} C_n v_s d(v^n)/dt. \tag{5}$$

The problem is to determine the constraints on p, s, ϕ, and the C_n for which $W_s \neq 0$.

NECESSARY CONDITIONS

The Manley-Rowe equations [3] for an ideal varactor are

$$\sum_{m=0}^{\infty} \sum_{n=-\infty}^{\infty} \frac{mW_{m,n}}{mf_1 + nf_0} = 0$$

$$\sum_{m=-\infty}^{\infty} \sum_{n=0}^{\infty} \frac{nW_{m,n}}{mf_1 + nf_0} = 0, \tag{6}$$

and it is useful to ask whether or not these equations illuminate the desired constraints. Each frequency of the system has the form $(mf_1 + nf_0)$, and $W_{m,n}$ is the average power absorbed at that frequency, m and n being integers. Frequencies f_1 and f_0 are reference frequencies only; it is not generally necessary that either f_1 or f_0 represent a system frequency. Specifically, for the two-frequency case, it is not sufficient to approach the Manley-Rowe equations by defining $f_1 = p/2\pi$, $mf_1 = s/2\pi$, as has been done, for this begs the question by assuming $s/p = m$. Instead, choose four integers a, b, c, d. Let $(af_1 + bf_0) = p/2\pi$, $(cf_1 + df_0) = s/2\pi$. Then,

$$\frac{aW_p}{p} + \frac{cW_s}{s} = 0,$$

$$\frac{bW_p}{p} + \frac{dW_s}{s} = 0, \tag{7}$$

and since $W_p + W_s \equiv 0$,

$$\frac{s}{p} = \frac{c}{a} = \frac{d}{b} = \frac{y}{z}. \tag{8}$$

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[1] There are exceptions. Page [1] gives general results for a varactor specified by $q = Cv^N$. Leeson and Weinreb [2] demonstrate that Nth-order harmonic generation requires an Nth-order varactor curvature.

[2] The network can be chosen, using filters as in Manley and Rowe [3] for example, so that the total force on the varactor is a finite sum of sinusoidal forces.

where y and z are two integers with no common factor other than unity.

It is important to remember that the Manley-Rowe equations are necessary but not sufficient conditions for the operation of a parametric network [4]. Each is satisfied if all $W_{m,n}=0$. A network that satisfies the equations can be proved operative only by showing that at least one $W_{m,n}\neq 0$. Thus, the Manley-Rowe equations show that a two-frequency varactor network is not impossible if the ratio of signal and pump frequencies is a rational number. The equations do not show which ratios permit operation.

ANALYSIS

In combination with the binomial theorem, the identity $2\cos x \equiv \exp(jx)+\exp(-jx)$ transforms (5) to

$$w_s = \sum_{n=1}^{\infty}\sum_{k=0}^{n}\sum_{a=0}^{k}\sum_{b=0}^{n-k}\binom{n}{k}\binom{k}{a}\binom{n-k}{b}(1/2)^{n+1}C_nE_p{}^kE_s{}^{n-k-1}$$
$$\cdot\{(2a-k)p+(2b-n+k)s\}\{\exp j[(2b-n+k)\phi+(\pi/2)]\}$$
$$\cdot\{\exp j[(2a-k)pt+(2b-n+k+1)st+\phi]$$
$$+\exp j[(2a-k)pt+(2b-n+k-1)st-\phi]\} \quad (9)$$

where

$$\binom{a}{b} = \frac{a!}{b!(a-b)!}\,. \quad (10)$$

$W_s=\langle w_s\rangle$ is nonzero only if at least one of the exponents within the last bracket of at least one term of this sum (9) is constant.

Each term is also proportional to $E_s{}^{n-k+1}$. If the signal termination of the network is linear, the power dissipated in the termination can be proportional to the first or second power of E_s and to no other. (This conclusion follows Ohm's Law. Let I_s be the amplitude of the current in a linear termination, and F_p the termination power factor. The average dissipation is $-W_s=(F_p/2)E_sI_s$. Now, I_s is either independent of E_s or it is not. If it is independent, then the dissipation is proportional to the first power of E_s. If I_s is not independent, then $-W_s=(F_p/2)E_s{}^2/Z$, where Z is the termination impedance, and the dissipation is proportional to the second power of E_s. Ohm's Law permits no formula for dissipation as a function of E_s other than these two. For a given varactor and a given E_p, the terms of (9) are proportional to integral powers of E_s alone, and one or more of these must match either the first or second of the dissipations above.) Thus, $(n-k+1)=1$, 2. In the first, nonregenerative case, the network is a stable source of current at the signal frequency. In the second, regenerative case, the network provides a negative conductance at the signal frequency. The words nonregenerative and regenerative denote whether or not the network is generally stable with variation of the signal termination impedance.

The Nonregenerative Case

For $(n-k+1)=1$, $k=n$, and $b=0$. Eq. (9) becomes

$$w_s = \sum_{n=1}^{\infty}\sum_{a=0}^{n}\binom{n}{a}(1/2)^{n+1}C_nE_p{}^nE_s(2a-n)p$$
$$\cdot\{\exp j[(2a-n)pt+st+(\pi/2)+\phi]$$
$$+\exp j[(2a-n)pt-st+(\pi/2)-\phi]\}. \quad (11)$$

If $W_s\neq 0$, then $(2a-n)p+s=0$, or $(2a-n)p-s=0$, and s/p is an integer. Let $s/p=N$. Then $a=(n-N)/2$, or $a=(n+N)/2$. Since $0\leq a\leq n$, $n\geq N$, and n is odd or even as N is odd or even. Thus,

$$W_s = \sum_{\substack{n=N\\ \left(\substack{\text{all odd}\\ \text{or even}}\right)}}^{\infty}\left(\frac{n}{\frac{n-N}{2}}\right)(1/2)^nsC_nE_p{}^nE_s\sin\phi,\quad s/p=N \quad (12)$$

which is the average signal power for nonregenerative operation. A power output ($W_s<0$) is obtained only at a harmonic of the pump frequency.

The Regenerative Case

For $(n-k+1)=2$, $k=n-1$, and $b=0$, 1. Eq. (9) becomes

$$w_s = \sum_{n=1}^{\infty}\sum_{a=0}^{n-1}\binom{n-1}{a}(1/2)^{n+1}nC_nE_p{}^{n-1}E_s{}^2$$
$$\cdot\{2[2a-n+1]p\exp j[(2a-n+1)pt+(\pi/2)]$$
$$+[(2a-n+1)p-s]$$
$$\cdot\exp j[(2a-n+1)pt-2st+(\pi/2)-2\phi]$$
$$+[(2a-n+1)p+s]$$
$$\cdot\exp j[(2a-n+1)pt+2st+(\pi/2)+2\phi]\}. \quad (13)$$

If $W_s\neq 0$, then

$$(2a-n+1)p-2s=0,$$

or

$$(2a-n+1)p+2s=0,$$

and s/p is an integer divided by 2. Let $s/p=U/2$. Then $a=(n-1+U)/2$, or $a=(n-1-U)/2$. Since $0\leq a\leq n-1$, $n-1\geq U$, and n is odd or even as U is even or odd. Thus,

$$W_s = \sum_{\substack{n=U+1\\ \left(\substack{\text{all odd}\\ \text{or even}}\right)}}^{\infty}\left(\frac{n-1}{\frac{n-1-U}{2}}\right)(1/2)^nnsC_nE_p{}^{n-1}E_s{}^2\sin 2\phi,$$
$$s/p=U/2 \quad (14)$$

which is the average signal power for regenerative operation. The output ($W_s<0$) can be developed either at a harmonic of the pump frequency (U even) or at an odd multiple of one-half the pump frequency (U odd).

DISCUSSION

It is clear that if the signal termination is linear, the ratio of signal and pump frequencies for the two-frequency varactor must be $M/2$, where M is an integer. When M is even, the mode of operation can be either nonregenerative as shown by (12) or regenerative as shown by (14). When M is odd, operation can be regenerative only. The only possible subharmonic generation occurs when M is unity.

Noiseless harmonic generation is a recurring design goal, and varactor frequency multipliers are often relatively efficient. There is a possible difficulty, however, since the harmonic output can consist of both regenerative and nonregenerative components. The two components differ in phase by either $\pi/4$ or $3\pi/4$ radians. Ordinarily the nonregenerative component will predominate, since it depends importantly on varactor constants C_n of lower order. But if sufficient regeneration exists to amplify noise in the network, the sum of the two components will be a signal that appears to be noise-modulated in both phase and amplitude. For odd-harmonic generation, it may not be possible to avoid this difficulty entirely, other than by seeking a varactor with small C_n for $n > N$. For generating even harmonics, it is always possible to pair like varactors in a balanced network such that $C_n = 0$ for n odd, thereby eliminating regeneration at all even harmonics.

It is possible to generate oscillations or to amplify regeneratively at signal frequencies that are odd multiples of one-half the pump frequency [5]. This kind of operation is attractive because of the relatively low pump frequency required. However, such applications want a varactor for which C_{U+1} is appropriately large, and as U becomes large, the negative conductance becomes more sensitive to pump amplitude. Operation at large s/p ratios therefore requires more precise control of pump amplitude, which may be a disadvantage.

If the varactor characteristic is completely specified, the effects of strong pumping are predictable. Eqs. (12) and (14) can be written

$$W_s = (E_s \sin \phi) \sum_{n=N}^{\infty} \binom{n}{\dfrac{n-N}{2}} sC_n(E_p/2)^n$$

$$= (E_s \sin \phi) I(E_p) \tag{15}$$

$$W_s = (E_s^2/2)(\sin 2\phi) \sum_{n=U+1}^{\infty} \binom{n-1}{\dfrac{n-1-U}{2}} nsC_n(E_p/2)^{n-1}$$

$$= (E_s^2/2)(\sin 2\phi) G(E_p), \tag{16}$$

respectively, where the polynomial $I(E_p)$ is a velocity (current), and the polynomial $G(E_p)$ a conductance. In general, these polynomials can be expected to display one or more maxima in E_p, corresponding to limiting behavior as the pump amplitude is increased. Note that $I(E_p)$ and $G(E_p)$ are independent of E_s, an independence which implies unlimited power capability for the varactor network. This result follows from the assumption of zero network dissipation. In practice, even if the varactor were lossless, it would not be possible to provide lossless frequency-selective networks for the pump and signal terminations; and losses in these elements would limit the available signal power and the power absorbed from the pump.

CONCLUSION

For a two-frequency varactor network with a linear signal termination, the ratio of signal and pump frequencies is $M/2$, where M is an integer. For M odd, the network is possibly regenerative at the signal frequency. For M even, the network is possibly either regenerative or nonregenerative, or both, at the signal frequency.

Operation is not possible for other frequency ratios. A nonregenerative frequency divider (subharmonic generator) is not possible. A regenerative frequency divider is not possible, except for division by 2.

Frequency multipliers (harmonic generators) may be both regenerative and nonregenerative. A strictly nonregenerative frequency multiplier is possible for even harmonics. A strictly nonregenerative frequency multiplier may not be possible for odd harmonics.

Regenerative amplifiers and oscillators are possible at signal frequencies that are odd multiples of one-half the pump frequency. The oscillation may have either of two phases separated by a half-cycle of the oscillation.

The effects of strong pumping are specified by a polynomial in the pump amplitude using constants of the power series expansion of the varactor characteristic.

REFERENCES

[1] C. H. Page, "Frequency conversion with nonlinear reactance," *J. Res. NBS*, vol. 58, pp. 227–236; May, 1957.

[2] D. B. Leeson and S. Weinreb, "Frequency multiplication with nonlinear capacitors—a circuit analysis," PROC. IRE, vol. 47, pp. 2076–2084; December, 1959.

[3] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements—part I. General energy relations," PROC. IRE, vol. 44, pp. 904–913; July, 1956.

[4] P. Penfield, Jr., "Frequency-Power Formulas," John Wiley and Sons, Inc., New York, N. Y.; 1960.

[5] G. F. Montgomery, "Parametric amplification with a low-frequency pump," PROC. IRE, vol. 49, pp. 1214–1215; July, 1961.

Correspondence

Comments on "A Synthesis Procedure for an n-Port Network"*

The above note[1] by Hazony and Nain produces the following comments:

1) In the literature the word "port" usually refers to a pair of terminals. Furthermore, the classical result of Bott and Duffin applies to one-port networks. Thus, there is an inconsistency in the first paragraph of the note by Hazony and Nain.

2) The authors do not mention ideal transformers. However, the Bott and Duffin procedure did indeed eliminate the ideal transformer from the synthesis of a positive real driving point function. It is inferred from the first paragraph of their note that Hazony and Nain have succeeded in finding necessary and sufficient conditions for the realizability of an RLC n port without ideal transformers.

If this inference is correct, the authors are to be congratulated. Unfortunately, congratulations will only be minimal unless the achievement is sufficiently advertised by discussing the role of ideal transformers.

3) Hazony and Nain exhibit their conditions 1)–4) as sufficient for their n-port realization. These conditions are also necessary.

However, these conditions are not precisely equivalent to the statement that the matrix in question is a positive real matrix. In particular, their condition 3) is necessary for the matrix to be positive real, but when coupled with their conditions 1), 2), and 4), the entire set of conditions is not sufficient.

Since positive real matrices have been predominant in the related literature, it is of interest to relate 1)–4) to the usual requirements for a matrix to be positive real.

4) This author is sure that the excellent note by Hazony and Nain has stimulated the curiosity of all people interested in n-port synthesis. A discussion of the three items cited above will clarify the contribution made by their note.

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* Received by the IRE, September 15, 1961.

[1] D. Hazony and H. J. Nain, PROC. IRE (Correspondence), vol. 49, pp. 1431–1432; September, 1961.

Authors' Comments[2]

Prof. Slepian in his note points out correctly that the term, *one-port* synthesis, should have been used in conjunction with the Bott-Duffin procedure.

With reference to comment No. 3, the constraint 4) in the original note should have read: $|\mathrm{Re}[Z(s)]| \geq 0$ for Re $s=0$. Further, the condition that the elements of $[Z(s)]$ are rational functions of s with real coefficients should have been added to this constraint. This constraint would then be equivalent to the usual requirement for a matrix function to be positive real.

In comment No. 2 the authors regret to say that this method does not guarantee synthesis without the use of ideal transformers for $n \geq 2$. However, a necessary condition for a transformerless synthesis by the method is, for an example of two ports, that $[Z(k)]([Z(s)]=$a minimum matric function) be relizable as a transformerless two ports.[3]

In view of some personal queries that the authors have received, they also wish to restate Theorems II and III as follows:

Theorem II: $[\zeta_1]$ and $[\zeta_2]$ as defined by (1) are also *realizable* matrix functions, and they satisfy the constraint $|Z(k)|^2 \times |\zeta_1| \times |\zeta_2| = 1$.

Theorem III: If $[Z(s)]$ is a "realizable," minimum resistive, matrix function $(\mathrm{Re}[Z(s)] = [0]$ at $s = j\omega_0)$, then it is possible to find a constant $k(k>0)$ such that the elements of $[\zeta_1]$ and $[\zeta_2]^{-1}$ have either the same j axis zeros or poles. This k is obtained by solving either $|Z_1(j\omega_0)| = 0$ or $|Z_2(j\omega_0)| = 0$ in (1).

The authors gratefully acknowledge Prof. Slepian's interest in their note and regret the inconvenience caused because of these corrections.

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[2] Received by the IRE, September 27, 1961.

[3] Indeed, the method seems to promote transformerless solutions; for example, the authors were able to synthesize the following impedance matrix

$$[Z(s)] = \begin{bmatrix} \dfrac{s^2+s+2}{2s^2+s+1} & \dfrac{1}{2}\,\dfrac{s^2+3s+4}{4s^2+3s+1} \\[2ex] \dfrac{1}{2}\,\dfrac{s^2+3s+4}{4s^2+3s+1} & \dfrac{s^2+2s+3}{3s^2+2s+1} \end{bmatrix}$$

without transformers.

A Cryotron Memory Cell*

Haynes of IBM has recently described a cryotron memory which will perform a number of logical operations such as shifting within the memory and simultaneous access to any three words—two for reading and one for writing.[1] The memory cell which he proposed is constructed with five cryotrons and three superconductive loops. This note describes a memory cell which will accomplish similar functions with three cryotrons and one superconductive loop.

WRITE-IN

The circuit diagram of the new cell is shown in Fig. 1. It contains a persistent current loop *abcd* in which a clockwise persistent current represents the storage of a ONE and no persistent current represents a ZERO. A persistent current is established in loop *abcd* by simultaneously energizing the "write" and "enable" lines. Since current in the enable line makes cryotron E resistive, the write current takes the path *abcd*. If the enable current is removed before the write current is removed, a persistent current can be established in the loop. The magnitude of the persistent current is chosen so that it is equal to $1/2\ I_{CRIT}$, where I_{CRIT} is the critical control current for cryotrons R_1 and R_2.

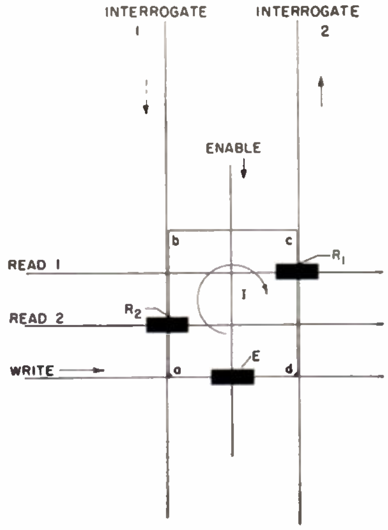

Fig. 1—Three cryotron memory cell.

READ-OUT

The memory cell has two interrogate lines and two read lines and the stored information can be sensed on either the read 1 or the read 2 output, depending upon which interrogate line is energized.

The two interrogate lines are deposited on top of side *ab* and side *cd* of the memory cell. Output cryotrons R_1 and R_2 are therefore double-control cryotrons; their gates are quenched when the sum of the interrogate current and the current in the loop *abcd* is equal to I_{CRIT}.

The cell geometry is arranged so that when the interrogate 1 line is pulsed with a current equal to I_{CRIT} then a current of $I_{CRIT}/2$ is induced in the loop. The induced current $(I_{CRIT}/2)$ plus the persistent current $(I_{CRIT}/2)$ add so that cryotron R_1 is made resistive. If there is no persistent current then there would be only an induced current in the loop (equal to $I_{CRIT}/2$) and R_1 would remain superconductive.

It should be noted that gate R_1 is on the opposite side of the memory cell from the interrogate 1 line. Gate R_2 is not quenched

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* Received by the IRE, October 25, 1961.

[1] M. K. Haynes, "Cryotron storage, arithmetic and logical circuits," *Solid-State Electronics*, vol. 1, pp. 399–408; September, 1960.

when the interrogate 1 line is pulsed because the current in the interrogate 1 line is opposing the current in the persistent current loop. Thus the interrogate 1 input will control the R_1 output.

TRIPLE ACCESS MEMORY

Fig. 2 is a circuit diagram of a memory containing three 3-bit words. Any two of these words can be interrogated and information written into the third one simultaneously. In the example shown, words 1 and 2 are being interrogated so that their contents will be sensed on the read 1 and read 2 outputs, respectively.

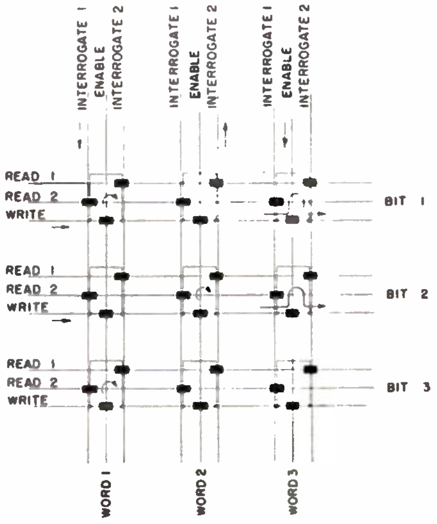

Fig. 2—Circuit diagram of memory.

To write 110 into word 3, current is applied to the bit 1 and bit 2 write lines. This current will preferentially flow through the enable gates of the memory cells in words 1 and 2 rather than the higher inductance branch that crosses the R_1 and R_2 cryotrons. Thus the write current does not disturb the output cryotrons in the memory cells through which it passes. In word 3, however, the enable gates are resistive and the write current is forced through the higher inductance branch of these memory cells so that the word 110 can be written into this location. Since the write current is equal to 1/2 I_{CRIT}, it is not large enough to give a false output in the R_1 or R_2 gates of word 3.

CIRCUIT LAYOUT

A single memory cell measures 0.025 by 0.050 inch, suggesting a packing density of 600 bits per square inch. The circuit layout for this cell is shown in Fig. 3. The dimensions are based on a minimum line width of 0.005 inch and a similar spacing between lines. This size is compatible with present masking techniques and, if it seemed desirable to do so, could be reduced by a factor of two to five.

SUMMARY

The memory cell described is capable of a variety of functions such as triple access and register-to-register transfer within the memory.[2] It also appears to have minimum complexity for the number of functions which can be performed. A circuit containing four memory cells is presently being tested in order to determine its operating characteristics.

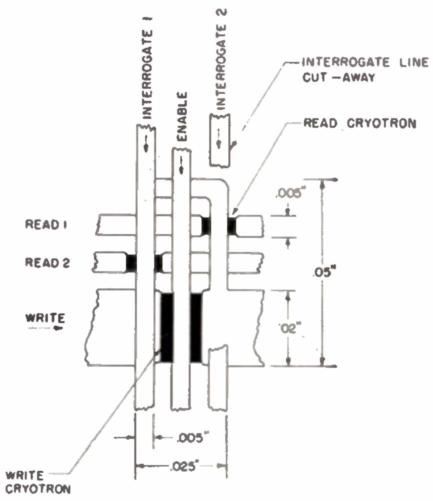

Fig. 3—Layout of memory.

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[2] *Ibid.*, p. 402.

Diffused Silicon Mesa Diodes for Use in Refrigerated Parametric Amplifiers*

Previous experiments with parametric amplifiers [1, 2] have indicated that the excess noise temperature of the amplifier can be substantially reduced by refrigeration of the varactor diode. The noise figure was found to decrease steadily as the diode temperature was decreased from room temperature to temperatures near 100°K. This was found to be the case for both diffused germanium mesa diodes [1] and for gallium-arsenide point-contact diodes [2]. However, previous results (to be discussed below) obtained by authors using refrigerated diffused silicon mesa diodes were much less favorable, indicating either little improvement in, or an increase in noise figure with cooling. As discussed below, the prediction of the low-temperature performance of a diffused silicon mesa diode is not entirely straightforward. However, at the outset, operation of such a diode (suitably doped) to temperatures as low as 100°K without degradation of the quality factor seems feasible. Accordingly, further investigation of the noise figure reduction obtainable with refrigeration is indicated.

Results are reported in this letter which indicate that a substantial decrease in noise figure can be obtained with refrigeration of a diffused silicon mesa varactor to temperatures near 100°K (a temperature range including temperatures accessible by thermoelectric means). Further, the noise figure obtained at room temperature using the best silicon unit measured is comparable with that of an average gallium-arsenide unit. It is felt that these results are significant, in view of the proven stability and ease of fabrication of diffused silicon devices.

One example of previous pessimistic results using refrigerated diffused silicon mesa diodes is shown in Fig. 1. The experiments were performed using 5.84 Gc signal and 11.7 Gc pump frequencies. The amplifier used for these experiments has been described in a previous report [1]. In Fig. 1, curves F and I_d are the measured noise figure and rectified diode current plotted as functions of diode temperature. The curve designated "Pump" indicates the pump power required to maintain the amplifier gain constant at 16 db. As the diode temperature was lowered, it was necessary to increase the pump power to maintain constant gain (in contrast to similar experiments with refrigerated germanium and gallium-arsenide diodes where pump power decreased or remained almost constant). The assumption that changes in pump power with diode temperature probably indicate a continuous decrease in the dynamic quality factor ($\tilde{Q}$) of the diode [3] seems to be borne out by the temperature dependence of noise figure indicated in Fig. 1. The values of noise figure measured at room temperature indicate that these diodes have substantially lower $\tilde{Q}$ values than do the germanium and gallium-arsenide diodes.

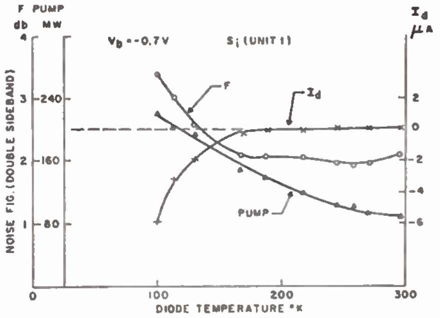

Fig. 1—Noise figure as a function of diode temperature.

Improvement in the $\tilde{Q}$ of the diode can be effected by reducing the mesa area. This results because the spreading resistance does not increase as rapidly with decreasing mesa diameter as the capacitance decreases, and the RC product of the wafer decreases with decreasing mesa diameter [5].

The prediction of the temperature dependence of $\tilde{Q}$ for a diffused mesa diode involves considerable complication. It is clear that the bulk series resistance in the wafer can be made almost temperature independ-

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* Received by the IRE November 16, 1961.

ent to 10°K by doping the silicon sufficiently. For example, the results of Morin and Maita [6] indicate only small changes in conductivity (σ) down to 10°K for p-type silicon doped to a level of about 1.5×10^{19} cm^{-3}. However, there is a heavily compensated impurity distribution near the space charge boundary of the diffused junction. The lowest conductances in this region are near the space charge boundaries, and the net carrier concentrations are probably lower at these boundaries than they are in the bulk semiconductor. The measurements of Morin and Maita indicate considerable differences in the temperature dependence of σ for uncompensated silicon for doping levels lower than 10^{19} cm^{-3}. Similar complicating changes in the temperature dependence of σ for compensated doping are probable.

In view of these uncertainties, the temperature dependence of the noise figure for a variety of diffused silicon varactor diodes with starting doping level in the degenerate region was investigated. Encouraging results have been obtained with a group of units fabricated using phosphorus diffusion into p-silicon of 0.001 ohm-cm.

Several diodes of this kind were first examined at room temperature, to determine the noise figure as a function of bias voltage. The best noise figure measured was 0.7 db at 5.84 Gc (double channel). The dynamic quality factor of this diode, on the basis of the data obtained, is better than 6.5, and the cutoff frequency is higher than 130 Gc at zero bias. The high $\bar{Q}$ and low-noise figure are believed to result from the combination of high doping level and very small mesa diameter. A mechanical failure prevented determination of noise temperature below 200°K. However, the noise temperature decreased steadily as the temperature decreased toward this value. Another similar diode was satisfactorily operated down to about 93°K, and the experimental data are plotted in Fig. 2. In this case, excess noise temperature (T_n) is plotted as a function of diode temperature. T_n decreased from 93°K to about 30°K as the diode was refrigerated from 296°K to 93°K. The pump power necessary to maintain the gain at 16 db decreased continuously down to 180°K and remained constant at lower temperatures. This probably indicates improvement of the diode $\bar{Q}$ with cooling in the higher temperature range.

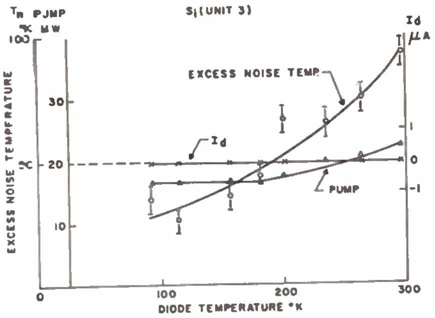

Fig. 2—Noise temperature as a function of diode temperature.

These results, admittedly limited, indicate that refrigeration of the diffused silicon varactor diode can decrease the noise figure of a varactor amplifier substantially. Further, the cutoff frequency of the diffused silicon mesa varactor can be well over 100 kMc at zero bias, a value which ensures a noise figure close to that of presently available gallium arsenide diodes.

ACKNOWLEDGMENT

The authors wish to acknowledge the assistance of R. C. Swenson and T. W. Mohr in fabrication of the diodes and in carrying out the noise figure measurements.

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REFERENCES

[1] M. Uenohara and A. E. Bakanowski, "Low-noise parametric amplifier using germanium p-n junction diode at 6 Kmc," PROC. IRE, vol. 47; pp. 2113–2114; December, 1959.

[2] M. Uenohara and W. M. Sharpless, "An extremely low-noise 6-kmc parametric amplifier using gallium arsenide point-contact diodes," PROC. IRE, vol. 47; pp. 2114–2115; December, 1959.

[3] K. Kurokawa and M. Uenohara, "Minimum noise figure of the variable capacitance amplifier," *Bell Sys. Tech. J.*, vol. 40, p. 695; May, 1961. If the time dependent junction capacitance is assumed of the form $C(t)=C_0-C\cos\omega_p t$, where ω_p is the pump frequency, then $\bar{Q}=Qf(\delta)$ where Q is the diode Q at the signal frequency and $\delta=|C|/C_0$. The nature of $f(\delta)$ depends on the assumptions made regarding terminating impedances for higher harmonics.

[4] A. E. Bakanowski, *et al.*, "Diffused silicon nonlinear capacitors," IRE TRANS. ON ELECTRON DEVICES, vol. ED-6, pp. 384–390; October, 1959.

[5] N. G. Cranna and A. E. Bakanowski, "Series losses in diffused silicon varactor diodes," ch. 2, 12th Interim Rept., "Microwave Solid State Devices," Signal Corps Contract DA-36-039 SC-73224; January 15, 1960.

[6] F. J. Morin and J. P. Maita, "Electrical Properties of Silicon Containing Arsenic and Boron," *Phys. Rev.*, vol. 96, pp. 25–35; October 1, 1954.

Reverse-Current Anomaly in Germanium Diodes*

Experiments have recently been performed at 156 Mc with experimental germanium varactors and a reverse-current anomaly was observed. In April, 1960, Hefni[1] reported a similar anomaly at 4 kMc in a silicon mesa-type varactor and proposed that minority carrier lifetime was responsible. Measurements on nearly 100 germanium diodes, however, suggest that this may not be the complete explanation.

The germanium varactors had normal static characteristics; however, when RF driving power was superimposed upon the dc bias voltage, a current much larger than the normal reverse saturation current flowed through the diode and an area of negative resistance was observed. Avalanche breakdown current was eliminated as a possible source of this large reverse current since the effect was observed at bias voltages much less than the avalanche breakdown voltage.

The diode I-V characteristics were displayed on an oscilloscope while operating in a 156- to 312-Mc frequency doubler circuit with the bias voltage across the diode swept by a sawtooth voltage of approximately 100 cps. A few typical displays are shown in Fig. 1. Figs. 1(a) and 1(b) are of a normal diode which did not exhibit the anomaly. Figs. 1(c) and 1(d), however, show the I-V characteristics of a diode exhibiting this reverse current anomaly. This effect was observed for RF power inputs as low as 14 μw.

The minority carrier lifetime was measured to determine if any correlation existed between lifetime and the negative resistance effect. Minority carrier lifetime was measured on a Tektronic oscilloscope equipped with a Tektronic diode recovery tester plug-in adapter. For most diodes, the trace was as shown in Fig. 2(a). This was as expected. For others, however, the trace appeared as in Fig. 2(b) with a spike which, at times, was of sufficient magnitude to interfere with accurate lifetime measurement. This spike was later correlated with long lifetime and the negative resistance phenomenon.

The diodes tested may be separated into three basic device groups depending upon the treatment of the n-type germanium used in their fabrication. These groups are homogeneous-base devices, diffused-base devices, and homogeneous gold-doped devices.

The homogeneous-base devices, fabricated from n-type germanium of various resistivities, exhibited lifetimes from 13 to 356 nanoseconds. Of these, every diode which had a minority-carrier lifetime greater than 50 nsec also had a spike on the lifetime measurement display and exhibited the reverse current anomaly under RF drive. Those units with lifetimes less than 50 nsec had normal lifetime displays and did not exhibit the reverse current anomaly.

The gold-doped units were fabricated from n-type, gold-doped germanium with a resistivity of approximately 2 ohm-cm. These units were found to have normal lifetime characteristics with lifetimes in the order of 30 nsec and showed no evidence of the reverse-current anomaly.

The diffused-base units were fabricated from homogeneous n-type germanium which had a thin $n+$ layer diffused into the surface. Although these diodes had minority carrier lifetimes between 26 and 260 nsec, no spikes on the lifetime measurement display nor negative resistance effects under RF drive were observed. These results were contrary to those expected by the simple correlation with minority-carrier lifetime. Since, with the exception of the surface treatment, these units were identical to the homogeneous-base devices, these results suggest that the appearance of the reverse current anomaly and the spike on the lifetime display are dependent upon surface effects.

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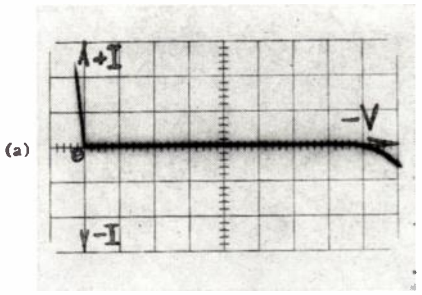
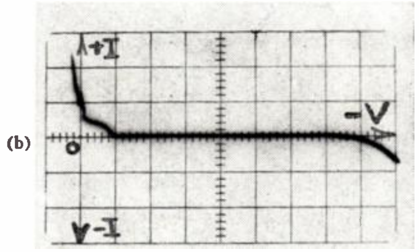
* Received by the IRE, September 21, 1961; revised manuscript received, October 27, 1961.

[1] I. Hefni, "Effect of minority-carriers on the Dynamic characteristic of parametric diodes," Electronic Engrg., vol. x, pp. 226–227; April, 1960.

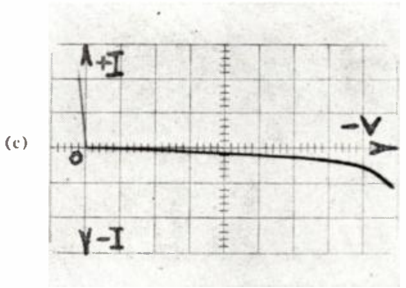


Diode No. 1:

(a) Reverse characteristics without RF driving power. Horizontal = −10 v/division. Vertical = 0.1 ma/division.

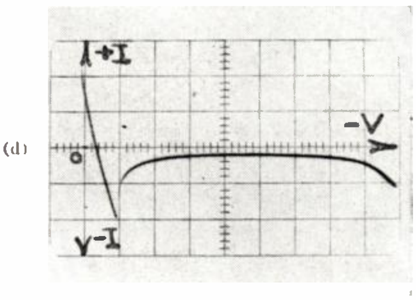


(b) Reverse characteristics with RF driving power = 35 mw. Horizontal = −10 v/division. Vertical = 0.1 ma/division.



Diode No. 2:

(c) Reverse characteristics without RF driving power. Horizontal = −10 v/division. Vertical = 0.1 ma/division.



(d) Reverse characteristics with RF driving power = 35 mw. Horizontal = −10 v/division. Vertical = 0.1 ma/division.

Fig. 1—Reverse characteristics of varactor diodes.

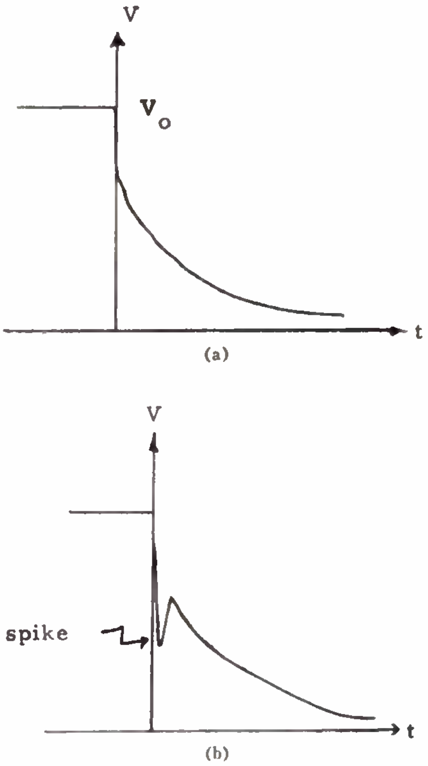


Fig. 2—Minority carrier lifetime measurement display.

This investigation has shown that, although minority-carrier lifetime is related to the reverse-current anomaly, it may not be the determining factor. Although the anomaly was not observed for any diodes with lifetimes less than 50 nsec, it was shown to be dependent on surface treatment for diodes with longer lifetimes. This would suggest that the reverse-current anomaly and the spike on the lifetime display depend upon surface effects in addition to the minority-carrier lifetime.

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Note on "Complementary Series"*

The question whether "Complementary Series"[1] of length 26 exist has been answered in the affirmative with the following example:

11100111010000010110010000
00011000101101010110010000

Marcel J. E. Golay
Technical University
Eindhoven, Netherlands

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* Received by the IRE, October 13, 1961.

[1] M. J. E. Golay, "Complementary series," IRE Trans. on Information Theory, vol. IT-7, pp. 82–87; April, 1961.

On the Resonant Frequency Modes of Ruby Optical Masers*

When the input radiation from a ruby optical maser is passed through a Fabry-Perot interferometer which is not crossed with a dispersing apparatus such as a spectrograph, a multitude of rings is observed as reported by several workers in the field.[1–3] In such a setup the various orders of the interferometer overlap. As a result one cannot extract sufficient information from these patterns.

In consequence, it was decided to pass the laser beam through a high-resolution spectrograph. The spectrograph chosen was the 40-foot MIT high-resolution instrument using a 10-inch Harrison and Stroke diffraction grating which has an apparatus line-width of only 0.02 wave numbers for 6943Å radiation. With this instrument, a detailed picture of the frequency modes can be obtained (see Fig. 1). Here, two exposures of the laser output, taken at different times, were recorded on the same photographic plate by moving the plate between exposures. A Kodak Type I-N spectroscopic plate was used. To facilitate the evaluation of this photograph, the reticule of an eyepiece was placed on the film after development. The upper scale shown in Fig. 1 represents distance in millimeters where one millimeter corresponds to 0.8 wave numbers.

It can be seen that the laser output consists of approximately 10 lines with a constant separation of 0.08 wave numbers between them. These lines represent the different axial Fabry-Perot interferometer modes excited within the ruby cavity itself. These appear at the wavelengths which satisfy (1).

$$n\frac{\lambda_0}{2} = (n+1)\frac{\lambda}{2} = \cdots = (n+k)\frac{\lambda_k}{2} = \cdots \qquad (1)$$

where

$$\frac{\lambda_k}{2} = \left(\frac{\lambda_0}{2} + k\frac{\Delta\lambda}{2}\right) \quad \text{when} \quad k \ll n.$$

These modes fall within the 16-wave-number line-width of the 6934Å fluorescence line. The separation of different lines, $\Delta\lambda/2$, corresponds to 0.08 wave numbers and may be accounted for by the length of the ruby rod. Furthermore, we find that each line has a line width of 0.02 wave numbers, but, since the instrument also has a line width of 0.02 wave numbers, we are limited by the resolution of the instrument. Therefore, to avoid this difficulty, this spectrograph should be used in conjunction with a high-resolving power Fabry-Perot interferometer.

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* Received by the IRE, November 20, 1961. This note was presented at the Electron Devices Research Conference, Troy, N. Y., June, 1961.

[1] R. J. Collins, et al., "Coherence narrowing, directionality, and relaxation oscillations in the light emission from ruby," *Phys. Rev.*, vol. 5, pp. 303–305; October, 1960.

[2] T. H. Maiman, *et al.*, "Stimulated optical emission in fluorescent solids II," *Phys. Rev.*, vol. 123, pp. 1151–1157.

[3] W. Kaiser, *et al.*, "Fluorescence and optical effects in CaF$_2$:Sm^{++}," *Phys. Rev.*, vol. 123, pp. 766–776; August, 1961.

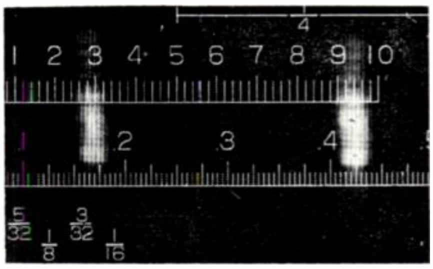

Fig. 1—The resonant frequency modes of ruby optical maser operated near threshold.

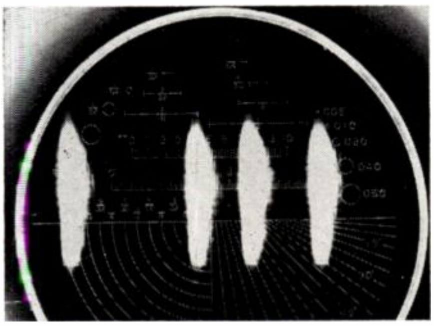

Fig. 2—Same as Fig. 1, but operated at higher input energy and showing four exposures.

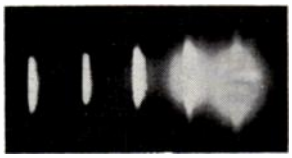

Fig. 3—Same as Fig. 1, but operated at still higher input energies and with successively increasing (*from left to right*) slit width of the spectrograph.

From these experiments, it is clear that the output consists of lines, as observed by the use of the spectrograph, and these are due to the different Fabry-Perot interferometer modes sustained within the ruby cavity itself. Fig. 1 now clarifies the origin of the numerous rings obtained by the Fabry-Perot interferometer. The number of modes shown on this pattern is dependent on the input energy. At higher input energies, more modes are excited, as is shown in Fig. 2. If all the modes within the fluorescence line width were excited, one would obtain some 400 of these lines. Fig. 3 illustrates a case where almost all of these lines have been excited.

For this picture, the slit of the spectrograph had been gradually opened for successive exposures from left to right resulting in an increasing loss in resolution. For instance, the exposure on the very right was obtained with a slit-width corresponding to 0.3 cm^{-1} resolution. The reason for successively opening the slit of the spectrograph wider and wider was to gain sufficient intensity and observe whether or not the weakly excited resonant modes falling under the tail of the fluorescence line shape were at all present. In view of the fact that the fluorescence output of the laser was not intense enough to be photographed, one concludes that the above exposure indicates the presence of all the frequency modes falling under the fluorescence line width. From such experiments it was estimated that, by controlling the threshold input electrical energy to better than 1/10th of 1 per cent, one should be able to obtain not only a single interferometer mode but also a single spike. Such single spikes have been observed. Since this interrelationship between frequency modes and spiking patterns is a very interesting one, further experiments are planned to study this effect in some detail.

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WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table below. The corrections reported have been arrived at 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, 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; a time advancement of 50 msec was made at 0000 UT on August 1, 1961.

WWV Frequency
With Respect to U. S. Frequency Standard

October 1961	Parts in 10^{10}†
1	−149.5
2	−149.4
3	−149.3
4	−149.4
5	−149.3
6	−149.3
7	−149.3
8	−149.2
9	−149.1
10‡	−149.0
11	−149.7
12	−149.6
13	−149.8
14	−150.0
15	−150.1
16	−150.2
17	−150.4
18	−150.6
19	−150.5
20	−150.6
21	−150.4
22	−150.5
23	−150.5
24	−150.5
25	−150.1
26	−150.1
27	−149.8
28	−149.8
29	−149.7
30	−150.0
31	−150.0

† A minus sign indicates that the broadcast frequency was low. The uncertainty associated with these values is $\pm 5 \times 10^{-11}$.
‡ Frequency adjusted -0.7×10^{-10} on October 10.

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Boulder, Colo.

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* Received by the IRE, November 17, 1961.

[1] Refer to "National Standards of Time and Frequency in The United States," Proc. IRE, vol. 48, pp. 105–106; January, 1960.

The Relationship Between Emitter Junction Recovery and "Wiggle" Effect*

An anomalous behavior of the saturated transistor collector waveform during turn-off has been observed previously and has been referred to as the "wiggle" effect.[1,2] Several attempts have been made to explain this phenomenon.[2,3] The purpose of this note is to report on a mechanism explaining this effect and to present supporting experimental evidence.

Curve (1) of Fig. 1 shows the collector waveform for a transistor which does not exhibit the "wiggle" effect; curve (2) shows the waveform when the "wiggle" effect is present. This effect may be explained by considering the emitter junction to become momentarily reverse biased upon application of the "off" pulse, and to subsequently become forward biased once again. Waveforms, characteristic of the base current and voltage in a common emitter circuit with base speed-up capacitance, are shown in Fig. 2. Waveforms V_{BE} and I_B have been observed experimentally. Curve $V_{B'E}$, rep-

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* Received by the IRE, November 13, 1961.

[1] R. Beaufoy and J. J. Sparkes, "The junction transistor as a charge controlled device," *ATE J.*, vol. 13, pp. 310–327; October, 1957.

[2] J. J. Sparkes, "A study of the charge control parameters of transistors," Proc. IRE, vol. 48, pp. 1696–1705; October, 1960.

[3] R. J. Bosselaers, "A simple method for compensation of the wiggle effect," *The Solid State J.*, vol. 2, pp. 32–35; April, 1961.

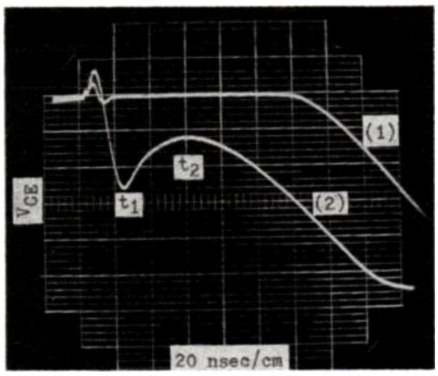

Fig. 1—Collector waveform.

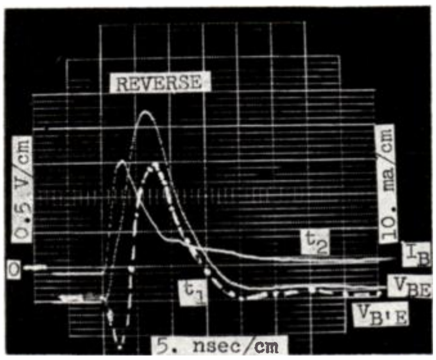

Fig. 2—Base waveforms.

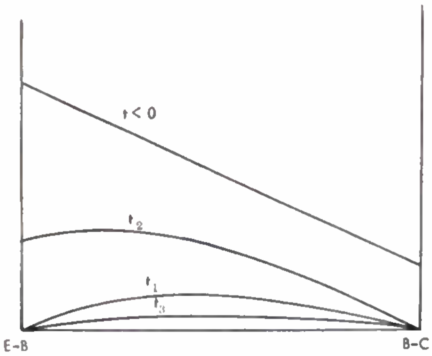

Fig. 3—Base minority carrier distribution.

resenting the internal emitter junction voltage (*i.e.*, excluding the voltage drop across the experimentally determined extrinsic base resistance), has been calculated by measuring the instantaneous base current. The base current was measured using a small sampling resistor and a differential sampling oscilloscope to assure that the current waveform was not being integrated.

The initial reversal of $V_{B'E}$ indicates that the internal emitter junction does recover momentarily when sufficient overdrive is provided. Note that, if the contribution of the total capacitance between collector and ground is recognized and considered, good agreement exists between the time when the emitter junction becomes forward biased and when the collector waveform begins to rise (denoted by t_1 in Figs. 1 and 2). This agreement establishes the relationship between the "wiggle" effect and emitter junction recovery.

Fig. 3 shows the distribution of base minority carriers implied by the collector waveform (curve 2) of Fig. 1 and the $V_{B'E}$ waveform of Fig. 2. The large charge transferred because of the speed-up capacitor causes the emitter and collector junctions to recover almost immediately. A short time later the emitter junction again becomes forward biased and causes an increase in collector current. Finally, the transistor goes through its normal recovery phase. Variations from the collector waveform, shown as curve (2) in Fig. 1, have been observed. These variations can be explained by appropriate changes in the distribution of base minority carriers.

The model for the "wiggle" effect proposed here also satisfactorily explains other features of the turn-off waveform. For example, the initial fast fall is a consequence of the recovery of emitter and collector junctions.[4]

It is anticipated that this work will be published in greater detail in the near future.

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[4] R. P. Nanavati, "A fundamental lower bound for junction transistor fall time," *Proc. Natl. Electronics Conf.*, vol. 17, pp. 178–181; October, 1961.

Optical Maser Characteristics of Tm^{+3} in CaWO$_4$*

Recently[1] stimulated emission has been observed in the infrared fluorescence line at 1.063 microns (77°K) of Nd^{+3} in CaWO$_4$. Optical maser effects were observed both at room temperature and 77°K. Later publications[2,3] report optical maser characteristics of Nd^{+3} in two additional host lattices: CaF$_2$ and SrMoO$_4$.

This note and the succeeding one[4] describe optical maser characteristics of two other trivalent rare earth ions in CaWO$_4$: thulium and holmium.

To observe stimulated emission[1] a cylindrical rod of CaWO$_4$:Tm^{+3} with polished silvered spherical ends is pulsed with high-intensity illumination from a GE FT524 helical xenon flash lamp and the light emerging from the lightly silvered end (~5 per cent transmitting) is focussed onto the entrance slit of a grating spectrometer. The analyzed beam is focussed onto a gold-doped germanium photoconductive detector[5] cooled to 77°K and the signal displayed on an oscilloscope.

When irradiated by visible light, trivalent thulium fluoresces strongly in CaWO$_4$ in the region 1.7–2 microns. The infrared fluorescence spectrum at 77°K, shown in Fig. 1, corresponds to $^3H_5 \rightarrow {^3H_6}$ transitions

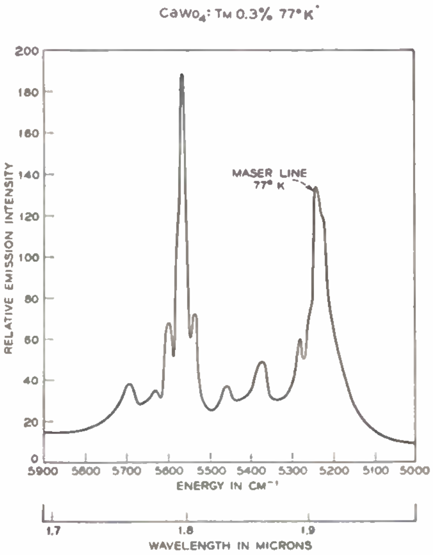
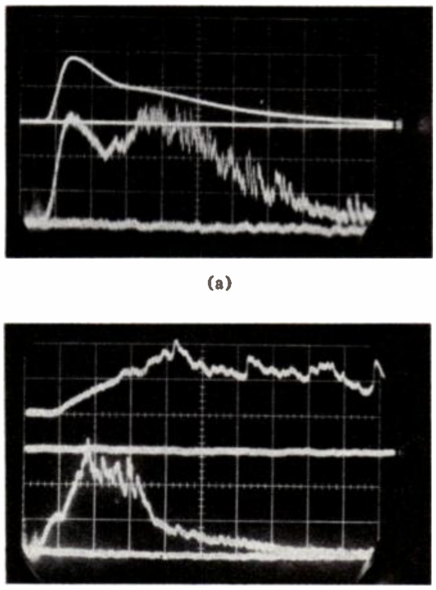


Fig. 1—Infrared fluorescence of Tm^{+3} in CaWO$_4$ at 77°K. The line at 1.911 microns exhibits stimulated emission.



(a)

(b)

Fig. 2—Stimulated emission in CaWO$_4$:Tm^{+3} at 77°K.
(a) Upper trace: variation of lamp intensity with time
Lower trace: stimulated emission at 1.911 microns superimposed on natural fluorescence and stray lamp light.
Time scale for both traces: 100 μsec/division; pump intensity 50 per cent above threshold.
(b) Fluorescence and stimulated emission at 1.911 microns, 300 per cent above threshold.
Time scale—upper trace: 10 μsec/division lower trace: 100 μsec/division.

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* Received by the IRE, November 20, 1961.

[1] L. F. Johnson and K. Nassau, "Infrared Fluorescence and Stimulated Emission of Nd+3 in CaWO$_4$," Proc. IRE, (Correspondence), vol. 49, pp. 1704–1706; November, 1961.

[2] L. F. Johnson, "Optical Maser Characteristics of Nd^{+3} in CaF$_2$" (to be published).

[3] L. F. Johnson and R. R. Soden, "Optical Maser Characteristics of Nd^{+3} in SrMoO$_4$" (to be published).

[4] L. F. Johnson, *et al.*, "Optical Maser Characteristics of Ho^{+3} in CaWO$_4$," Proc. IRE (Correspondence), vol. 50, pp. 87–88; January, 1962.

[5] We are indebted to Dr. H. Levinstein of Syracuse University for providing us with a crystal of Au-doped germanium.

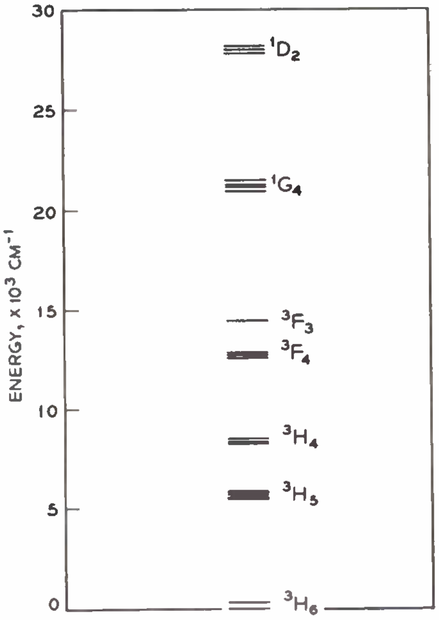

Fig. 3—Energy level diagram for Tm^{+3} in $CaWO_4$.

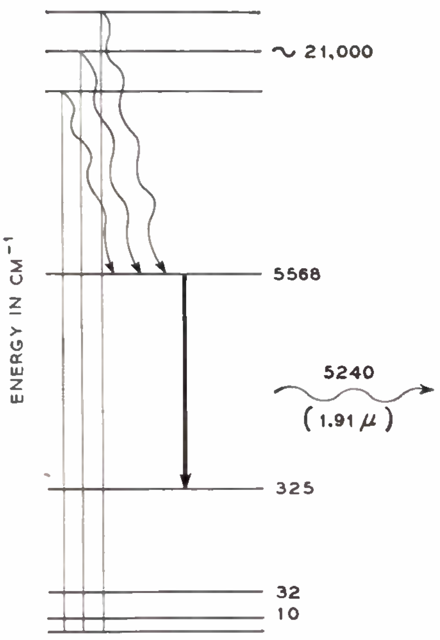

Fig. 4—Optical maser action in $CaWO_4:Tm^{+3}$.

for the free ion. At 77°K stimulated emission is observed in the fluorescence line at 1.911 microns, designated by the arrow in Fig. 1. No maser action could be obtained at room temperature. An oscilloscope trace of stimulated emission at 1.911 microns is shown in Fig. 2(a); a quartz prism was employed for wavelength isolation of the maser beam (lower trace). Both traces are triggered simultaneously. The stimulated emission is shown in more detail in Fig. 2(b) where 1.911 microns has been isolated from natural fluorescence and stray light by means of a grating. In the lower trace it is seen that the fluorescence rises with lamp intensity until a critical density of inverted population is attained between the two states involved in the transition. A series of sharp spikes or bursts of stimulated emission then appear. The upper trace of Fig. 2(b) displays the character of these spikes on an expanded time scale of 10 μsec/division. The randomness and irregularity is quite unlike oscillations in the stimulated emission of $CaWO_4:Nd^{+3}$. The threshold for stimulated emission in the helical xenon lamp configuration was 125 joules of energy into the lamp for the one crystal examined to date.

An energy level diagram for the $4f \rightarrow 4f$ transitions of Tm^{+3} in $CaWO_4$ based on preliminary absorption data combined with fluorescence measurements at 77°K is shown in Fig. 3. The term assignments are taken from the work of Gruber and Conway[6] on thulium salts. Absorption is quite weak for the concentration used (0.5 per cent), the strongest lines occurring between 1.7 to 1.8 microns. In the experiment described pumping probably arises principally from the near coincidence of weak absorption lines in the region 4600 to 4800 A and xenon lines from the lamp. On this basis a diagram depicting optical maser action in $CaWO_4:Tm^{+3}$ is shown in Fig. 4. The terminal state for the $^3H_5 \rightarrow {}^3H_6$ transition involved in stimulated emission lies 325 cm^{-1} above the ground state.

We wish to thank Miss D. M. Dodd for obtaining the absorption data and P. M. Ness for preparing the maser rod. The technical support of R. A. Thomas and A. M. Broyer is also appreciated.

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[6] J. B. Gruber and J. G. Conway, "Electronic energy levels and crystal quantum states of Tm (IV)," *J. Chem. Phys.*, vol. 32, pp. 1178–82; April, 1960.

Optical Maser Characteristics of Ho^{+3} in $CaWO_4$*

When trivalent holmium in $CaWO_4$ is excited by visible light intense infrared fluorescence is produced in the range 1.9 to 2.1 microns. The fluorescence spectrum at 77°K is shown in Fig. 1. The lines are $4f^{10}$ transitions and correspond to $^5I_7 \rightarrow {}^5I_8$ transitions of the free ion. The assignments are taken from McClure.[1]

A rod of $CaWO_4:Ho^{+3}$ (0.5 per cent) was examined for optical maser characteristics in the same manner as $CaWO_4:Tm^{+3}$. At 77°K stimulated emission is observed in the infrared fluorescence line at 2.046 microns, designated by the arrow in Fig. 1. No maser action was found at room temperature. An oscilloscope trace taken with the beam passed through a grating spectrometer set

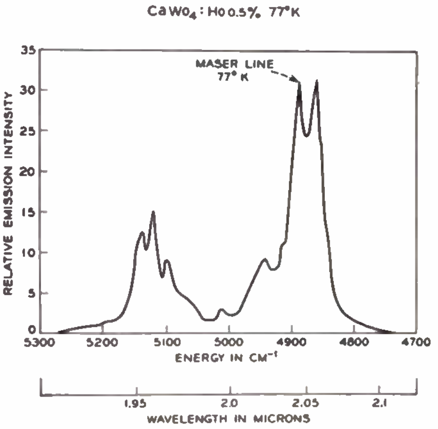

Fig. 1—Infrared fluorescence spectrum of Ho^{+3} in $CaWO_4$ at 77°K. The line at 2.046 microns exhibits stimulated emission.

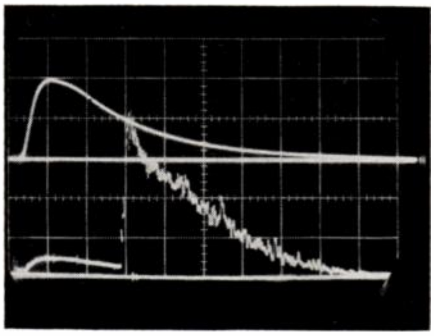

Fig. 2—Stimulated emission in $CaWO_4:Ho^{+3}$ at 77°K.
Upper trace: variation of lamp intensity with time.
Lower trace: stimulated emission at 2.046 microns superimposed on natural fluorescence and stray light from the lamp.
Time scale: 100 μsec/division.
Pump intensity: 30 per cent above threshold.

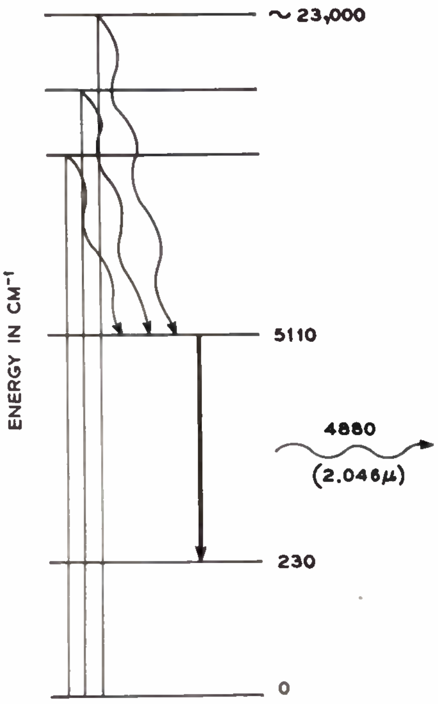

Fig. 3—Optical maser action in $CaWO_4:Ho^{+3}$. The principal pumping region lies in the range 4400 to 4600 A (~23,000 cm^{-1}). The terminal state of the $^5I_7 \rightarrow {}^5I_8$ transition exhibiting stimulated emission lies 230 cm^{-1} above the ground state.

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* Received by the IRE, November 20, 1961.

[1] D. S. McClure, "Electronic spectra of molecules and ions in crystals," in "Solid State Physics," vol. 9, pp. 400–525; 1959.

at 2.046 microns is shown in Fig. 2. Spikes in the pattern are quite irregular in both amplitude and frequency, similar to those shown in the preceding note[2] for $CaWO_4:Tm^{+3}$. The threshold for stimulated emission was 300 joules of energy into the helical xenon lamp.

The strongest absorption lines of Ho^{+3} in $CaWO_4$ lie in the range 4400 to 4600 A. By combining preliminary absorption data with the fluorescence spectrum of Fig. 1 a diagram depicting optical maser action in $CaWO_4:Ho^{+3}$ is obtained (Fig. 3). The terminal state of the $^5I_7 \rightarrow {}^5I_8$ transition involved in stimulated emission lies 230 cm^{-1} above the ground state. Such a small splitting implies that the threshold at liquid hydrogen temperature may be considerably lower than at 77°K.

We wish to thank Miss D. M. Dodd for the absorption measurements and R. A. Thomas for preparing the maser rod.

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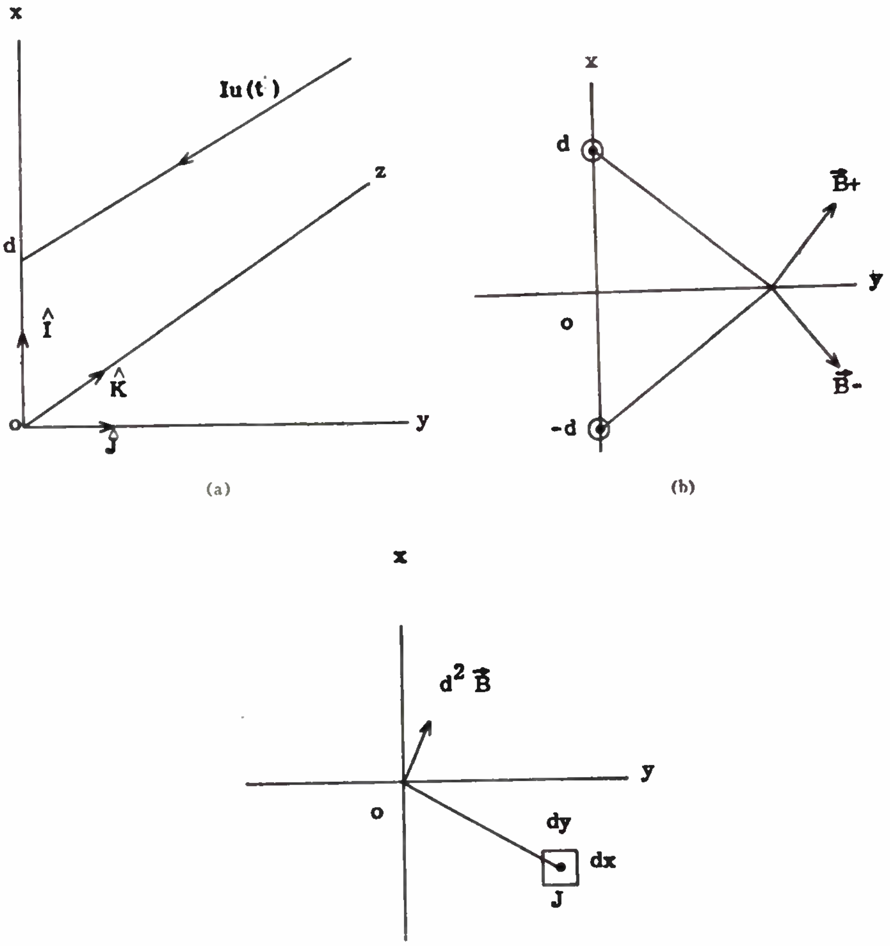
[2] L. F. Johnson, G. D. Boyd and K. Nassau, "Optical maser characteristics of TM^{+3} in $CaWO_4$," Proc. IRE (Correspondence), vol. 50, pp. 86–87; January, 1962.

The Field Induced by Eddy Currents in a Semi-Infinite Solid*

The current interest in the switching of thin permalloy films poses the problem of the transient decay of the magnetic field produced by eddy currents in the ground plane when the drive current in a wire above ground is switched. This problem is considered by Smythe[1] for the case of a source above a thin ground plane of surface resistivity s, but as far as the author knows has not been treated for a thick ground plane. We consider here the problem of a current-carrying wire a distance d above a semi-infinite solid of conductivity σ and permeability μ_0. At zero time a step of current I is switched on.



(a) (b)

(c)

Fig. 1.

The field may be described by a vector potential A. We have

$$E = -\frac{\partial A}{\partial t} = J/\sigma, \tag{1}$$

where E and J are electric field and current density. Of course

$$\begin{aligned} \sigma &= \text{constant} \quad & x \leq 0 \\ \sigma &= 0 & x > 0. \end{aligned} \tag{2}$$

Eqs. (1) and (2) imply that A for $x<0$ cannot change instantaneously since this would require infinite current density in the solid. Therefore the magnetic field B in the solid cannot change in zero time, and eddy currents are induced on the surface of the solid such that no magnetic field penetrates into the solid. From outside the solid this initial surface eddy current is indistinguishable from an image current of opposite sign in the solid.

If we neglect the displacement current (we are interested in small line-solid spacings and the displacement current in the solid is negligible with respect to the conduction current) the initial magnetic field is given by

$$\nabla^2 B = 0 \qquad (B)_{x<0} = 0,$$

which has a solution, the field due to two line sources distance $2d$ apart for $x>0$.

We find the initial surface current K from the boundary condition,

$$(\imath \times B)_{x=0} = \mu_0 K.$$

Since

$$B_{x=0} = \hat{j}\frac{\mu_0 I}{\pi} d/(x^2+d^2),$$

$$K = -\hat{k}\frac{I}{\pi}\frac{d}{x^2+d^2}.$$

This amount of surface current just cancels the initial B field in the solid due to the line source, at $x=d$.

In the solid[1] the current density is given by

$$\nabla^2 J = \sigma\mu_0 \frac{\partial J}{\partial t}.$$

A Greens function for this equation is[2]

$$G(x, y; x', y, t) = \frac{1}{4\pi a^2 t} \cdot \exp\left\{\frac{(x-x')^2+(y-y')^2}{4a^2 t}\right\},$$

where $a^2 = 1/\sigma\mu$.

Thus the current density in the solid at a general time becomes

$$J(x, y, t) = -\hat{k}\frac{Id}{2\pi^2 a^2 t} e^{-(x^2/4a^2 t)} \cdot \int_{-\infty}^{\infty} \frac{e^{-(y-y')^2/4a^2 t}}{y'^2+d^2} dy'. \tag{3}$$

Here we have doubled the magnitude of the initial surface current to account for the absence of solid for $x>0$.

A current filament at x, y produces a horizontal (y) component at $x=y=0$ of

$$d^2 B_h = \frac{\mu_0 J x dx dy}{2\pi(x^2+y^2)}. \tag{4}$$

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* Received by the IRE, September 5, 1961.

[1] W. R. Smythe, "Static and Dynamic Electricity," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 390 ff.

[2] P. M. Morse and H. Feshbach, "Methods of Theoretical Physics," McGraw-Hill Book, Co., Inc., New York, N. Y., p. 861; 1953.

Here the vector sign on J is dropped.

If we define,

$$\xi = \frac{x}{2a\sqrt{t}}, \quad \eta = \frac{y}{2a\sqrt{t}}, \quad \lambda = \frac{d}{2a\sqrt{t}} \quad \text{etc.,}$$

we have by (3) and (4), since vertical components of B cancel when integrated over the solid,

$$B_h = B_0 \frac{2\lambda^2}{\pi^2} \int_{-\infty}^{\infty} \int_0^{\infty} \int_{-\infty}^{\infty} \cdot \frac{e^{-\xi^2-(\eta-\eta')^2} \xi d\eta' d\xi d\eta}{(\eta^2+\xi^2)(\eta'^2+\lambda^2)}, \tag{5}$$

where

$$B_0 = \frac{\mu_0 I}{2\pi d} \cdot$$

With the aid of

$$\int_0^{\infty} \frac{e^{-\xi^2}\xi d\xi}{\eta^2+\xi^2} = \frac{e^{\eta^2}}{2} \int_{\eta^2}^{\infty} \frac{e^{-y}dy}{y},$$

$$\cdot \int_0^{\infty} \int_{-\infty}^{\infty} \frac{e^{-\xi^2-(\eta-\eta')} \xi d\xi d\eta}{(\eta^2+\xi^2)(\eta'^2+\lambda^2)}$$

$$= \frac{e^{\eta^2}}{2} \mathcal{E}i(-\eta^2) \int_{-\infty}^{\infty} \frac{e^{-(\eta'-\eta)^2}}{\eta'^2+\lambda^2} d\eta'$$

$$= \sqrt{\pi} \left(\frac{1}{\lambda^2} + \frac{2e^{-\eta^2}}{\lambda^2} \sum_{n=1}^{\infty} \frac{(-1)^n}{(2\lambda)^{2n}} \cdot \left(\frac{d}{dn}\right)^{2n-1} \eta e^{-\eta^2} \right),$$

one finds from (5)

$$B = B_0 \left(1 + \frac{2\sqrt{\pi}}{\pi^2} \sum_{n=1}^{\infty} \frac{(-1)^n}{(2\lambda)^{2n}} I_n \right), \tag{6}$$

where

$$I_n = \int_{-\infty}^{\infty} \mathcal{E}i(-\eta^2) \left(\frac{d}{d\eta}\right)^{2n-1} \eta e^{-\eta^2} d\eta$$

$$= -2 \int_{-\infty}^{\infty} \frac{e^{-\eta^2}}{\eta} \left(\frac{d}{d\eta}\right)^{2n-2} \eta e^{-\eta^2} d\eta.$$

The first three I_n are

$$I_1 = \sqrt{2\pi}, \quad I_2 = -5\sqrt{2\pi}, \quad I_3 = 169\sqrt{\pi}/\sqrt{2}.$$

Thus an asymptotic expansion for B in descending powers of λ is

$$B \approx B_0 \left(1 - \frac{1}{\sqrt{2\pi}\lambda^2} - \frac{5\sqrt{2}}{8\pi\lambda^4} - \frac{169\sqrt{2}}{64\pi\lambda^6} \right).$$

In terms of the time t,

$$B \approx B_0 \left(1 - \frac{2\sqrt{2}\,t}{\pi d^2 \sigma \mu_0} - \frac{10\sqrt{2}\,t^2}{\pi d^4 \sigma^2 \mu_0^2} - \frac{169\sqrt{2}\,t^3}{\pi d^6 \sigma^3 \mu_0^3} \right). \tag{7}$$

For the case of aluminum ($1/\sigma = 2.86 \times 10^{-8}$ ohm meters) and a 7 mil line-plane spacing, the induced field has decayed by 1/10 in a time,

$$t = 170 \text{ nanoseconds.}$$

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Some Characteristics of Certain Electrostatically-Focused Devices*

In the course of a recent investigation of E-type devices [1], the authors have considered the possibility of employing an electrostatically-focused fast-wave device, such as that shown in Fig. 1, in the generation and amplification of microwave frequencies. A comparison of this embodiment with the helitron, described by Watkins and Wada [2], and with the "Cyclotron-Resonance Oscillator," described by Pantell [3], [4] and Chow [5], reveals that it possesses certain features resembling both devices. In particular, the ribbon-shaped electron beam is injected along a helical path such that the centrifugal force of the electron is balanced by the radial electric field force, just as with the helitron. However, certain differences in the electromagnetic field modes are necessitated in the proposed device by the fact that the "principal," or TEM, wave in a coaxial transmission system does not provide the required azimuthal RF electric field [6], while the parallel-wire geometry of the helitron supports a TEM wave possessing an azimuthal interacting field.

The suggestion that the ribbon-shaped electron beam might be arranged to interact with the transverse component of the TE_{11}-mode wave is, of course, not a new innovation. An excellent discussion of an electron beam interacting with the TE_{11}-mode in a circular waveguide, containing no center conductor, has been presented by Pantell [3], with the ribbon-shaped beam being confined to a helical trajectory by means of an axial magnetic field. The experimental work conducted by Pantell and Chow [3], [5] was carried out using a rectangular waveguide, although it should certainly be possible to duplicate the experiment in a circular guide.

The cutoff wavelength λ_c for the TE_{11} mode, which is presumed to propagate in the coaxial transmission system shown in Fig. 1, is approximately given by [6]

$$\lambda_c \approx \frac{1.873\pi r_s}{2} \left(1 + \frac{r_c}{r_s}\right) \text{cm}, \tag{1}$$

where r_s and r_c are in cm. Since this relation holds under the condition that r_s/r_c is equal to 3-to-1, λ_c is approximately (3.92 r_s). Hence the diameter of the outer circular guide, necessary to support the TE_{11} mode in the coaxial system, is $15.3/f_c$ where the cutoff frequency f_c is expressed in Gc. If the radial width of the ribbon beam is much less than the center-of-the-beam radius r_0, and if the latter is the geometric mean of r_s and r_c, then r_0 becomes $4.32/f_c$.

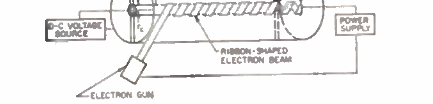

Fig. 1.

The voltage tuning curve can be computed from

$$f \approx \frac{e^{\sec^2\Phi/2}}{2\pi r_s} \sqrt{\frac{-\eta(V_c - V_s)}{\ln(r_s/r_c)}} \cdot \left(\frac{r_c}{r_s}\right)^{V_s/(V_c-V_s)}, \tag{2}$$

where η is the magnitude of the charge-to mass ratio of the electron, V_c and V_s are the cathode-to-center-conductor and cathode-to-outer-conductor potentials, and Φ is the beam pitch angle measured relative to a cross-sectional plane of the coaxial system. One can also use the basic equations of Watkins and Wada to obtain

$$\frac{V_s}{V_c} = \left[\frac{\sec^2\Phi + 2\ln(r_0/r_s)}{\sec^2\Phi + 2\ln(r_0/r_c)}\right], \tag{3}$$

which reduces to the case of purely circular motion obtained by the authors [1] when Φ vanishes.

Thus, at an operating frequency of 5 Gc, the above results lead to

$$r_s = 1.55 \text{ cm}, \quad r_0 = 0.865 \text{ cm}, \quad r = 0.517 \text{ cm}.$$

These values of transmission-line dimensions are relatively easy to fabricate and the beam radius is sufficiently small to provide a beam transmission of the order of fifty per cent.[1] If the beam pitch angle is taken to have a typical value of 10°, then the values of radii given above may be used in conjunction with (3) to yield a value of -0.148 for the ratio of V_s/V_c. Hence V_c and V_s are equal to 343 kv and -50.7 kv, respectively, in view of (2). These excessively large operating potentials limit the useful range much below the frequency chosen for computation. Similar conclusions were reached by Pantell [4] pertaining to a related device.

If the ratio of radii (r_s/r_c) is maintained at the value specified above, then as the frequency is reduced the beam radius r_0 must likewise increase. Thus, while the beam and equilibrium potentials are reduced by a decrease of frequency, the beam stiffness is adversely affected by this increased radius.

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References

[1] W. M. Nunn, Jr. and J. E. Rowe, "A small-signal analysis of E-type traveling-wave devices," IRE Trans. on Electron Devices, vol. ED-8, pp. 508–520; November, 1961.
[2] D. A. Watkins and G. Wada, "The helitron oscillator," Proc. IRE, vol. 46, pp. 1700–1705; October, 1958.
[3] R. H. Pantell, "Backward-wave oscillations in an unloaded waveguide," Proc. IRE (Correspondence), vol. 47, p. 1146; June, 1959.
[4] R. H. Pantell, "Electron beam interaction with fast waves," *Proc. Symp. on Millimeter Waves*, Polytechnic Inst. of Brooklyn, N. Y., vol. 9, pp. 301–312; April, 1959.
[5] K. K. Chow and R. H. Pantell, "The cyclotron-resonance backward-wave oscillator," Proc. IRE, vol. 48, pp. 1865–1870; November, 1960.
[6] N. Marcuvitz, "Waveguide Handbook," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., New York, N. Y., vol. 10, pp. 66–80; 1951.
[7] R. H. Pantell, "Small-signal analysis of the helitron oscillator," IRE Trans. on Electron Devices, vol. ED-7, pp. 22–29; January, 1960.

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* Received by the IRE, November 9, 1961.

[1] The experimental device constructed by the authors [1] was capable of providing an average beam transmission of 54 per cent over a spatial angle of approximately 270° with a beam radius of 1.38 cm.

Physical Basis for Electromagnetic Theory*

Without denying the truth of the very attractive treatments proposed by two recent authors,[1,2] it is necessary to amend their notes and point out the places where knowledge of the nature of the real physical world is required to complete the derivation of Maxwell's equations. Sugai states: "Vector operation (2) onto (6) gives (9)," where Vector operation (2) is

$$\text{div curl}\,\overline{F} = \nabla\cdot(\nabla\times\overline{F}) = 0 \qquad (1)$$

and (6) is

$$\nabla\times\overline{E} = -\frac{\partial\overline{B}}{\partial t}, \qquad (2)$$

while (9) is

$$\nabla\cdot\overline{B} = 0. \qquad (3)$$

Examining these relations in a bit more detail, we see that the first two lead to

$$\nabla\cdot(\nabla\times\overline{E}) = \nabla\cdot\left(-\frac{\partial\overline{B}}{\partial t}\right) = 0,$$

and, interchanging the order of differentiation,

$$-\frac{\partial}{\partial t}(\nabla\cdot\overline{B}) = 0$$

and integrating

$$\nabla\cdot\overline{B} = \text{constant}, \qquad (4)$$

which is not quite the same as (3). In fact, if we compare this with

$$\nabla\cdot\overline{D} = \rho, \qquad (5)$$

the similarity of the equations leads us to the reason for the vanishing of the constant in (4). This is exactly the place where we must call on physical experience for help. If we take the magnetic analog of the charge density, ρ, we have a collection of magnets each with two poles, and hence the net magnetic "charge" must be zero.

I know of no purely mathematical or logical argument for the nonexistence of an isolated magnetic pole. This physical fact arises from details of atomic structure. Eq. (3) can be derived, it is true, for the field due to a current carrying wire without recourse to the details of atomic structure,[3] but this is admittedly not a general derivation.

Another scheme for the logical development of electromagnetic theory, which should certainly be mentioned in this connection, is the one presented by Page and Adams,[4] where the theory is developed as the natural extension of Milne's work in the theory of relativity. In this treatment, a "moving element" is postulated with an "observer" or group of observers, and the equations of the theory of relativity are derived. Subsequently, an electric field is defined or, as the authors say, "represented graphically by continuous lines of force originating on positive charges and terminating on negative charges." Each of these lines of force is "considered as the locus of a dense linear aggregate of moving elements, or points."

From these assumptions, Maxwell's equations are developed; in particular,

$$\overline{H} = \nabla\times\overline{A}, \qquad (6)$$

and hence,

$$\nabla\cdot\overline{H} = 0 \qquad (7)$$

for free space.

The physical assumptions here are quite different from those required in the classical derivations. They include the assumption of a unique and constant velocity of signal transmission (the velocity of light) and certain characteristics of an observer, such as the ability to assign a temporal scale to events observed.

It would be interesting to speculate about the relationships between these assumptions and the "physical facts" of atomic structure. Is there some common foundation from which electromagnetic field theory and atomic structure both arise as the only possible description of the physical world due to limitations imposed by our methods of observation?

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Comments by Iwao Sugai[5]

The author would like to thank Dr. Leen for his attention to the recent letter,[2] which employs the universally accepted understanding of div $\overline{B}=0$ for the set of Maxwell's equations for macroscopic physical facts.

Leen's note appears to seek a basic answer for div $\overline{B}=0$ in macroscopic descriptions and div $\overline{B}=$constant, for microscopic phenomena. His aim seems to be unification of the field theories of electromagnetic phenomena and the field theories related to atomic structures. By the well-known correspondence principle, the prediction made by quantum physics (microscopic) approaches the description of classical physics (macroscopic) if the quantum number involved becomes very large, but not in the reverse direction. Therefore, if he maintains div $\overline{B}=C$ (constant), this C must become a function of quantum number in order to realize this smooth transition. This already requires that field quantities such as $\overline{E}$, $\overline{H}$, $\overline{D}$, and $\overline{B}$ must become functions of position, time, and quantum numbers.

Because $-(\partial/\partial t)\nabla\cdot\overline{B}=0$, mathematically speaking, $\nabla\cdot\overline{B}$ must be independent of time as an instantaneously observed fact. The question to be settled here is that it may be possible to say that $\nabla\cdot\overline{B}$ can be a finite quantity in a microscopic world. However, this assumption (or nonassumption) must be supported by experiment as well as by theory. If EM theory is a mathematical identity, it needs no physical experimentation. However, EM theory is the basis of the present electronics technology, and it represents a physical world. I think any further discussion on this fundamental unifying EM-atomic structure theory is rather specialized for the interest of the general public of this journal, despite the fact it is a worthy project.

There are many books to prove div $\overline{B}=0$ for macroscopic cases; yet the author found that the recent book by Langmuir[6] had a very clear and elegant proof. If the permeability constant depends on position and time for macroscopic cases, the expression of div $\overline{B}=0$ is different from div $\overline{H}=0$. Interestingly enough, attempts[7] to derive wave equations for stationary general linear media will deal with the commutabilities of linear scalar differential operators.

May I take this opportunity to thank Farhat for his kindness in pointing out a typographical error in the previous letter.[2] The correct expression for curl curl $\overline{F}$ is $\nabla(\nabla\cdot\overline{F})-\nabla^2\overline{F}$.

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Comment by P. A. Clavier[8]

While the author's work[1] is only mentioned in Leen's article, and not referred to directly, Leen misses one point the author tried to make.

The attempt was to show that the number of possible forms for physical laws was very small indeed, so that one could well develop all possible physical theories without mentioning the physical world itself. When, however, one wants to select a possible theory to represent a given phenomenon, there is first a choice between the small number of possible theories. When a given theory is selected, the values of the universal constants can still be chosen within that theory so as best to fit the experimental data. Such a constant is, for instance, the velocity of light.

As for $\nabla\cdot B$, the introduction of a new scalar is only necessary if the description obtained with 3 vectors and one scalar (E, H, i and ρ) is not satisfactory. (The introduction of ρ forces the introduction of i.)

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* Received by the IRE, August 2, 1961.

[1] P. Clavier, "Electromagnetic theory from a mathematical viewpoint," PROC. IRE (Correspondence), vol. 48, pp. 1494–1495; August, 1960.

[2] I. Sugai, "Vectors for waves and electrons," PROC. IRE (Correspondence), vol. 49, pp. 628–629; March, 1961.

[3] G. P. Harnwell, "Principles of Electricity and Electromagnetism," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 289–290; 1938.

[4] L. Page and N. I. Adams, "Electrodynamics," D. Van Nostrand Co., Inc., Princeton, N. J.; 1940.

[5] Received by the IRE, August 16, 1961.

[6] R. V. Langmuir, "Electromagnetic Fields and Waves," McGraw-Hill Book Co., Inc., New York, N. Y., p. 69; 1961.

[7] I. Sugai, "Operators for wave equations in general linear media," IRE TRANS. ON ANTENNAS AND PROPAGATION (Communication), vol. AP-9, pp. 501–502; September, 1961.

[8] Received by the IRE, August 17, 1961.

Explanation of One Type of RF Induced Negative Resistance in Junction Diodes*

In a previous correspondence[1] we reported the observance of negative resistance regions in the I-V characteristics of some p-n junction diodes when they were pumped by an RF signal. We propose to give here the reason for the observed negative resistance.

In our experiment the diode was driven via an impedance matching network which was resonant at the pump frequency. The diode behaves as a nonlinear capacitance in this resonant circuit. The effect of capacitive nonlinearity in a resonant circuit has been treated extensively.[2,3] The important result for our purpose is that shape of the response curve is distorted from that of the linear response curve to the shape displayed by the experimental response curve of Fig. 1. This curve was obtained by driving the circuit from a sweep frequency generator and loosely coupling a detector to the circuit.

If the driving frequency is kept fixed and the bias voltage is swept, the observed response voltage as a function of bias voltage is similar to the curve of Fig. 1 because sweeping the bias voltage has the effect of sweeping the resonant frequency of the circuit, the diode capacitance being a function of the bias voltage. An experimental curve of response versus bias voltage for the diode is shown in Fig. 2.

The RF response voltage appears across the diode. If the driving power is large enough, the response voltage drives the diode into the nonlinear resistance region, resulting in rectification of the RF current. This current rectification results in a forward direct current in the bias circuit.

A negative resistance in the bias circuit can now occur in the following way. The response voltage as a function of bias voltage has a negative slope in two regions, denoted by A and B in Fig. 2. The scope trace jumps across region A because the response curve exhibits an instability in that region.[2,3] If the negative slope of the response curve is steep enough in these regions, the rectified current will increase even though the bias voltage is decreasing. Hence the increase of the response with decreasing bias voltage in region A of Fig. 2 gives rise to an increasing rectified current with decreasing bias voltage in region A of Fig. 3, which is the bias circuit I-V characteristic observed, while the response curve of Fig. 2 occurred. Similarly, the negative slope of the response in region B of Fig. 2 causes the rectified current to increase as bias voltage decreases in region B of Fig. 3. In Fig. 3 the negative resistance at A is current controlled and that at B is voltage controlled.

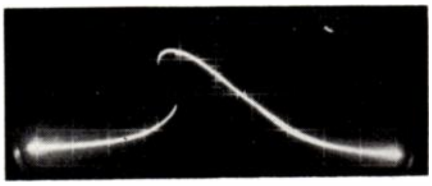

Fig. 1—Experimental response curve. Vertical scale is power sampled by the detector. Horizontal scale, representing frequency, is calibrated at 4 Mc/cm centered at 320 Mc. Input power 40 mw. Diode bias is −9 v. The diode used was a type TI C64.

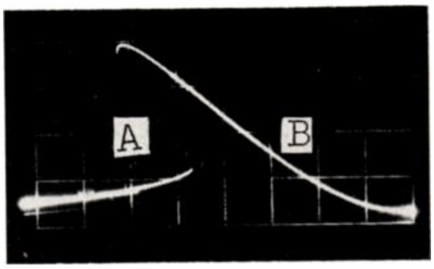

Fig. 2—Experimental response of the circuit as a function of back-bias voltage on the diode. Vertical scale is sampled power at the driving frequency of 310 Mc. Horizontal scale is negative bias voltage calibrated at 2 v/division with zero volts at the extreme right. Input power 40 mw.

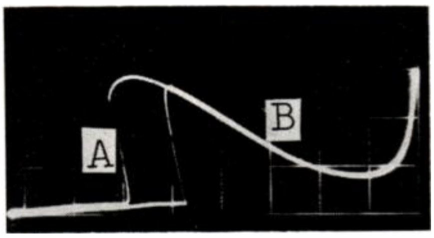

Fig. 3—Bias circuit I-V characteristic resulting from the response of Fig. 2. Horizontal scale same as Fig. 2. Vertical scale is 0.1 ma/division.

It should be emphasized here that the negative resistance effect described above is different from another negative resistance effect, due to minority carrier storage, which has frequently been observed in some variable-capacitance diodes.[4–6] That the two effects are of different nature can be seen from the following experimental observations. The resonance type of negative resistance described here does not exist unless there is a resonant circuit at the driving frequency, whereas the minority carrier negative resistance can exist even when there is not a resonant circuit at the driving frequency. The resonance type of negative resistance seems to be observable at any frequency[7] as long as the diode capacitance is appreciable compared to any linear capacitance in the circuit, while the minority carrier negative resistance disappears when the period of the driving signal is appreciably longer than the minority carrier lifetime. An explanation of the minority carrier negative resistance can be found in Siegel.[5]

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* Received by the IRE, October 23, 1961.

[1] J. C. McDade, "RF-induced negative resistance in junction diodes," Proc. IRE (Correspondence), vol 49, pp. 957–958; May, 1961.

[2] J. J. Stoker, "Nonlinear Vibrations," Interscience Publishers, Inc., New York, N. Y., pp. 81 ff.; 1950.

[3] W. J. Cunningham, "Introduction to Nonlinear Analysis," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 171 ff.; 1958.

[4] I. Hefni, "Effect of minority carriers on the dynamic characteristic of parametric diodes," *Electronic Engrg.*, vol. 32, pp. 226–227; April, 1960.

[5] K. Siegel, "Anomolous reverse current in varactor diodes," Proc. IRE (Correspondence), vol. 48, pp. 1159–1160; June, 1960.

[6] I. Hefni, "The negative resistances in junction diodes," Proc. IRE (correspondence), vol. 49, pp. 1427–1428; September, 1961.

[7] E. O. Keizer, "A carrier energized bistable circuit using variable capacitance diodes," *RCA Rev.*, vol. 18, pp. 475–485; December, 1957.

Cutoff Frequency of a Junction Transistor when Mobilities are not Constant*

Sugano and Koshiga[1] have presented a very interesting analysis of the cutoff frequency of a drift transistor, taking account of the mobility variation in the base region. This decrease in mobility in the heavily-doped base region, indeed, does reduce the cutoff frequency, as the authors point out. However, the effect of this mobility variation in practical structures may not be as serious as indicated by their numerical example.

In their numerical example, Sugano and Koshiga assume a built-in field of $8(kT/qW)$ where W is the base width, and assume that the mobility is decreasing with increasing impurity concentration as $N^{-1/3}$ throughout the entire base region. However, inasmuch as the mobility in germanium is essentially constant for $N<10^{16}$ cm^{-3}, these assumptions imply that the impurity concentration at the collector edge of the base region must be at least 10^{16} cm^{-3} and the impurity concentration at the emitter edge of the base region must be $\epsilon^8=3\times10^3$ as large, or 3×10^{19} cm^{-3}. In practical structures, however, such a very large value of concentration would *not* be used because of the difficulty of obtaining good emitter efficiency and of the very low emitter-base breakdown voltage and high emitter transition capacitance that would result.

In practical germanium drift-type structures, the impurity concentration at the base edge of the emitter-base junction would normally range from 5×10^{16} to 5×10^{17} cm^{-3}, while the impurity concentration at the collector-depletion-layer edge of the base region will generally range upward from 3×10^{14} cm^{-3}. Consequently, in most practical structures, the mobility variation will be important only through a *part* of the active base region. To illustrate the order of magnitude of the effect in typical structures, the cutoff frequency[2] ω_T (as distinct from the alpha-cutoff frequency ω_α considered by Sugano and Koshiga) can be calculated approximately by the method of stored charge, as shown by Varnerin[3] and as extended by Beale,[4] for a base structure in which the impurity concentration and minority-carrier diffusion constant vary as follows:

$$N(x) = N_0\epsilon^{-\beta x} \qquad 0 \le x < W \tag{1}$$

$$D_p(x) = \begin{cases} D_0\epsilon^{-\beta x/3} & 0 < x \le W_1 \\ D_1 & W_1 \le x < W \end{cases}. \tag{2}$$

Here x is distance measured through the base from the emitter, W is the base width, and W_1 is the distance from the emitter edge of the base region to the point in the base at

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* Received by the IRE, September 11, 1961.

[1] T. Sugano and F. Koshiga, "The calculation of cutoff frequencies of minority-carrier transport factors in drift transistors when the mobilities are not constant," Proc. IRE (Correspondence), vol. 49, p. 1218; July, 1961.

[2] The frequency at which the magnitude of the common-emitter short-circuit forward current transfer ratio $|h_{fe}|$ can be extrapolated to 1.

[3] L. J. Varnerin, "Stored charge method of transistor base transit analysis," Proc. IRE, vol. 47, pp. 523–527; April, 1959.

[4] J. R. A. Beale, "The calculation of transit times in junction transistors when the mobilities are not constant," Proc. IRE (Correspondence), vol. 48, p. 1341; July, 1960.

TABLE I

Case	N_0	N_W	βW_1	βW	$(\omega_T W^2/D_1)$	
	(cm^{-3})				Eq. (3)	Eq. (4)
(a)	2×10^{17}	5×10^{14}	3	6	5.9	7.2
(b)	10^{17}	5×10^{14}	2.3	5.3	5.8	6.5
(c)	5×10^{16}	5×10^{14}	1.6	4.6	5.55	5.85
(d)	—	—	8	8	2.2	9.1

which the impurity concentration has decreased to a value such that mobility, *i.e.*, D_p, no longer varies with impurity concentration, *e.g.*, for germanium, W_1 occurs in the base at the point where $N(x)=10^{16}$ cm^{-3}. In (1) and (2) β is a constant which is proportional to the built-in electric field, and can be evaluated in terms of the impurity concentration at the emitter and collector edges of the base region, N_0 and N_w respectively, as $\beta W=\ln(N_0/N_w)$. By substituting (1) and (2) into Beale's (2) for calculating the hole concentration in the base and by substituting the resulting expressions[5] for hole concentration in Varnerin's (1) the transit time, τ, which is the reciprocal of the cutoff frequency ω_T, may be written as

$$\tau = (1/\omega_T) = \frac{W^2}{D_1}\frac{1}{(\beta W)^2} \cdot \{(9/4)\epsilon^{\beta W_1/3} - (1/4)\epsilon^{-\beta W_1} + \epsilon^{-\beta W} + (\beta W - \beta W_1) - 3\}. \quad (3)$$

Note that this expression reduces to Beale's (7) if $W_1 = W$, *i.e.*, if mobility varies throughout the entire base region. Alternatively, if $W_1=0$, *i.e.*, if mobility is constant throughout the entire base region, (3) reduces to

$$\tau = (1/\omega_T) = \frac{W^2}{D_1}\frac{1}{(\beta W)^2} \cdot \{(\beta W - 1) + \epsilon^{-\beta W}\}. \quad (4)$$

This result could be extracted from Varnerin's analysis, and interestingly enough also was derived by te Winkel[6] using classical techniques, rather than the stored-charge approach!

To illustrate the order of magnitude of cutoff-frequency reduction for practical structures, the results of a series of calculations based on (3) and (4) are presented here in Table I.

Cases (a)–(c) represent typical germanium drift structures[7,8] and show that the decrease in cutoff frequency due to mobility variation is, at most, of the order of 5 to 20 per cent. Case (d) is that of Sugano and Koshiga and is included only to show that (3) above also predicts a rather severe decrease in cutoff frequency due to mobility variation for such a hypothetical structure. In fact, the results in Table I indicate that the cutoff frequency ω_T is reduced to 0.24 times the value corresponding to constant mobility in the base, rather than to the 0.3 factor quoted by Sugano and Koshiga for ω_α. However, this difference probably can be attributed to the fact that the result in Table I is obtained in terms of the cutoff frequency, ω_T, rather than in terms of the alpha-cutoff frequency, ω_α.

In conclusion, it also might be of interest to point out that a resaonable estimate of the effect of mobility variation on cutoff frequency can be obtained by employing the simpler (4), which assumes constant mobility, and using a value of mobility corresponding to the *mean* value of the impurity concentrations at the emitter and collector-depletion-layer edges of the base region.

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Authors' Comment[9]

In the preceding communication, Pritchard has discussed our numerical example. It seems worthwhile to make clear why we assumed such a hypothetical structure as previously reported.[1] A built-in field of 4 in terms of d/f, where d was the base width, f was $2kT/qE$ and E was the electric field, was assumed for convenience to compare our calculations with Krömer's results and in order to compare these under the severest condition. The mobility was assumed to decrease with increasing impurity concentration as $N^{-1/3}$ throughout the entire base region for convenience to compare our calculations with Beale's results[4] and to calculate modified Bessel functions numerically.

We are in complete agreement with Pritchard's opinion that the effect of minority carrier mobility variation in practical drift transistors is not as serious as indicated by our numerical example.

We have calculated the effect of mobility variation on the minority-carrier transport factor as the function of d/f, the parameter of which is ν, defined by (1) of our previous correspondence. Namely,

$$\mu_p = \mu_{p0}\left(\frac{N_d}{N_{a0}}\right)^{-\nu}.$$

The results are shown in Fig. 1, where κ is the improvement factor defined by

$$f_\alpha = \kappa \frac{D_{p0}}{2\pi\alpha^2},$$

where f_α is the cutoff frequency of the transport factor and D_{p0} is the diffusion constant of minority carriers at the collector edge of the base region.

Furthermore Pritchard has presented an interesting extension of Varnerin's stored-charge base transit time analysis. Here it must be pointed out that the base transit time τ is not the reciprocal of the cutoff angular frequency ω_α.

The dotted line in Fig. 1 shows the improvement factor when the cutoff angular frequency is defined by the reciprocal of the base transit time, which shows the great discrepancy.

It appears difficult to relate the cutoff frequency to the base transit time directly, because calculations based on the diffusion-mobility model allow for the broadening of the input signal and provide thereby a more detailed and accurate determination of phase lag for an ac signal,[10] but the stored-charge method does not.

However, the phase differences between the emitter current and the collector current at the cutoff frequency is an experience. Namely, the approximation

$$\omega_\alpha = \frac{1}{\tau}$$

is rewritten to give

$$\omega_\alpha \tau = 1,$$

which indicates that the phase angle of the transport factor at the cutoff frequency is one radian.

The relation is a good approximation for the diffusion transistor, because the phase angle of the transport factor at the cutoff frequency is 57°, or 0.997 radian.

For the case of the diffusion transistor, τ is given by

$$\tau = \frac{\alpha^2}{2D_p},$$

therefore the cutoff frequency is obtained as follows:

$$f_\alpha = \frac{D_p}{\pi\alpha^2}.$$

This formula is approximately equal to the result obtained from the diffusion equation.

For the case of the drift transistor, however, the phase angle θ_α of transport factor at the cutoff frequency is the function of the built-in field. Therefore, the relation of the cutoff frequency to the base transit time for the drift transistor should be

$$f_\alpha = \frac{\theta_\alpha}{2\pi\tau}.$$

But the usual way of deriving the expression for the phase angle of transport factor is to use the diffusion equation. Consequently, from the stored-charge method, we believe,

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[5] Note that there are two solutions, one valid for $0<x<W_1$, and a second expression for $W_1<x<W$.

[6] J. te Winkel, "Drift transistor—simplified electrical characterization," *Elec. and Radio Engr.*, vol. 36, p. 382, Eq. (17); August, 1959.

[7] A. L. Kestenbaum and N. H. Ditrick, "Design construction and high-frequency performance of drift transistors," *RCA Rev.*, vol. 18, p. 13; March, 1957.

[8] C. G. Thornton and J. B. Angell, "Technology of micro-alloy diffused transistor," PROC. IRE, vol. 46, p. 1171; June, 1958.

[9] Received by the IRE, October 3, 1961.

[10] L. J. Varnerin, private communication.

Fig. 1—Improvement factor of the cutoff frequency as the function of a built-in field.

the cutoff frequency has not been obtained as yet.

In summary, Pritchard's contribution of a more practical analysis for the case of variable mobility is interesting and illuminating. The relation of the cutoff frequency to the base transit time, however, should be discussed more rigorously.

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A 5-Ampere Peak-Current Tunnel-Diode Oscillator at VHF*

Germanium tunnel diodes with peak currents of about 5 amperes have been used as oscillators at frequencies up to 100 Mc.

The diodes are "solution-grown" squares with dimensions of 0.02 in × 0.02 in × 0.011 in. One face is soldered to the end of a short piece of No. 13 copper wire. Another wire makes a mercury wetted contact to the other face. The capacitance of the diodes (estimated as the ratio of peak-minus-valley current to the maximum slope of the constant current switching transient) was about 0.02 μf. The series parasitic inductance was calculated and measured to be about 20 $\mu\mu$h, and the series resistance was about 0.01 ohms.

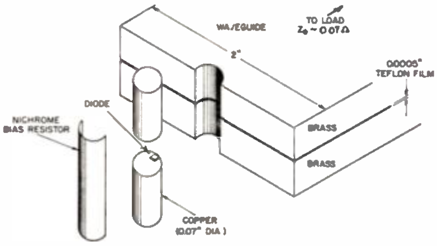

Fig. 1—Oscillator structure (not to scale).

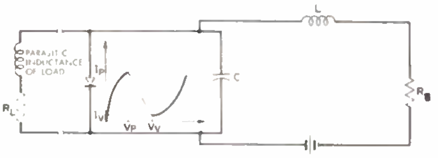

Fig. 2—Simplified equivalent circuit.

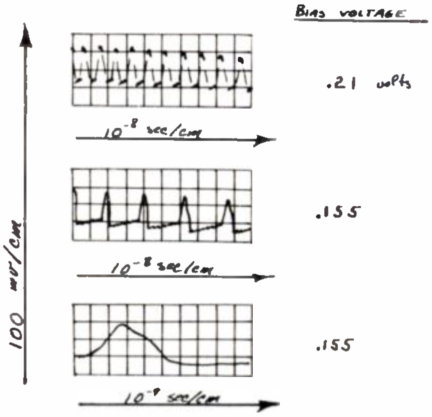

Fig. 3—Unloaded output waveforms.

The oscillator structure is shown in Fig. 1. The oscillator circuit is formed by pressing a thin strip of nichrome tightly against the copper wires to serve as bias resistance and to complete a small-loop inductance of 100 to 200 $\mu\mu$h. The low-impedance output transmission line allows loading of the diode.

Such an oscillator can be crudely analyzed by assuming that the series parasitic inductances of the diode and the load are negligible, and the v-i characteristic has constant slope in the negative resistance region as shown in Fig. 2. The load conductance can be increased until

$$\frac{1}{R_N} - \frac{1}{R_L} = \frac{C}{L} R_B,$$

at which point the power into R_L will be

$$P_0 = (1 - r/l)P_{\max}, \tag{1}$$

where $r = R_B/R_N$, $l = L/R_N^2C$, $P_{\max} = (V_v - V_p)^2/8R_N$ and $R_N = (V_v - V_p)/(I_p - I_v)$.

If $l \ll 1$, the oscillation is sinusoidal with frequency

$$f = \frac{1}{2\pi\sqrt{LC}}.$$

If $l \gg 1$ the unloaded oscillation is of the relaxation type, and under specified conditions of loading, the frequency should be proportional to R_N/L. Thus the frequency should have the functional form

$$f \approx \frac{f_d}{\sqrt{l + al^2}}, \tag{2}$$

where $f_d = 1/2\pi R_N C$. Measurements of frequency vs inductance for low peak-current diodes indicate that $a \approx 0.12$.

Unloaded output waveforms are shown in Fig. 3 for diode No. 68-3 ($I_p = 5.2$ a, $R_N = 0.034$ ohm, $C = 0.016$ μf, so $f_d = 290$ Mc and $R_N^2C = 19$ $\mu\mu$h). The frequency shows the dependence upon bias voltage which is characteristic of relaxation oscillation.

With diode No. 68-2 ($I_p = 4.35$ a, $R_N = 0.066$ ohm, $C = 0.014$ μf and $P_{\max} = 92$ mw, so $f_d = 175$ Mc and $R_N^2C = 60$ $\mu\mu$h) the output power is measured as

Frequency	f/f_d	l	r	Power (measured)	output (calculated)
55	0.31	6	0.77	87	80
100 Mc	0.57	2.4	0.4	65 mw	76 mw

The "calculated" power output is estimated from (1) and (2). There is fair agreement with the measured power, although the measured power appears to be falling off more rapidly with frequency. This may be because the effect of the load parasitic inductance has been entirely neglected.

The author would like to express his gratitude to the staff of the Bell Telephone Laboratories at Murray Hill, N. J., for the opportunity to use their facilities to fabricate the tunnel diodes considered here.

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* Received by the IRE, September 25, 1961. This note is based on a thesis submitted in partial fulfillment of the requirements of the degree of Doctor of Science in the Dept. of Elec. Engrg. at the Mass. Inst. Tech., Cambridge, Mass., on August 21, 1961.

On the Probability of Error for Largest-of Selection*

Quite often in the processing of radar and communications signals, a group of channel outputs are compared and the largest selected as the channel containing the signal. Inherent in this process is the possibility of making an error in selection. The probability of making such an error is the subject of this note in which a closed-form expres-

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* Received by the IRE, September 5, 1961.

sion is derived and a family of curves presented.

A model consisting of N channels is assumed. All channels contain noise; one of the channels contains a signal as well. The noise in the N channels are independent and Gaussian with mean zero and average power ψ_0. The signal is a sinusoid of amplitude P. The outputs of the N channels are envelope detected and the largest selected as the channel containing the signal.

The probability of making an error, P_E, can be computed from the probability densities given by Rice,[1] for the envelope of a sinusoid immersed in Gaussian noise. These are: for the N-1 channels containing only noise

$$p_1(V) = \frac{V}{\psi_0} e^{-(V^2/2\psi_0)}, \qquad (1)$$

and for the channel containing signal and noise

$$p_2(X) = \frac{X}{\psi_0} e^{-(X^2+P^2)/2\psi_0} I_0\left(\frac{XP}{\psi_0}\right). \qquad (2)$$

It is interesting to note that, since the relative signal amplitudes will not be effected by detectors whose output is a monotonic function of the input amplitude, any monotonic detector could have been used in the model. The envelope detector was selected for ease of analysis.

An expression for probability of error is obtained by finding the probability of making the correct selection, and then subtracting this quantity from unity. The probability of correct selection is obtained by, first, computing the probability of success for each possible value of the signal and then averaging these probabilities over all possible values of the output of the signal channel. This operation is expressed functionally by

$$P_e = 1 - \int_0^\infty p_2(X)\left[\int_0^X p_1(V)dV\right]^{N-1} dX. \qquad (3)$$

By inserting (1) and (2) into (3), we obtain

$$P_e = 1 - \int_0^\infty \frac{X}{\psi_0} e^{-(X^2+P^2)/2\psi_0} I_0\left(\frac{XP}{\psi_0}\right) \cdot\left[\int_0^X \frac{V}{\psi_0} e^{-V^2/2\psi_0} dV\right]^{N-1} dX. \qquad (4)$$

It is easily shown that

$$\int_0^X \frac{V}{\psi_0} e^{-(V^2/2\psi_0)} dV = 1 - e^{-(X^2/2\psi_0)}$$

and using the Binomial series, we obtain

$$\left[1 - e^{-(X^2/2\psi_0)}\right]^{N-1} = \sum_{n=0}^{N-1} (-1)^n \binom{N-1}{n} e^{-(nX^2/2\psi_0)}. \qquad (5)$$

Combining (5) and (4) and reversing the order of summation and integration, we have

$$P_e = 1 - \sum_{n=0}^{N-1} (-1)^n \binom{N-1}{n} \int_0^\infty \frac{X}{\psi_0} \cdot e^{-[X^2(n+1)+P^2]/2\psi_0} I_0\left(\frac{XP}{\psi_0}\right) dX. \qquad (6)$$

Watson[2] has shown that

$$\int_0^\infty \frac{X}{\psi_0} e^{-[x^2(n+1)+P^2]/2\psi_0} I_0\left(\frac{XP}{\psi_0}\right) dX = \frac{1}{n+1} e^{-(n/n+1)P^2/2\psi_0}. \qquad (7)$$

Combining (7) and (6) yields

$$P_e = 1 - \sum_{n=0}^{N-1} \frac{(-1)^n}{n+1} \binom{N-1}{n} e^{-(n/n+1)P^2/2\psi_0}.$$

By letting $R = P^2/\psi_0$ (note that R corresponds to $2E/N_0$ of Woodward[3] and matched filter theory) and recognizing that the first term of the series is unity, we have

$$P_e = \sum_{n=1}^{N-1} \frac{(-1)^{n+1}}{n+1} \binom{N-1}{n} e^{-(n/n+1)R/2}. \qquad (8)$$

For small values of N there is little difficulty in calculating P as a function of R; however, for N larger than about 7, it is usually more expedient to use an electronic computer. The results of such a program are shown in Fig. 1 for R from zero to 50 and logarithmic increments of N from 2 to 27.

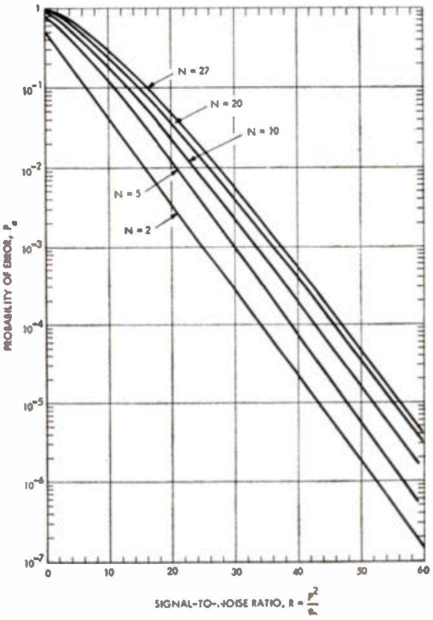

Fig. 1—Probability of error for largest of selection.

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[1] N. Wax, "Noise and Stochastic Processes," Dover Publications, Inc., New York, N. Y.; p. 238, 1954.

[2] G. N. Watson, "Theory of Bessel Functions," Cambridge University Press, London, Eng., p. 393; 1948.

[3] P. M. Woodward, "Probability and Information Theory," Pergamon Press, Inc., New York, N. Y., p. 103; 1955.

Doppler Studies of the Ionosphere with Vertical Incidence*

Several articles have appeared recently[1,2] discussing ionospheric investigations by means of the Doppler shifts produced by movements of the layers. It may be of interest that similar investigations are being carried out at the Boulder Laboratories of the National Bureau of Standards, Colo. The method in use is essentially that described by Watts and Davies[3] for oblique propagation except that the transmitter and receiver are separated by about 15 km, thus giving essentially vertical incidence. Briefly, the technique consists of recording, on slow-moving magnetic tape, the beat frequency produced by the received carrier and a local off-set oscillator of high stability. The beat frequency (about 5 cps) is converted into an audio tone by fast playback, and the frequency spectrum is analyzed by standard audio techniques. Time variations in the phase path (P) of the wave produce equivalent frequency variations of the received signal. The off-set oscillator makes possible the determination of the senses of these frequency variations. Sample Doppler "ionograms" produced by this technique are shown in Fig. 1 on carrier frequencies of 2.10, 4.00 and 5.05 Mc. The traces on 4.0 and 5.05 Mc are similar in shape but are

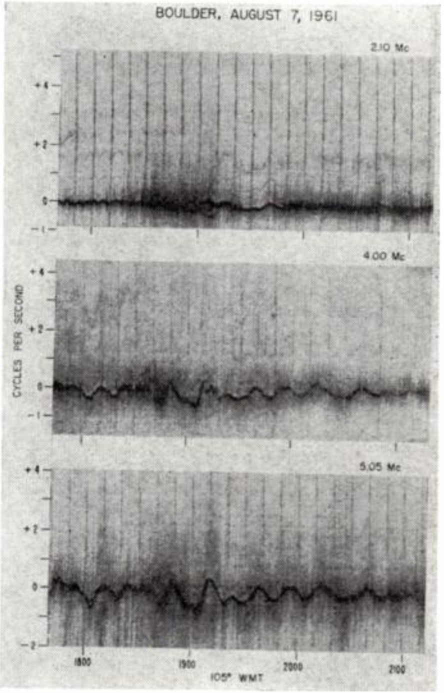

Fig. 1—Frequency fluctuations of signals reflected from the ionosphere with normal incidence.

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* Received by the IRE, October 5, 1961.

[1] T. Ogawa, S. Ando and A. Yoshida, "Vertical incident Doppler ionogram," PROC. IRE (Correspondence), vol. 49, p. 643; March, 1961.

[2] J. W. Findlay, *et al.*, "Vertical incidence Doppler ionogram," PROC. IRE (Correspondence), vol. 49, p. 1220; July, 1961.

[3] J. M. Watts and K. Davies, "Rapid frequency analysis of fading radio signals," *J. Geophys. Res.*, vol. 65, pp. 2295–2301; August, 1960.

similar to the 2.10-Mc trace only near 1930 (105° West Meridian Time).

Several frequencies can be used to give some information about the level at which the observed effects originate. The observed instantaneous frequency shift (δf) of a carrier frequency f, Mc, is given by the formula

$$\delta f = \frac{-f}{c}\frac{dP}{dt}, \tag{1}$$

where

$$P = \int_s n ds \cos \alpha, \tag{2}$$

n being the phase refractive index at a point on the ray path where the ray makes an angle α with the direction of the wave normal. (Note that this difference in direction arises from the anisotropy produced in the ionosphere by the earth's magnetic field.) c is the free-space velocity of light. For vertical propagation, $ds \cos \alpha = dh$ where dh represents an increment in the actual height (h); thus

$$P = 2\int_h n dh. \tag{3}$$

From (3) it can be seen that with vertical incidence, variations in phase path may be brought about either by changes in height of reflection only or by changes in refractive index along the path. Consider the case in which only the height of reflection varies.

$$\delta f = -2\frac{f}{c}\frac{dh}{dt} \tag{4}$$

so that, for a given ionospheric height variation, the frequency fluctuations are proportional to the carrier frequency. On the other hand, consider the case in which ionization is produced in a slab, of thickness t, and in which the electron density is N/m^3, well below the level of reflection, *e.g.*, the D region, where the refractive index, n, is approximately unity so that

$$n = \sqrt{1 - k\frac{N}{f^2}} \simeq 1 - \frac{1}{2}k\cdot\frac{N}{f^2}, \tag{5}$$

where

$$k = 8.06 \times 10^{-11}.$$

Now

$$\frac{dP}{dt} = \frac{1}{2}t\frac{k}{f^2}\frac{dN}{dt}.$$

Therefore

$$\delta f = \frac{1}{2}\frac{t}{c}\frac{k}{f}\frac{dN}{dt}, \tag{6}$$

so that, under these conditions, the magnitude of the frequency fluctuations is inversely proportional to the frequency of the carrier wave.

By observing several different carrier frequencies it should be possible to determine, at least to some extent, whether the observed phenomena are due to changes in height of reflection or to ionization changes in the lower ionosphere. Measurements of the frequency fluctuations, in Fig. 1, near 1930 and 1940, respectively, show that the ratios are approximately the ratios of the frequencies, indicating that most of the changes are due to variations in the levels of reflection. One other point of interest concerns the measurement of phase path. As has been pointed out by Ogawa, Ando and Yoshida,[2] the Doppler technique measures time rates of change of phase path. However, it is possible to obtain net changes in phase path (ΔP) by integrating (1):

$$\int \delta f \cdot dt = -\frac{f}{c}\Delta P. \tag{7}$$

Now the integral on the left-hand side is the area under the δf-time curve, which can be determined experimentally.

The method described above has several advantages over the system described by Ogawa, Ando and Yoshida, most important of which are: first, that the sense of the changes is determined and, second, that the total phase-path change can be easily obtained. It lacks the virtual height determination possible in Findlay's technique.

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Measuring Phosphor Resolution*

The resolution of settled phosphor screens can be measured with an electron gun and a fine metallic mesh. The screen to be tested is illuminated through the mesh by a flood beam from the electron gun (Fig. 1). As successively finer meshes are used, the shadow of the mesh ceases to be resolved through the phosphor, and a mesh resolution number is assigned to the screen corresponding to the just discernible web width.

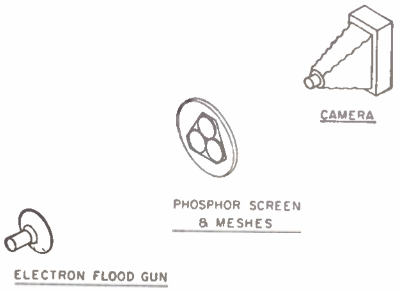

Fig. 1—The screen to be tested is illuminated through the mesh by a flood beam.

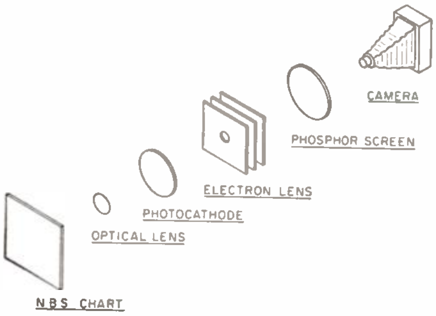

Fig. 2—A commonly used system focuses an electron image of a NBS chart onto the screen.

The resolution of the system is limited by the scattering of the electrons at the mesh. The error due to this scattering can be kept small if the separation of the screen and the mesh is only a few mils and the drift space between the gun and the screen is several inches. In contrast, one commonly used system projects a National Bureau of Standards lens resolution chart on a transparent photocathode and focuses the resultant electron image on the screen to be tested (Fig. 2). With this method the resolution of the optical lens, the photocathode, and/or the electronic lens might be worse than that of the screen.

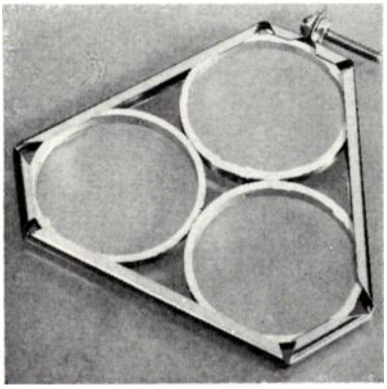

Fig. 3—The meshes are mounted behind the phosphor in this holder.

The meshes are cemented to rings, which are mounted in a holder that accommodates three rings simultaneously, so that the resolution of the phosphor can be bracketed (see Fig. 3). To test the resolution of the camera, the holder is mounted between the phosphor and the camera and backlighted with light from the phosphor. In our tests, the 1000-lines-per-inch mesh with a web width of 310 microinches was discernible when photographed in this manner. However, when the phosphor was placed between the holder and the camera, this pattern was no longer visible, but the next finer mesh was clearly "seen" through the phosphor. Since this mesh was a 500-line, 450-microinch mesh, the phosphor was assigned a mesh resolution figure of 450. These electroformed nickel meshes are available from Buckbee Mears, St. Paul, Minnesota, in web widths that cover the resolution range of settled phosphors; intermediate widths can be purchased on special order. Meshes with unequal horizontal and vertical webs that would effectively double the number of test patterns that could be mounted behind the phosphor are also available.

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* Received by the IRE, October 9, 1961. This work was done under the auspices of the U. S. Atomic Energy Commission.

No data were taken to correlate *mesh resolution figure* with *line pairs per millimeter*. For the purpose of evaluating settling techniques, the mesh system was simple, inexpensive, and relatively independent of the resolution of the system.

The author wishes to thank L. Mancebo and R. Kalibjian for their advice and Technician R. Reiss for his assistance.

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Analysis of AGC Loops*

In an earlier paper,[1] Victor and Brockman presented a detailed analysis of AGC loops. This communication contains an alternative method of derivation and some additional details of performance which may be of value in gaining a better understanding of loop behavior. The loop being analyzed has parts as shown in Fig. 1.

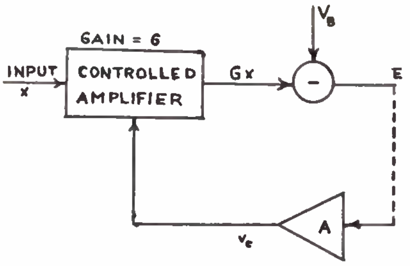

Fig. 1—Ports of AGC loop.

Definitions of Terms

x = Amplitude of input signal to IF amplifier.
G = Gain of controlled amplifier.
Gx = dc output of AGC detector.
V_B = AGC bias ("delay") voltage.
E, E' = dc error voltage.
A = Gain of AGC loop filter. (In general, A is frequency dependent.)
v_c = Control voltage applied to gain-controlled amplifier.

Loop Equations

$$E = Gx - V_B \tag{1}$$

$$v_c = AE. \tag{2}$$

We assume

$$G = G_0 10/\frac{KV_c}{20}, \tag{3}$$

where G_0 is gain with no signal present, and K is a gain-control constant with units of db/v.

Combining (1), (2) and (3) results in

$$E = xG_0 10^{KAE/20} - V_B. \tag{4}$$

Differentiate E with respect to x:

$$\frac{dE}{dx} = 10^{KAE/20}G_0\left[1 + \frac{xKA}{20}\ln 10\frac{dE}{dx}\right], \tag{5}$$

and take common logarithms of both sides of the equation:

$$\log\frac{dE}{dx} = \frac{KAE}{20} + \log G_0 + \log\left[1 + \frac{xKA}{20}\ln 10\frac{dE}{dx}\right]. \tag{6}$$

Differentiate (6) with respect to x once again:

$$\frac{\frac{d^2E}{dx^2}}{\frac{dE}{dx}\ln 10} = \frac{KA}{20}\frac{dE}{dx} + \frac{\frac{KA}{20}\ln 10\left[\frac{dE}{dx} + x\frac{d^2E}{dx^2}\right]}{\ln 10\left[1 + x\frac{KA}{20}\ln 10\frac{dE}{dx}\right]}, \tag{7}$$

and manipulate to obtain the nonlinear differential equation of the AGC loop:

$$\frac{d^2E}{dx^2} - 2\frac{KA\ln 10}{20}\left(\frac{dE}{dx}\right)^2 - x\left(\frac{KA\ln 10}{20}\right)^2\left(\frac{dE}{dx}\right)^3 = 0. \tag{8}$$

This formidable-appearing expression can be shown to have a solution:

$$E = -\frac{20}{KA}\log\frac{x}{x_0}, \tag{9}$$

where x_0 is the amplitude of the input signal required in order to just barely overcome the AGC delay bias and initiate gain control action. That is,

$$x_0 = \frac{V_B}{G_0}. \tag{10}$$

To the extent that (3) is correct, the result for static error (9) is exact and involves no linearizing approximation. Of course, the assumption (3) of the exponential gain-control characteristic is not really exact, even though it may be very nearly so in a precision AGC loop.

Gain error in db is given by

$$20\log\left(1 + \frac{E}{V_B}\right). \tag{11}$$

These expressions are readily applied to calculate steady-state errors under static conditions. However, under dynamic conditions the mathematics quickly become intractable and recourse must be had to a linear approximation. The following argument is intended to derive the linearized open-loop gain.

Assume a steady signal $x(x > x_0)$ causes an error voltage E. In a hypothetical experiment E is measured, the loop opened at the dashed lines, and a voltage $E^1 = E$ applied for control purposes. No steady-state conditions have been altered, so the error voltage remains at E.

The small-signal, open-loop gain is defined as the ratio of the change in E, due to a change in E^1. From (1), (2), and (3),

$$E = G_0 10^{KAE'/20}x - V_B, \tag{12}$$

so the open-loop gain is

$$\frac{dE}{dE^1} = xG_0 10^{KAE'/20}\frac{KA}{20}\ln 10 = xG\frac{KA}{20}\ln 10 \tag{13}$$

which is still a nonlinear function, since $G = G(x)$.

However, if AGC action is effective, E is very small and xG is approximately equal to V_B. Therefore, the linearized open-loop gain is given by

$$\frac{dE}{dE'} = \frac{V_BKA}{20}\ln 10 = 0.115\,V_BKA. \tag{14}$$

This result agrees with that derived by Victor and Brockman, as, indeed, it should. Conventional feedback analysis may now be applied, in order to determine the closed-loop transfer function and dynamic loop behavior.

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* Received by the IRE, September 19, 1961.

[1] W. K. Victor and M. H. Brockman, "The application of linear servo theory to the design of AGC loops," Proc. IRE, vol. 48, pp. 234–238; February, 1960.

Air-to-Undersea Communication*

Since the classical paper of Sommerfeld,[1] many problems have been solved with varying degrees of accuracy involving the determination of electromagnetic fields caused by a transmitter located in air above an arbitrary "ground." In a recent report[2] the author has obtained solutions for the Sommerfeld integrals over the "quasi-near" range using an extension of the method employed by Baños and Wesley[3] for the fields produced by a submerged antenna.

The method depends on the evaluation of certain auxiliary integrals, in terms of asymptotic series, obtained by the saddle-point method of integration in the complex plane. The integrals and the paths of integrations are carefully chosen such as to permit term-wise differentiation or integration to obtain new asymptotic series for their derivatives or integrals. By choosing suitable arguments of the relevant functions and by carrying out the operations indicated in the relevant expressions, two of the integrals (U, V) developed by Baños and Wesley[4] can

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* Received by the IRE, October 9, 1961. This work was performed under Contract Nonr 2798(01).

[1] A. Sommerfeld, "Über die Austreitung der Wellen in der drahtlosen Telegraphie," *Annalen der Physik*, vol. 28, p. 44; 1909.

[2] S. H. Durrani, "Air-to-Undersea Communication: Electromagnetic Fields in the Two Media, Caused by Vertical and Horizontal Electric Dipoles in Air," Engrg. Exper. Sta., University of New Mexico, Albuquerque, Tech. Rept. EE-61; September, 1961.

[3] A. Baños and J. P. Wesley, "The Horizontal Electric Dipole in a Conducting Half-Space," Marine Physical Lab. Repts. No. 55-33 and 54-31: University of California, Los Angeles, 1953–54.

[4] *Ibid*. Chs. 8, 9.

be modified to yield values of the $\bar{\pi}$-vector components for the present problem.

The "quasi-near" range is defined as the range which is less than a wavelength in air but more than a wavelength in the sea; that is, $k_ar<1<|k_sr|$. Here subscripts a and s refer to air and sea, k is the wave-number $(\mathrm{Re}\,\bar{k}>0)$, and r the range. (See Fig. 1, which shows both rectangular and cylindrical coordinates.) The fields are required at P for an electrical dipole—vertical or horizontal—radiating from S. A time dependence of $e^{j\omega t}$ is assumed and rationalized MKS units are used. Superscripts v and h stand for the vertical and horizontal dipoles, respectively. Then Sommerfeld's results[1,5,6] for the first case become

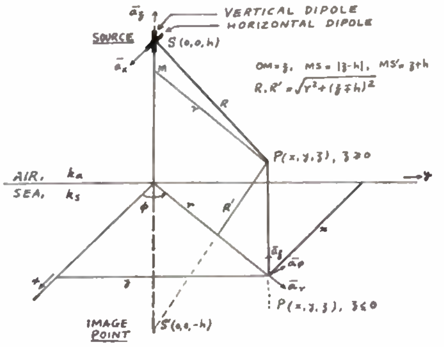

Fig. 1—Coordinate system.

$$\pi_{za}{}^v = \frac{e^{-jk_aR}}{R} + \frac{e^{-jk_aR'}}{R'} - 2\int_0^\infty \frac{J_0(\lambda r)e^{-\gamma_a(z+h)}}{n^2\gamma_a+\gamma_s}\frac{\gamma_s}{\gamma_a}\lambda d\lambda$$

$$\pi_{zs}{}^v = 2\int_0^\infty \frac{J_0(\lambda r)e^{\gamma_s z-\gamma_a h}}{n^2\gamma_a+\gamma_s}\lambda d\lambda,$$

where $\gamma_i=\sqrt{\lambda^2-k_i^2}$, Re $\gamma_i>0$, $i=a$, s, and $n^2=k_s^2/k_a^2$; r, R, R' as in Fig. 1. Similar expressions are obtained for $\pi_{xa}{}^h$, $\pi_{xs}{}^h$, $\pi_{za}{}^h$ and $\pi_{zs}{}^h$.

On applying the method outlined above, all $\bar{\pi}$ components are evaluated to within an error of 0.5 per cent. Use of the familiar formulas

$$\bar{E} = \nabla\nabla\cdot\bar{\pi} + k^2\bar{\pi}, \qquad \bar{H} = j\omega\underline{\epsilon}\nabla\times\bar{\pi}$$

leads to the following approximate results for $\bar{E}, \bar{H}$ components in the quasi-near range with h, $|z|<r$:

$$E_{ra}{}^v \simeq \frac{3r(z-h)}{R^5} + \frac{3r(z+h)}{R'^5} - \frac{2jk_s}{n^2r^2}$$

$$E_{za}{}^v \simeq -\frac{r^2-2(z-h)^2}{R^5} - \frac{r^2-2(z+h)^2}{R'^5}$$

$$H_{\phi a}{}^v \simeq j\omega\epsilon_0\left(\frac{r}{R^3}+\frac{r}{R'^3}\right)$$

$$E_{rs}{}^v \simeq -2e^{jk_sz}jk_s/(n^2r^2)$$

$$E_{zs}{}^v \simeq -2e^{jk_sz}/(n^2r^3)$$

$$H_{\phi s}{}^v \simeq j\omega\epsilon_0e^{jk_sz}/r^2$$

and

$$E_{ra}{}^h \simeq \cos\phi\left\{\frac{2}{R^3}-\frac{2}{R'^3} - 2[1+jk_s(z+h)]/(n^2r^3)\right\}$$

$$E_{\phi a}{}^h \simeq \sin\phi\left\{\frac{1}{R^3}-\frac{1}{R'^3} + 2[1+jk_s(z+h)]/(n^2r^3)\right\}$$

$$E_{za}{}^h \simeq \cos\phi\left\{\frac{3r(z-h)}{R^5}-\frac{3r(z+h)}{R'^5}\right\}$$

$$E_{rs}{}^h \simeq \cos\phi e^{jk_sz}\{2(1-{}^{jk_sh})/(n^2r^3)\}$$

$$E_{\phi s}{}^h \simeq \sin\phi e^{jk_sz}\{2(2+jk_sh)/(n^2r^3)\}$$

$$E_{zs}{}^h \simeq \cos\phi e^{jk_sz}\{-6h/(n^2r^4)\}.$$

Similarly for the H^h components.[2]

A comparison of the E fields shows that a vertical electric dipole produces the strongest component in the sea, this being the horizontal component $E_{rs}{}^v$. It has previously been shown by Moore and Blair[7] that a horizontal submerged transmitter is better than a vertical one and produces the strongest field component in air along the vertical direction. Combining the two results, it is concluded that from field-strength considerations alone, the best intercommunication system is obtained by using a vertical electric dipole in air along with a horizontal electric dipole in the sea.

Using the field-strength values given by the above formulas, simplified estimates can be made of the communication range, *i.e.*, the distance over which satisfactory signal strength is obtained, provided noise figures, bandwidth, temperature and other data regarding transmitter and receiver are available. An example[2] illustrates the method.

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[5] A. Sommerfeld, "Partial Differential Equations in Physics," Academic Press, Inc., New York, N. Y., §§32–33; 1949.

[6] J. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y.; 1941.

[7] R. K. Moore and W. E. Blair, "Dipole radiation in a conducting half-space," *J. Res. NBS*, Sec. D, vol. 65, pp. 547–563; November–December, 1961.

Symmetrical RC Distributed Networks*

The class of distributed RC networks, more specifically distributed RC ladder networks, which is electrically symmetrical can be identified by the characteristic that the distributed series resistance is proportional to the distributed shunt capacitance. The proof of this is presented and the symmetrical network analyzed.

Proof

The linear differential equation for the model of the distributed RC ladder network is

$$v'' - (r'/r)v' - srcv = 0,$$

where r is the series distributed resistance per unit length and c the shunt distributed capacitance per unit length and where v is the voltage across the network which has an assumed time dependence of e^{st}. The prime denotes differentiation with respect to x.

Let the solution for the symmetrical line be in the form

$$v(x,s) = v_1D(x,s) - i_1B(x,s)$$

in which v_1 and i_1 are the necessary two arbitrary constants, in this case the voltage and current at the input of the network. The choice of the symbols D and B is consistent with their interpretation as the chain parameters of a network of length x.

The current in the series resistance of the network is related to the voltage by

$$i(x,s) = -v'(x,s)/r(x);$$

consequently,

$$i(x,s) = -v_1D'(x,s)/r(x) + i_1B'(x,s)/r(x).$$

If the current is written in terms of chain parameters, *i.e.*,

$$i(x,s) = -C(x,s)v_1 + A(x,s)i_1,$$

then the identification

$$A(x,s) = B'(x,s)/r(x)$$

can be made.

For a two-port network to be symmetrical $A=D$, hence

$$D(x,s) = B'(x,s)/r(x).$$

Substituting this back into the original form for the voltage, and dropping the functional dependence symbols,

$$v = v_1(B'/r) - i_1B.$$

Thus,

$$v' = v_1[(B''r - B'r')/r^2] - i_1B'$$

and

$$v'' = v_1[(B'''r - B'r'')/r^2 - 2(B''r - B'r')r'/r^3] - i_1B''.$$

As this is, by hypothesis, a solution to the differential equation make the substitutions and regroup:

$$(v_1/r)\{B''' - 3B''(r'/r) + B'[3(r'/r)^2 - (r''/r) - src]\} - i_1\{B'' - (r'/r)B' - srcB\} = 0.$$

Since v_1 and i_1 are arbitrary, the bracketed groups of terms above must both vanish. If the second is differentiated with respect to x and then subtracted from the first bracketed group the result is

$$B'' - (r'/r)B' - (1/2)s(r/r')(rc)'B = 0.$$

Comparing this with the second bracketed group, *i.e.*,

$$B'' - (r'/r)B' - srcB = 0,$$

* Received by the IRE, October 3, 1961. This material represents a portion of the work in progress on a Ph.D. dissertation at the Polytechnical Institute of Brooklyn, Brooklyn, N. Y.

it is seen that,

$$src = (1/2)s(r/r')(r'c + rc');$$

hence,

$$(r'/r) = (c'/c)$$

and finally

$$r(x) = kc(x)$$

$$k = \text{constant}.$$

Thus, the electrically symmetrical RC distributed ladder network has its series distributed resistance proportional to its shunt distributed capacitance.

Analysis of "Proportional" Network

Let both series resistance and shunt capacitance have the same functional dependence $f(x)$ on the distance along the line, *i.e.*,

$$r = r_0 f(x) \qquad c = c_0 f(x).$$

The differential equation in this case becomes

$$v'' - (f'/f)v' - sr_0c_0f^2v = 0.$$

Reduction to a Riccati equation by the usual[1] substitution

$$v' = uv$$

leads to

$$u' + u^2 - (f'/f)u - sr_0c_0f^2 = 0.$$

However, it is more fruitful if the substitution is made

$$v' = yfv,$$

leading to the separable equation,

$$y' + (y^2 - sr_0c_0)f = 0$$

$$dy/(sr_0c_0 - y^2) = fdx$$

and its solution

$$y = \sqrt{sr_0c_0}\tanh\left(\sqrt{sr_0c_0}\int_0^x fdx + K\right).$$

Incidentally, $y = v'/fv = -r_0/(v/i)$ is the normalized admittance at any point along the network.

Continuing,

$$dv/v = f\sqrt{sr_0c_0}\tanh\left(\sqrt{sr_0c_0}\int_0^x fdx + K\right)dx$$

$$v = v_0 \operatorname{Cosh}\left(\sqrt{sr_0c_0}\int_0^x fdx + K\right)$$

in which v_0 and K are constants of integration. The current is found from the relation

$$i = -(v'/r_0f);$$

it is

$$i = -\sqrt{sc_0/r_0}\,v_0 \operatorname{Sinh}\left(\sqrt{sr_0c_0}\int_0^x fdx + K\right).$$

Substitute the initial conditions of $v = v_1$ and $i = i_1$ when $x = 0$, and place the voltage and current equations into a form suitable for identification of the network chain parameters:

$$v = v_1 \operatorname{Cosh}\sqrt{sc_0/r_0}\int_0^x rdx$$

$$- i_1\sqrt{r_0/sc_0}\operatorname{Sinh}\sqrt{sc_0/r_0}\int_0^x rdx$$

$$i = -v_1\sqrt{sc_0/r_0}\operatorname{Sinh}\sqrt{sc_0/r_0}\int_0^x rdx$$

$$+ i_1 \operatorname{Cosh}\sqrt{sc_0/r_0}\int_0^x rdx.$$

Therefore, the chain parameters of the network are

$$A = \operatorname{Cosh}\sqrt{sc_0/r_0}\int_0^x rdx$$

$$B = \sqrt{r_0/sc_0}\operatorname{Sinh}\sqrt{sc_0/r_0}\int_0^x rdx$$

$$C = \sqrt{sc_0/r_0}\operatorname{Sinh}\sqrt{sc_0/r_0}\int_0^x rdx$$

$$D = \operatorname{Cosh}\sqrt{sc_0/r_0}\int_0^x rdx$$

Note that the uniform network for which $f = 1$, *i.e.*, $r = r_0$, $c = c_0$, is in this class since the series resistance is proportional to the shunt capacitance. It is seen that the above relations for the chain parameters give the correct results in this special case.

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[1] M. Golomb and M. Shanks, "Elements of Ordinary Differential Equations," McGraw-Hill Book Co., Inc., New York, N. Y., p. 246; 1950.

Higher Modes in Millimeter Wave Isolators*

In millimeter wavelengths, it is almost impossible to insert a probe into the inside of the waveguide to measure the field distribution because of the extremely small waveguide size. One of the purposes of this note is to report that a new method can be used for obtaining the general field distribution in a millimeter wave field displacement isolator. Other observations of the experimental results obtained by this method are also reported. The method involved changing the mounting position of a ferrite strip.

Some of the effects of ferrite strip mounting for millimeter wave field displacement isolators which employed single crystal barium ferrite have previously been reported.[1] In this case a polycrystalline titanium cobalt substituted barium ferrite,[2] $[B_aO \cdot 0.1(T_iC_0O_3) \cdot 5.9\ Fe_2O_3]$, was used in a field displacement isolator. The polycrystalline ferrite was prepared by the usual ceramic processes. The sample used was cut in the form of a slab whose *E* plane coincided with the direction of pressing. The anisotropy field of this ferrite was estimated to be 16 kilo-oersteds using the data of DiBitetto, *et al.*[3]

Using the isolator configuration shown in Fig. 1, data of forward and reverse attenuation as a function of ferrite strip distance from the guide wall was taken at a frequency of 58 kMc and an applied dc magnetic field of 5000 oersteds. From the results, shown in Fig. 2, it is seen that the forward attenuation falls off much faster than the backward attenuation at distances ranging from 0.01 inch to 0.03 inch from the guide wall. Also, the data indicates an optimum mounting position of 0.081 L which, to our knowledge, is closer to the guide wall than had been previously reported by other authors.[1,4–7]

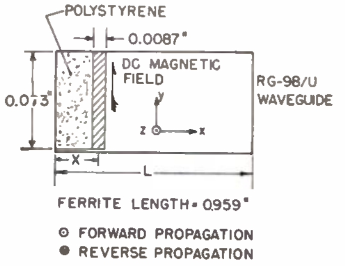

Fig. 1—Field displacement isolator mounting configuration.

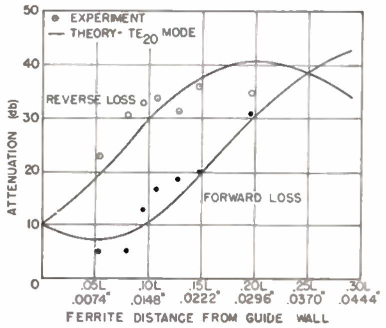

Fig. 2—Propagation loss in both directions as a function of ferrite distance from the guide wall. For the theoretical curve, $A = 40$ db, $B = 10$ db, $C = 10$ db.

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* Received by the IRE, October 16, 1961. This work was supported by an A. O. Smith Corp. research grant.

[1] K. Ishii, J. B. Y. Tsui, and F. F. Y. Wang, "Effects of ferrite strip mounting positions on millimeter wave isolator characteristics," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES (Correspondence), vol. MTT-9, p. 362; July, 1961.

[2] Supplied by A. O. Smith Corp.

[3] D. J. DiBitetto, *et al.*, "Hexagonal Magnetic Material for Microwave Applications," Philips Labs., New York, N. Y., Rept. No. 2, U. S. Army Signal Corps Contract No. DA 36-039 SC-85279; July 14, 1960.

[4] R. F. Soohoo, "Theory and Application of Ferrites," Prentice Hall Inc., Englewood Cliffs, N. J.; 1960.

[5] B. Lax, "Frequency and loss characteristics of microwave ferrite devices," PROC. IRE, vol. 44, pp. 1368–1386; October, 1956.

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

[7] K. J. Button, "Theoretical analysis of the operation of the field displacement isolator," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-6, pp. 303–308; July, 1958.

Similar results were repeatedly noted with other samples of the same ferrite material.

It appears from Fig. 2 that the directivity will change somewhere between 0.20 and 0.25 L indicating that the TE$_{20}$ mode and higher modes are propagated in this region. Comstock and Fay[8] have shown by theory and the probe method at a lower frequency (6.175 kMc) that for low values of resistive coating on the ferrite strip the TE$_{10}$ mode is highly attenuated and that all propagation is in the TE$_{20}$ mode or higher modes. The data obtained at 53 kMc agrees with their findings, although no resistive coating was used on the ferrite strip; the ferrite itself being used as the dissipative material. At X band, the field probe measurements of Angelakos[9] for a ferrite without resistive coating also indicated that the TE$_{20}$ mode is propagated.

The general shape of the experimental curves indicated that an equation developed by Lax[5] for the propagation loss in the two directions as a function of ferrite distance from the guide wall could be used to approximate the experimental data. This equation is

$$\alpha_{\pm} = A \sin^2 kx + B \cos^2 kx \pm C \sin^2 kx,$$

where

$$A = C'\chi_{xx}''\beta^2$$
$$B = C'\chi_{yy}''k^2$$
$$C = C'\chi_{xy}''\beta k$$
$$k = \frac{n\pi}{L}\cdot$$

Using the TE$_{20}$ mode in this equation the "constants" A, B, and C were chosen to get an approximate fit to the experimental data shown in Fig. 2. Although this equation cannot fit the experimental data perfectly, this equation used in conjunction with higher modes probably could give a better fit.

Thus, at millimeter wavelengths propagation in a field displacement isolator is mainly in the TE$_{20}$ mode with the TE$_{10}$ mode attenuated as was found by Comstock and Fay at 6 kMc. Also, Lax's equation, verified at X band, can be used to describe the backward and forward losses of a field displacement isolator at millimeter wavelengths.

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[8] R. L. Comstock and C. E. Fay, "Operation of the field displacement isolator," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 605–611; November, 1960.

[9] D. J. Angelakos, "Transverse electric field distributions in ferrite loaded waveguides," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 390–391; July, 1959.

Contributors

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Loren S. Bearce was born in Brown County, Kan., on March 22, 1927. He received the B.S. degree in physics, with honors, from Kansas State University, Manhattan, Kan., in 1949. His subsequent graduate study has been at the University of Maryland, College Park.

From 1945 to 1946, he served in the U. S. Navy as an aviation electronics technician. In 1948, he joined the scientific staff of the Radio Techniques Branch of the U. S. Naval Research Laboratory, Washington, D. C., where he has been engaged as an Electronic Scientist in research on radio communication systems. His initial projects were in the field of instrumentation, particularly development of precision capacitance standards and improved techniques of communication receiver measurement. Later, he was concerned with speech compression techniques, receiver detection circuitry, and the development of more linear RF amplifiers with large dynamic range to improve the strong signal selectivity of communication receivers. Much of his more recent work has been in the field of UHF, VHF, and VLF electromagnetic wave propagation; in connection with this work, he has developed an automatic programming system for NRL's NAREC high-speed digital computer. His present activity in VLF propagation research stems from a project for extension of the theoretical treatment of submerged communications reception. He has contributed to many phases of LOFTI I, including design and technical coordination of the experiment and reduction and analysis of the data.

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Donald L. Birx was born in Baltimore, Md., on November 30, 1924. He received the B.E.E. degree from The Johns Hopkins University, Baltimore, Md., in 1944. In 1945 he entered the U. S. Navy and was employed as an Instructor in the Radio Materiel School, Washington, D. C. In 1946 he re-entered The Johns Hopkins University and received the Dr.Engrg. degree in electrical engineering in 1951.

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Since 1951 he has been with the Franklin Institute, Philadelphia, Pa., where he is the Manager of the Electronics Laboratory. He has been working on accurate measurements of low-level RF powers, and the utilization of natural radiation in the navigation of space vehicles. Prior to this work, he was engaged in a variety of projects which have included the statistical design of experiments for the analysis of optical and electrical systems, the synthesis of networks utilized in the design of RF noise filters, and the study of nonlinear aspects of rate gyroscopes.

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He has been in the Radio Division of the U. S. Naval Research Laboratory, Washington, D. C., since 1942. From 1942 to 1950 he worked in the field of microwave techniques and instrumentation where he contributed to the early development of microwave signal generators and devised methods and standards for measuring power, attentuation, frequency, harmonic content, impedance and modulation at microwave frequencies. From 1951 to 1958, he was Head of the Information Conversion and Reproduction Section and devised and developed techniques and systems to utilize computer methods for data transmission and processing. This work included a system permitting radar operators to extract data from a display in digital form and a completely automatic system called MATALOC (MAgnetic TArget LOCator) which would determine the center of radar targets in cartesian coordinates. In 1958, he became Head of the Satellite Communications Section of the Radio Techniques Branch and is presently directing the Laboratory's effort in four specific areas: transionospheric propagation of VLF radio waves (LOFTI), communication by passive satellite reflection, communication by moon relay (CMR), and participation in the 24-hour satellite program (ADVENT).

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Mr. Shen worked for Hewlett-Packard Company, Palo Alto, Calif., the summers of 1959 and 1960.

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Emerick Toth was born in New York, N. Y., on June 27, 1905. He received the B.S.E.E. degree from the Newark College of Engineering, Newark, N. J., in 1927.

For the following 11 years he was employed as physicist, research engineer, and design engineer by Bakelite Corporation, Bell Telephone Laboratories, DeForest Radio Company, Sylvania, Bendix Radio, and other industrial organizations. During this period he worked on the development of new RF measurement techniques for insulating materials, design of commercial radio receiving equipment (police and high-fidelity types), and development of military communication receivers and direction-finders (shore, shipboard, and airborne types). In 1938 he joined the scientific staff of the U. S. Naval Research Laboratory, Washington, D. C., as a communication receiver specialist. His projects have included development of sonar receiving equipment, research on underwater radio receiving systems and development of new techniques of communication system analysis, in addition to advanced receiver development work. Since 1957 he has been Head of the Radio Techniques Branch of NRL and presently directs research and development in satellite communication, centralized electronic control, precise frequency and time instrumentation, and communication system analysis.

E. Toth
SM'48-F'58

Mr. Toth has received the Navy Meritorious Civilian Service Award for development of VHF and UHF receivers during World War II. He has served as guest lecturer at the U. S. Naval Postgraduate School and Cornell University. He is a member of the Research Society of America and Commission VI of URSI. He is serving the IRE as a member of the Administrative Committe of PGCS, Awards Chairman of the Administrative Committee of PGMIL, and a member of the Awards Committee of the Washington Section of IRE.

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Toshio Utsunomiya was born in Nakatado-gun, Kagawa, Japan, on November 20, 1921. He received the B.Engrg. degree from the University of Tokyo in 1943 and the Engrg.D. degree in electrical engineering in 1961.

After receiving his Bachelor's degree, he remained with the University of Tokyo as a member of the teaching staff in electrical engineering. In 1948, he was appointed Assistant Professor of Electrical Engineering and Professor in 1961. In addition to his teaching assignments, he has undertaken considerable research. He has worked on microwave sources and power measurement and has conducted extensive research on dc

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M'60

amplifiers. More recently, he has carried on basic research on semiconductor and transistor circuitry. In 1959, he was granted a leave from the University of Tokyo to accept a one-year appointment to the research staff of the Electronics Research Laboratories of Columbia University, New York, N. Y. During this period his major effort was devoted to research on variable-reactance frequency multipliers. He also did significant work on the automatic control of dc amplifiers.

Dr. Utsunomiya is a member of the IEE of Japan, the Institute of Communication Engineers of Japan, and the Institute of Television Engineers of Japan.

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James R. Wait was born in Ottawa, Ont., Canada, on January 23, 1924. During World War II, he was in charge of a radar maintenance group at Kingston, Ont. He received the B.A.Sc. and M.A.Sc. degrees in engineering physics in 1948 and 1949, respectively, and the Ph.D. degree in electromagnetic theory in 1951, all from the University of Toronto, Ont. During this time he was associated with Newmont Exploration Ltd. of New York and Jerome, Ariz.

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From 1952 to 1955, he was a Section Leader in the Defence Research Telecommunications Establishment, Ottawa, primarily concerned with theoretical problems in radiation. Since joining the National Bureau of Standards, Boulder, Colo., in 1955, he has concentrated on the theoretical aspects of LF radio propagation. He is presently Consultant on radio wave propagation in the Office of the Director of the Boulder Laboratories. From 1958 through 1961, he served as the initial Editor of the *Radio Propagation Journal* (Section D of the *NBS Journal of Research*). He is now an Associate Editor of the *Journal of Geophysical Research.*

In the fall of 1960, he visited the Technical University of Denmark, Copenhagen, to help establish a research laboratory in electromagnetic theory. In 1961, he was appointed Professor (adjoint) of electrical engineering by the University of Colorado, Boulder. He is Chairman of the technical program of the URSI Symposium on Electromagnetic Theory to be held in Copenhagen in June, 1962.

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Shui Yuan was born in Kwang Tung, China, on June 15, 1931. He received the B.S.E.E. degree from the University of California, Berkeley, in 1952, and the M.S.E.E. degree from Columbia University, New York, N. Y., in 1956.

From 1953 to 1955, he served on the technical staff of the Radio Corporation of America, Camden, N. J., where he was engaged in the design and development of the UHF-slotted-cylindrical television broadcasting antennas and related components, including waveguides and transmission lines. Later, he became associated with the Electronics Division of the Curtiss Wright Corporation, Caldwell, N. J., where he was engaged in their systems development program, designing systems for simulators, including switching circuits, analog computers, and servos. Since 1958, he has served in the capacity of Senior Research Engineer at the Electronics Research Laboratories of the School of Engineering, Columbia University, New York, N. Y., where his work includes advanced radar research with particular emphasis upon noise problems and the development of frequency multipliers employing variable reactances.

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Robert W. Zeek was born in Paterson, N. J., on March 12, 1920. He received the B.S.E.E. degree from the Newark College of Engineering, Newark, N. J., in 1943.

He joined the scientific staff of the U. S. Naval Research Laboratory, Washington, D. C., in 1943. His early work concerned primarily the evaluation of design and performance of Naval communication receivers. From 1945 to 1947 he was engaged in development of panoramic adaptors and frequency-shift converters for communication receivers. From 1947 to 1949, he participated in NRL's comprehensive investigation of the relative merits of AM and FM for naval communications, contributing a basic study and analysis of the mechanism of capture effect.

R. W. ZEEK
M'47–SM'55

In 1950 he led a group developing remote radio control means for synchronization of minesweepers and completed a research study of highly selective mechanical and electrical RF filter systems. He then directed the development of systems for communication receiver remote control by radio and wire link until 1955. From 1955 to 1959, he headed a research effort to extend theoretical treatment of the propagation of VLF radio signals from terrestrial transmitters to underwater antennas. Since 1959, he has provided technical mangement and coordination of the LOFTI satellite program and is now Technical Program Coordinator for all communication satellite projects of the Radio Techniques Branch.

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Books

Linear Algebra and Group Theory, by V. I. Smirnov; R. A. Silverman, Ed.

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 418 pages+6 index pages+x pages+2 bibliography pages+10 appendix pages+27 pages hints and answers. Illus. 6¼×9¼. $12.50.

This book is a selection of material from Professor Smirnov's encyclopedic six volume "Course of Higher Mathematics." The present volume concerns itself with the usual topics of linear algebra—determinants, matrices and systems of equations—and also contains an introduction to group theory. A summary of the contents follows.

Chapter 1 discusses determinants and their properties including methods of expansion and the multiplication theorem. In Chapter 2, the author examines the solution of systems of linear equations. Topics include systems of m equations in n unknowns, linear vector spaces, the Gram determinant, Hadamard's inequality, linear differential equations with constant coefficients, and Jacobians.

Chapter 3, on linear transformations, considers rotations in three dimensions, general linear transformations, tensor concepts including convariant and contravariant vectors and cartesian tensors, elements of matrix algebra, unitary and orthogonal transformations, Schwarz's inequality, scalar product and norm. Chapter 4 deals with quadratic forms. The subjects discussed are reduction of a quadratic form to a sum of squares, classification of quadratic forms, simultaneous reduction of two quadratic forms, Jacobi's formula, small oscillations, extremal properties of the eigenvalues, hermitian and unitary matrices, functions of matrices.

In Chapter 5, the author investigates infinite-dimensional spaces. He discusses Hilbert and function spaces and their relation, convergence of vectors, complete systems of orthonormal vectors, linear transformations in infinitely many variables, and linear operators. Chapter 6 contains the proof of the classical reduction of any matrix to a canonical form. A lengthy numerical example is included to illustrate the procedure.

Chapter 7 is an introduction to the general theory of groups. The topics considered are groups of linear transformations, polyhedral groups, Lorentz transformations, permutations, abstract groups, subgroups, classes and normal subgroups, isomorphic and homomorphic groups, the unitary group and the rotation group, the unimodular group and the Lorentz group. Chapter 8 contains an introduction to the theory of group representations. Among the subjects treated here are the basic theorems, Abelian groups and one-dimensional representations, representations of the two-dimensional unitary group, of the rotation group and of the two-dimensional unimodular group, direct products, the orthogonality property, characters, regular representation of a group, and proofs that the rotation group and the Lorentz group are simple. The elements of the theory of continuous groups are developed in Chapter 9. The topics under discussion include continuous groups and their structure constants, infinitesimal transformations, construction of a group from its structure constants, integration on a group and the orthogonality property.

The exposition is characterized by concreteness and clarity. The editor of the English edition has increased its utility greatly by the addition of a large collection of problems, many with answers or hints for the solution.

The first four chapters and Chapter 6 comprise the usual course in linear algebra (oriented towards applications) on a senior undergraduate or possibly first-year graduate level. The material on group theory is somewhat unusual. Here, besides the general theory, the specific groups discussed are those of greatest interest to physics. The chapter on group representations is an excellent extensive introduction to the subject. The material on continuous groups is largely introductory but also discusses integration on a group.

The development of the material treated in the book is characterized by a concrete approach which should make it more accessible to the non-specialist. The orientation of the author is always directed towards applications of the algebra to analysis, geometry and physics. It should prove a welcome addition to the library of the applied mathematician, engineer or physicist.

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Elektronisches Rauschen, Teil I, by H. Pfeifer.

Published (1959) by B. G. Teubner Verlagsgesellschaft, Leipzig, Germany. 248 pages+13 index pages +vi pages+23 bibliography pages+16 appendix pages. Illus. 6×8¼.

It is very worthwhile to have another book available on noise which covers as wide a scope as this one since, at the present time, there are only a few adequate textbooks. The present book (Volume I) deals with the basic aspects of noise in semiconductors, semiconductor devices, vacuum tubes, and high-frequency devices such as klystrons, traveling wave tubes, masers, etc. This book will be useful for a course in noise phenomena, as well as for the experimenter in this field. Both experimental and theoretical material is presented very completely within the intended scope of the book. Of special merit is the excellent list of references, including current literature. The theory is mainly directed towards the device under discussion, in a manner which may be desired by the engineer or experimental physicist. Apparently, the book does not aim at exposing broadly the fundamentals of stochastic processes; hence, the theoretician should not advise this book for those aspects of noise which are linked with statistical mechanics, irreversibility, Markoffian chains, etc. In view of this, the mathematical introduction has been kept short (12 pages) which is not satisfactory in the reviewer's opinion (spectral analysis is *not* equivalent with Fourier analysis of a fluctuating quantity, but includes many more powerful tools).

The basic chapters on noise sources are well written and will acquaint one with the field of noise. At some places, however, the German *Gründlichkeit*, in compiling too many theories including antiquated ones, will lead to confusion. Examples are found in the discussion of 1/f noise (Bess' theory on dislocation pipes, etc.) and in the discussion of generation recombination noise in semiconductors. No clear distinction is made between shot effects (as occurring during ambipolar sweep-out) and thermal carrier fluctuations. More critical display would have been welcome in these instances.

The circuit aspect of noise has been kept short since it will form the contents of Volume II, which will possibly also be a welcome addition to the literature on noise.

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Functional Circuits and Oscillators, by Herbert J. Reich

Published (1961) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 419 pages+12 index pages+xix pages+2 appendix pages+28 pages of problems. Illus. 6¼×9¼. $12.50.

In the reviewer's opinion, this book is an excellent contribution to electronics engineering literature, which is characterized by many good to excellent books, and *very many* books that are mediocre or even less. While it seems to be somewhat too specialized for use as a college text, *per se*, it may find worthwhile application in graduate work and be useful as a supplementary reference text for a course in nonlinear circuits. Its chief value lies in the outstanding way in which it presents fundamental theory and technical information regarding a practically complete stable of elementary circuits, which a large and rapidly growing body of electronics speciality engineers must master in order to be in a position readily to synthesize the complicated circuit combinations that today's technology demands, and that will be even more essential in the future.

This book is a bit unusual in that it is not divided into chapters because, as the author states, such a breakdown "would be difficult and often artificial." There are 93 subject headings with 345 subheadings, plus 111 problems which are a very essential part of the text. In studying one of the several subject headings with which the reviewer was not familiar, an adequate understanding of the text was not obtained until after the referenced problems were worked out in detail. This seems to be a most excellent way of treating the subjects, and the text would probably have to be doubled in volume if text alone were to carry understanding without the problems.

This book gives an excellent and entirely adequate theoretical treatment of negative-resistance devices, summing circuits, differential amplifiers, differentiators, integrators, clippers, gates, and nonlinear circuit simulators, but is quite brief (although by no means superficial) in its treatment of sine wave oscillators. Colpitts and Hartley are mentioned, but presumably the author felt that their long history rendered a detailed treatment redundant, and therefore he confines himself to a brief generalized analysis. Mathematics, including differential equations, is used freely throughout the text, but not unnecessarily nor exhaustively, and only in cases where the mathematics makes a definite contribution. The author and subject indices are particularly well done and convenient, and the bibliography inserted as footnotes seems to be entirely adequate.

This book does not contain a compendium of "practical" circuits and in only a few cases are circuit valves mentioned; hence the "handbook engineer" will find little of use to him. However, the well grounded electronics engineer will find this book adequate as a foundation for specialization in any one of a number of electronics circuit fields.

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Plasmas and Controlled Fusion, by David J. Rose and Melville Clark, Jr.

Published (1961) by The Technology Press, Mass. Inst. Tech., Cambridge and John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. 463 pages +19 index pages+xiv pages+11 appendix pages+references and problems by chapter. Illus. 6×9¼. $10.75.

In this book, the recent flood of research papers on plasma physics has been carefully and fully digested and presented as an organized textbook. The book is clearly and avowedly intended as a textbook on the problems of controlled fusion, and appears to be a success in this aim. It is to be recommended further as a reference containing many of the derivations and descriptions previously available only in collections of papers with diverse notation and variable quality of presentation. A notable feature is that the descriptive text is clear and often appealing to the intuition, yet the derivations are complete and accurate to the best of the reviewer's knowledge. In the rare cases where mathematical details are left out, the reader is told in words the difficult steps, and references are given.

After an introductory chapter about energy sources, nuclear and non-nuclear, the book treats the basic parts of analytic plasma theory in the following chapters: 2) and 3) Collision Phenomena, 4) Velocity Distribution, 5) Maxwell's Equation, 6) Hydromagnetic Equations, 7) Macroscopic Motions, 8) Coulomb Interactions, 9) Plasma Waves of Small Amplitude, 10) Motion of Individual Charges, 11) Radiation Losses, 12) Plasma Stability. The final four chapters (13–16) are devoted to engineering problems and the three major confinement schemes, Pinch, Mirror and Stellerator devices. The appendix of vector relations, physical constants, conversion and glossary is useful.

The book has been found useful as a reference in which most of the fundamental derivations may be found. It would appear to be equally useful as a text, although this reviewer would have preferred the chapters in a different order.

An attempt was made to find errors and sections which might confuse the reader. A few marginal examples were found, but in this respect the book rates well above most current publications. It is to be recommended as a text and reference, and to those interested in seriously learning about plasma physics. To the latter group, a suggestion is offered that the book is thorough and can profitably be read carefully, equation by equation. The effort thus expended should be well rewarded.

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Transient Circuit Analysis, by Y. H. Ku

Published (1961) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 342 pages+9 index pages+xiv pages+88 appendix pages. Illus. 6¼×9¼. $13.00.

This book was written "to provide . . . engineers, physicists, and applied mathematicians with a board treatment of circuit analysis. . . ." The principal chapters are:

Solution of Ordinary Differential Equations
Linear Network Analysis
Fourier Series and Fourier Integral
Fourier Transform
Laplace Transform
Functions of a Complex Variable
Inverse Transform
Transients in Transmission Lines
The *Z*-Transform and Sampled-Data Systems

As suggested by the chapter headings, a more descriptive title would be "The Mathematical Basis of Modern Transient Circuit Analysis." The reader will find the text highly compressed and mathematical in character as befits a textbook prepared for a graduate level course on circuit analysis. The utility of the book as a reference work for the practicing engineer is therefore somewhat limited. Modern mathematical techniques for circuit analysis have provided many new tools for the system designer. In general, however, the "rules of the game" and the underlying physical concept are remembered long after the details of rigorous derivation and boundary conditions of the technique are forgotten. Occasionally this circumstance leads one into difficulty and, in such a situation Dr. Ku's book will be of assistance.

A common failing of mathematical texts directed towards a specific manifestation of the physical world is a preoccupation with mathematical manipulation and logic to the almost complete exclusion of the associated physical picture. As a simple example, while students we all learn sooner or later that the process of integration is an elegant way of computing the area under a curve. When the immediate recollections of our academic background have receded into the past it is usually the simple physical concept of "area" which enables us to recognize and to deal with a problem in which integration is required. By the same token, there are few, if any, modern mathematical techniques for circuit analysis which cannot be related to physical "things" such as time, amplitude, frequency, phase, growth, decay and the like. Concepts communicated with the assistance of physical imagery tend to endure long after the associated symbology has been forgotten. Dr. Ku's book is as deficient as many others in this respect. In this case it is particularly unfortunate because of the great wealth of material presented.

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Meteor Science and Engineering, by D. W. R. McKinley

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 283 pages+13 index pages+ix pages+13 bibliography pages+4 appendix pages. Illus. 6¼×9¼. $12.50.

In recent years there has been an increasing effort on the part of astronomers and radio scientists and engineers to determine the distributions and dynamics of meteoric particles in space, and their effects in the upper atmosphere. Progress has been rapid, with the newer techniques of radio augmenting the spectrum of measurable meteoric phenomena, and complementing other basic studies previously made only by visual and telescopic methods. Dr. McKinley's book, "Meteor Science and Engineering," provides a current status report on the methods, results, and applications of the radio and optical studies of meteors, with emphasis on the radio investigations. It comes at the right time for such a report, since we are just now entering a new phase in the study of meteoric particles, based on the use of space vehicles.

The first four chapters survey the history of meteor observations, provide some basic material on astronomy and radio, and summarize the visual, photographic, and radio techniques used in the basic studies. The next chapter presents observational data on meteors, although other results are included in other chapters to illustrate techniques or theoretical discussions.

In Chapter 6 the astronomical aspects of meteors are presented. For a more complete discussion of this subject, the reader may refer to Lovell's 1954 Oxford Press book, "Meteor Astronomy." The physical theory of meteors is summarized in the next chapter. Again there is a book available for the reader who may wish to expand on this subject—Öpik's, "Physics of Meteor Flight in the Atmosphere," Interscience Publishers, 1958.

The longest chapter (8) is on the theory of radio echoes from the ionized trails produced by meteoric particles in the upper atmosphere. The unified treatment and discussion of this material, most of which is not otherwise available in book form, provides a valuable reference source for those who are acquainted with the subject, and a training ground for newcomers to the discipline. In the discussion of the application of the theory to the highly variable measured quantities, there is a general emphasis on empiricism, a fact that the author does not fail to point out.

Forward scatter from meteor trails is the subject of the next chapter, where the applications of such scattering to new techniques in radio communications are emphasized. The final chapter treats briefly such other meteoric effects as their possible connection with rainfall, and their threat to space travel.

In his book McKinley covers the middle ground between the scientist and the engineer concerned with the techniques and results of studies and applications of various meteoric phenomena. It should prove valuable to specialists in any way concerned with meteors, and to the scientific layman. In his own work the author has contributed much to both the scientific and engineering areas. He is liberal in his references to the work of others. His writing is entertaining as well as factual, and the final product appears to have unusually few misprints.

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Linear Vacuum-Tube and Transistor Circuits, by A. J. Cote, Jr. and J. B. Oakes

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 402 pages+9 index pages+xxvi pages+references by chapter. Illus. 6⅛×9¼. $10.75.

Most of the several books that have been published recently concerning transistors cover both the device and applications to a greater or lesser degree. The present book is concerned only with the applications of transistors and, as the title indicates, only the linear circuit applications. Of necessity then, some of the material presented is circuit analysis not directly related to transistors. In fact, very little of the book is applicable to transistors only. Although the authors do not emphasize the general applicability of the development, they do emphasize the close similarity between transistors and vacuum tubes. This approach is good pedagogically and will be attractive to practicing engineers desiring to up-date their vacuum-tube amplifier knowledge. Self study is also facilitated by the extensive use of tabular information, worked-out examples, and a goodly supply of problems. The book could also be used as an undergraduate text, although for optimum use the curriculum should be developed so as to allocate a term to active circuit analysis.

The technical content of this book follows the conventional progression of two-port analysis, cascade connections, bandwidth considerations, feedback, stability, and oscillators. Two aspects of the authors' development depart from conventionality. First, the authors develop operating equations using general parameters—signals (current or voltage), immittances (impedance or admittance), etc. Second, the authors carry available power gain formulation over to voltage gain and current gain. Thus, transfer functions are defined as the ratio of output signal divided by the signal available from the source. On this basis the transcurrent quantity in *h* parameters is

$$A_I = h_{21} Y_L Z_s / [(h_{11} + Z_s)(h_{22} + Y_L) - h_{12} h_{21}]$$

as compared with the usual definition obtained by letting

$$Z_s \to \infty \text{ of } A_I = h_{21} Y_L / [h_{22} + Y_L].$$

A 12-page list of symbols is tabulated. Extensive use is made of straight line approximations to actual operating characteristics.

The authors are associated with the Applied Physics Laboratory of Johns Hopkins University where the material was developed and used for special course use.

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Energy Conversion for Space Power, Vol. 3 in Progress in Astronautics and Rocketry, Nathan W. Snyder, Ed.

Published (1961) by Academic Press, Inc., 111 Fifth Ave., New York 3, N. Y. 779 pages+xvi pages. Illus. 6⅛×9¼. $7.25.

This book is a collection of energy conversion papers presented at the American Rocket Society Symposium on space power systems held at Santa Monica, Calif., September 27–30, 1960. Volume 3 covers scientific and engineering principles. Volume 4 of the same series contains system papers presented at this meeting. In addition to the usual advanced energy conversion subjects, such as thermoelectricity, thermionics, solar cells, fuel cells, and magnetohydrodynamics, papers on dynamic as well as electrostatic generators are included. In each category the basic physical principles are adequately reviewed. Most of the papers report engineering studies with space power applications as the final objective. In each area the relationship of the important parameters and problems to system advantages is discussed.

In thermoelectrics, high-temperature generator materials are emphasized, as well as design of thermoelement modules considering heat transfer and thermal stress. Thermoelectric cooling in space applications is also discussed in one paper. In the thermionic section, the physics of thermionic diodes and parameter optimization, principally for the cesium diode, are emphasized. Some new experimental data is given. One paper describes actual operation of a nuclear fuel element-diode combination. The papers on solar cells cover advances in fabrication to achieve better performance and lower costs as well as radiation damage experiments. In fuel cells, the regenerative fuel cell is emphasized. Many possible cells are presented. There are also two papers on secondary batteries for energy storage in space applications. Dynamic engines discussed are the vapor turbine, Stirling engine, expansion engines, as well as an electromechanical energy storage system. One paper describes experimental work on zero gravity boiling. Experimental data are described where combustion gases are used as the working medium in magnetohydrodynamic generators. In addition, two papers cover cylindrical geometry with vortex flow of the working medium. The final section of the book covers parametric electrostatic generators.

The book is recommended for engineers concerned with space power systems or application of new energy conversion techniques in space. This emphasis distinguishes the volume from others in the advanced energy conversion field. Inclusion of papers on dynamic and electrostatic engines is particularly worthwhile. Some of the papers cover original experimental work of broader interest than just space applications.

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Transistors and Active Circuits, by John G. Linvill and James F. Gibbons

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. 501 pages+13 index pages+x pages. Illus. 6¼×9¼. $14.50.

This long awaited book by two members of the faculty of the Stanford Electrical Engineering Department has justified the expectations. The book is an excellent compromise between the existing texts on semiconductor device theory and, on the other hand, texts that are only concerned with circuit results. The authors have attempted with a minimum strain on the average beginning graduate electrical engineer's background to build a firm base for the understanding of most semiconductor active devices. Although the treatment given here relates principally to the bi-polar transistor, their method of solving the resulting differential equations by lumped approximation (difference equations) allows one readily to extend this to other forms of devices.

The book is divided into three major areas. Part 1, The Physics of Semiconductors, contains an introduction to the necessary theory to be used in the rest of the section and a complete discussion of the method of lumped representation that is used through the rest of the book.

Part 2, Two-Port Network Theory, covers the area of small signal theory and develops the results necessary for designing amplifier and oscillator circuits. It may be of some interest to note that Part 2 by itself could stand alone as a short textbook on the theory and design of two-port networks.

Part 3, Transistor Circuits, is the joining of the first two parts into a discussion of applications of transistors in a wide variety of circuit configurations. An interesting unifying approach to the solutions of problems relating to thermal drift, noise and non-linear distortion is presented here under the title of "Contaminating Signals in Amplifiers."

The book is very well written with the serious student in mind in that it avoids the "cook-book" approach and assumes that the reader has mastered all the material presented up to any point. The problems that are at the end of each chapter contribute to the fine presentation of the subject material.

The obvious boundary conditions of size and price of college textbooks probably caused the authors to give a somewhat inadequate coverage (considering the engineering importance) to device characterization and application in the switching mode. It is hoped that this could be remedied in an encore to "Transistors in Active Circuits."

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Introduction to Transients, by D. K. McCleery

Published (1961) by John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. 220 pages+4 index pages+xi pages+7 appendix pages+examples by chapter. Illus. 5¾×8¾. $7.50.

This volume adds substantially to the physical understanding of transients through the medium of Heaviside's operational techniques. The treatment emerges with special clarity without resorting to differential equations or Laplace transforms. The author is particularly sensitive to the usual difficulties that engineering students encounter in attempting to gain a fundamental comprehension of transient phenomena. He is critical of the present tendency to envelope such students with obscure mathematical formalism. "The hungry sheep look up and are not fed."

The textbook is organized in the form of eight chapters. Each chapter is well documented and includes a list of challenging exercises. Chapter one introduces the reader to the exponential law and its many implications. The intuitive approach is illustrated as well as analytical formulations of a variety of waveforms. In Chapter two, Heaviside's concepts are derived and applied to several examples involving the unit step function and the unit ramp function. The procedure is extended to include other types of driving functions. Oscillatory conditions are treated in Chapter three.

In Chapter four, transients in thermionic tube circuits are considered. Emphasis is placed on linear approximations as applied to class-A amplifiers, oscillators, and relaxation circuits. Chapter five deals with more complicated transient problems by utilizing Heaviside's expansion theorem. Traveling waves on transmission lines and shift operators are extensively covered in Chapter six.

Chapter seven concerns itself with the response of circuits to alternating stimuli. Carson's modified expansion theorem is also described and illustrated. In Chapter eight, the superposition principle is stressed and alternative forms of Duhamel's integral are elaborated upon. Carson's infinite integral theorem is also discussed and related to the Laplace transform.

This excellent, practical textbook is an invaluable introduction to the field of transients not only for college students but also for a wide professional readership. It provides essential background for the more abstract approaches which utilize Laplace transform concepts in order to gain deeper understanding.

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Electronic Engineering Principles, 3rd ed., by John D. Ryder

Published (1961) by Prentice-Hall, Inc., Englewood Cliffs, N. J. 409 pages+8 index pages+xi pages+11 appendix pages. Illus. 6¼×9¼. $12.65.

The fact that a work appears in third edition is in itself significant evidence of its merit. In this reviewer's opinion, J. D. Ryder's present revision of two earlier versions of a work well suited to teaching introductory electronics, is obviously a contribution.

After developing a workable understanding on the part of the reader of the behavior of charged particles in fields, the cathode-ray tube, and the physics of thermionic, photo, secondary, and high-field emission, the author introduces tube, semiconductor diode, and transistor fundamentals. From that point he skillfully carries the reader through small-signal amplifiers, feedback, and large-signal amplifiers in the context of both vacuum tubes and transistors.

Active circuit computations are handled conventionally in terms of the quasi-linear equivalent voltage and current generators without the use of transforms. Both z and h-parameter characterizations of transistors are treated.

The contents are well rounded out by adequate discussion of such topics as gaseous conduction, power rectifiers, smoothing circuits, and control and photoelectric devices. The omission of the cascode amplifier, the lack of emphasis on the class C amplifier, and the deceptively simple treatment of the tunnel diode are regarded by this reviewer as the least satisfactory aspects of a good work.

Overall, the book has much to commend it. The treatment of selected classic experiments adds much interest to the sections concerned with physical background material. Further, the documentation throughout is excellent, including references to classic as well as to current papers. The more than two hundred fifty problems included should afford ample illustration and practice. For convenience in reference, tabulation of physical constants and numerous tube and transistor graphical characteristics are also included. Ryder's work should prove to be a highly satisfactory vehicle for a first excursion into the area of electronic engineering.

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Recent Books

Brown, K. and Weiser, P., Eds. *Ground Support Systems for Missiles and Space Vehicles.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $15.00. Treatment directed toward the average engineer having no prior acquaintance with ground support systems. The text approaches the system as a whole unit; all subsystems are considered.

Carson, R. S. *Principles of Applied Electronics.* McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. $9.95. Provides a fully integrated introduction to the principles which govern the operation of the latest vacuum, gas, and semiconductor electron devices.

Kipiniak, W. *Dynamic Optimization and Control.* The M.I.T. Press, Mass. Inst. of Tech., Cambridge 39, Mass., $4.95. Develops a general control theory for dynamic systems of immediate interest to electrical, mechanical, and chemical engineers and to operations research analysts.

Koelle, H. H., Ed.-in-Chief. *Handbook of Astronautical Engineering.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $27.50. A handbook covering the major areas of space flight technology, organized in six parts; fundamentals of astronautical engineering, astrodynamics, astrionics, propulsion systems, space vehicles, and space flight operations.

Lanczos, Cornelius. *Linear Differential Operators.* D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. $12.75. A book designed for engineers, mathematical physicists and other applied mathematicians.

Ramey, R. L. *Physical Electronics.* Wadsworth Publishing Co., Inc., Belmont, Calif. $8.50.

Seifert, H. S. and Brown, K. *Ballistic Missile and Space Vehicle Systems.* John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. $12.00. Covers the disciplines involved in the development of a ballistic missile or space vehicle system from an engineering viewpoint; the purpose is to stress the relationships between the different parts of such a system, with particular emphasis on interface and systems design principles.

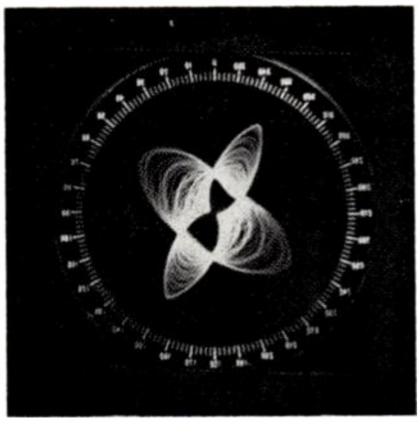
Snyder, N. W., Ed. *Space Power Systems,* Vol. 4 in *Progress in Astronautics and Rocketry,* M. Summerfield, Series Ed. Academic Press, Inc., 111 Fifth Ave., New York 3, N. Y. $6.00.

Vajda, S. *Mathematical Programming.* Addison-Wesley Publishing Co., Inc., Reading, Mass. $8.50. The book offers an introduction to linear and nonlinear programming, with emphasis on the mathematical aspects of the subject. It is designed for use as a textbook at the graduate level.

Yeager, Ernest, Ed. *Transactions of the Symposium on Electrode Processes.* John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. $20.00. The book includes all papers and discussions presented at the Symposium sponsored by The Electrochemical Society, Inc., and the U. S. Air Force Office of Scientific Research, held in Philadelphia, Pa., May, 1959.

Scanning the Transactions

The transmission of power without wires has been an unfulfilled dream of scientists and engineers for more than a half century. Interest in wireless power transmission dates back to 1899, when Nikola Tesla, with the aid of a grant from John Jacob Astor, constructed a large experimental station at Colorado Springs where he attempted to transmit power over long distances at a frequency of a few hundred kilocycles. His attempts failed. No provision had been made for directing the radiated energy. Indeed, at this low frequency a directional array would have been impossibly large. Today, we are readily able to direct electromagnetic energy into narrow beams by utilizing the microwave region of the spectrum, but the generation of large amounts of power at these high frequencies poses a formidable problem. An important step forward was made in the late 1950's with the discovery that large amounts of power can be obtained from relatively small sized microwave tube structures through the use of high-efficiency designs and new cooling systems, as exemplified by the Amplitron. Average power outputs of more than 10 kilowatts can now be attained at 10 centimeter wavelengths, and 200 kilowatt outputs are now confidently predicted in the foreseeable future. As a result, the feasibility of radiowave transmission of power is once again under serious study. Among a number of potential applications being investigated is one in which a helicopter device is sustained for an indefinitely long period of time at a point above the earth's surface by energy furnished to it by a microwave beam. A typical system envisioned for this application would involve a 350-foot transmitting antenna on the ground operating at 6 kMc and a 50-foot receiving antenna on a helicopter device hovering 65,000 feet above. Recent studies have indicated that such a vehicle could hover with the desired efficiency at an altitude as high as 102,000 feet. A flying platform 10 or more miles high might provide a number of technical and economic advantages for airborne early warning systems, air route surveillance radars, educational television, etc. In any event, it appears that a 60-year dream is about to come true. **(W. C. Brown "Power without wires," IRE STUDENT QUARTERLY, September, 1962; and H. Letaw, Jr., et al., "Some comments on the transmission of power by the use of microwave beams," IRE TRANS. ON ANTENNAS AND PROPAGATION, November, 1961.)**



Radio direction finding, although a well-established technique, is still undergoing improvement and refinement. One of the problems that has attracted investigators in recent years is that of resolving the directional components of a multicomponent field. It was found that the directions of two components of equal frequency can be resolved by using crossed loops and a twin-channel receiver. Recent work has now shown that similar results can be obtained by spinning a single-loop antenna at an appropriate rate and using a single-channel receiver. The unusual pattern in the accompanying photograph is the direction finder display which results when there are two components of equal frequency at bearings of 0° and 60°. In this type of CRO display, the loop antenna pattern is inverted so that nulls appear at the periphery and maxima at the center. The bearing of the two components is indicated by the corners of the black "bow tie" in the center. **(W. M. Sherrill and D. N. Travers, "Single-channel direction finding in a multicomponent field," IRE TRANS. ON ANTENNAS AND PROPAGATION, November, 1961.)**

Nuclear instrumentation, once a narrow subdiscipline among the byways of engineering and physics, has recently advanced to a more prominent and centralized position in the mainstream of science and technology, becoming an essential part of several important fields of activity. In fact, today, this subdiscipline can itself be subdivided into major areas of its own, such as reactor control, plasma diagnostics, radiochemical processing, experimental physics and radiation measurement. The rapid growth in the scope and importance of this subject led the IRE, AIEE and the Instrument Society of America to hold a Joint Nuclear Instrumentation Symposium at North Carolina State College last September. This was the first of what is likely to become an annual or biennial event. The papers presented at the meeting gave ample evidence of the breadth and variety of subject matter with which the nuclear instrumentation specialist is now concerned. The topics included such diverse areas as gamma-ray astronomy, digital computer control of nuclear power stations, gamma monitoring in radiochemical processing plants, and instrumentation for plasma physics research. The discussions were noteworthy not only for their diversity but for their depth. In the area of plasma physics research, for example, one finds descriptions of (1) instrumentation for measuring the electrical characteristics of a plasma during discharge experiments; (2) magnetic and electrostatic plasma probes; (3) spectroscopic techniques in the microwave, infrared, visible and ultraviolet regions; (4) measurements of high-energy radiation of neutrons, electrons and X rays; and (5) analog computer studies of plasma instrumentation problems. The 25 papers presented at this first joint symposium have now been published in the October issue of the IRE TRANSACTIONS ON NUCLEAR SCIENCE, providing a valuable report on current developments in a blossoming field.

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	IRE Members	Libraries and Colleges	Non Members
Antennas and Propagation	AP-9, No. 6	$2.25	$3.25	$4.50
Electron Devices	ED-8, No. 6	2.25	3.25	4.50
Information Theory	IT-7, No. 4	2.25	3.25	4.50
Reliability and Quality Control	RQC-10, No. 3	2.25	3.25	4.50
Nuclear Science	NS-8, No. 4	2.25	3.25	4.50
Space Electronics and Telemetry	SET-7, No. 4	2.25	3.25	4.50

Antennas and Propagation

Vol. AP-9, No. 6, November, 1961

Scattering by a Periodically Apertured Conducting Screen—R. B. Kieburtz and A. Ishimaru (p. 506)

This paper presents a generalized variational method for calculation of equivalent-circuit parameters or scattering coefficients of any thin obstacle. The equivalent susceptance is obtained from an eigenvalue of a general-energy operator. The method is applied here to the scattering of a plane wave by a conducting screen containing a doubly-periodic array of square apertures.

Previous variational formulations of analogous waveguide problems were largely restricted to cases in which only the dominant mode can propagate. In applying these methods to scattering problems, the ratio of aperture spacing to wavelength had to be restricted to less than unity at normal incidence so that no additional propagating modes were excited. The formulation presented here can be applied, however, for any number of propagating modes.

Numerical results are presented for scattering by a periodically apertured screen, giving scattering coefficients as a function of wave number of the incident wave. The distribution of energy into higher-order diffraction lobes is evaluated.

A Reaction Theorem and Its Application to Antenna Impedance Calculations—J. H. Richmond (p. 515)

The reaction theorem is generalized to allow the fields of an antenna in one environment to be employed in calculations of mutual impedance in another environment.

Several expressions for self-impedance and mutual impedance are presented. These are in the form of surface integrals or volume integrals of the field intensities or the current density. It is shown how the fields of an antenna in free space can be useful in calculating the impedance in the presence of scatterers.

Single-Channel Direction Finding in a Multicomponent Field—W. M. Sherrill and D. N. Travers (p. 521)

On the basis of theoretical argument and experimental results, the capability of a single-channel spinning-loop direction finder to resolve the directional components of a multicomponent field of plane waves is demonstrated. Calculated bearing displays are compared with corresponding displays obtained by experiment for the cases of two and three components. Directional resolution of components of equal and differing frequencies is demonstrated, and methods of presenting the bearing display are discussed.

A Solution to the Frequency-Independent Antenna Problem—B. Ru-Shao Cheo, V. H. Rumsey and W. J. Welch (p. 527)

A solution of Maxwell's equations is obtained for an antenna consisting of an infinite number of equally spaced wires in the form of coplanar equiangular spirals. Radiation amplitude patterns obtained from this solution agree closely with measurements on two-element spiral antennas. The phase pattern shows the approximate validity of a phase center at a distance behind the antenna which decreases with the tightness of the spiral. The current distribution clearly shows increased attenuation with increase in the tightness of the spiral, thus showing how the frequency-independent mode depends on the curvature. A remarkable feature of the solution is that the current consists of an inward traveling wave at infinity when the antenna is excited in that sense which produces an outward wave at the center.

The Radiation Fields from a Circumferential Slot on a Metal Cylinder Coated with a Lossy Dielectric—Charles M. Knop (p. 535)

The radiation fields produced by a sinusoidal distribution of axial-electric field along a thin circumferential slot, cut in a perfectly conducting infinite cylinder which is covered by a concentric dielectric coating, are found by applying the method of Wait with two modifications. Initially, a structure consisting of a finite coated cylinder containing the slot and exciting a radial waveguide is considered. The fields in this waveguide are expressed in terms of derivatives of two axial Hertz vectors, and for a finite radial wall spacing are seen to consist of a double Fourier series. The radial walls are then allowed to become infinitely spaced; in this process, the Fourier-series representation for the axial dependence of the fields becomes a Fourier integral. The radiation fields are then found by asymptotically evaluating the Fourier integral by the method of stationary phase rather than by the saddle-point method.

Expressions for the radiation fields are then found, but are given explicitly for the equatorial plane only. Calculated radiation patterns in this plane, for a coating having a fixed dielectric constant with the thickness as a parameter, and for a fixed thickness with the dielectric constant as a parameter, are given for the case of a cylinder of size $\beta_0 a = 3$. Generalizations based on these calculations are suggested. The case of a specific plasma coating is briefly considered, and an approximate solution, readily obtained, which gives the same azimuthal form for the equatorial patterns, is also noted.

Radiation from a Radial Electric Dipole Near a Long Finite Circular Cylinder—Hans H. Kuehl (p. 546)

By use of the currents on the infinite cylinder excited by a radial dipole, approximate expressions for the two components of the far-zone electric field of a radial electric dipole near a perfectly conducting finite cylinder are derived. The validity of the approximation depends on the conditions, $kl_2 \gg 1$ and $kl_1 \gg 1$, *i.e.*, the cylinder must be long compared to a wavelength. The expressions are initially in integral form but it is shown that they may be evaluated approximately over most of the range of θ by use of the saddle point method. The theoretical results are compared to experimental results for a particular cylinder.

ELF Waves in the Presence of Exponential Ionospheric Conductivity Profiles—Janis Galejs (p. 554)

Based on the theory of Nicolet and Aikin, the profile of ionospheric conductivity exhibits a nearly exponential variation with altitude through the *D* layer and the lower edge of the *E* layer. Propagation of ELF waves below this exponential layer is considered after calculating the surface impedance Z_s at an altitude h where the local refractive index n does not necessarily satisfy $|n| \gg 1$. The propagation constant is determined by an iteration process. This model of an isotropic nonhomogeneous ionosphere is a closer approximation of the propagation geometry than are earlier models, where Z_s was defined at h where $|n| \gg 1$. The present model accounts simultaneously for ELF attenuation rates as measured by Jean, and earth-ionosphere cavity resonances as observed by Balser and Wagner.

Some Remarks on Green's Dyadic for Infinite Space—J. Van Bladel (p. 563)

The validity of the often used dyadic

$$-\left(\mathfrak{J} + \frac{1}{k^2}\nabla\nabla\right)\frac{e^{-jkR}}{4\pi R}$$

to compute the electric field *inside* a current-carrying region is investigated. It is found that care must be exercised in the definition of the integrals, which should be taken as principal values around the field point.

Apparent Temperatures of Smooth and Rough Terrain—S. N. C. Chen and W. H. Peake (p. 567)

The apparent temperatures of smooth and rough terrain surfaces are calculated for frequencies between 1 and 75 kMc, for angles of incidence between 5° and 80°, and for observation altitudes between 2 km and 32 km. The attenuation and thermal radiation caused by the atmosphere are accounted for by an experimental model atmosphere, and the surface emissivities are based on measured complex dielectric constants (for the smooth surfaces) and measured radar return (for the rough surfaces). It is found that perpendicular polarization provides the greatest temperature contrast between rough and smooth surfaces, and that the contrast decreases as the altitude of observation increases; at $f = 75$ kMc (approximately the resonant frequency of the oxygen molecule), there is likely to be little, if any, contrast between rough and smooth surfaces.

Diffraction of a Plane Wave by a Perfectly Conducting Sphere with a Concentric Shell—Martin A. Plonus (p. 573)

A general solution to the problem of diffraction from a perfectly conducting sphere with a concentric shell spaced any distance from the surface of the sphere is presented. If the shell consists of a dielectric material the solution simplifies by using asymptotic expressions for the spherical Bessel function. Another simplification results when the shell is assumed to be thin.

Correction to "Sidelobe Reduction by Nonuniform Element Spacing"—Roger F. Harrington (p. 576)

Communications (p. 577)

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Papers to be Published in Future Issues (Inside Back Cover)

Electron Devices

Vol. ED-8, No. 6, November, 1961

Our Sincere Thanks to Dr. Earl Steele—W. A. Adock (p. 435)

Mylar Electron Beam Windows—William H. Steier (p. 436)

Thin sheets of Myler have been found a useful material for electron beam windows for beams in the 0.5- to 1.0-Mev energy range. The material finds special application in beam excited plasma devices where the plasma and the electron gun are to be held at different pressures.

Experiments are described in which the amount of beam scatter and beam debunching caused by the Mylar are determined. Data is presented on the damage to the Mylar caused by the electron beam and the resultant decrease in pressure differential that the bombarded Mylar can withstand. The experiments were also conducted using Mylar sheets coated with 4000–5000 A coatings of aluminum.

The Magnetron as a Negative-Resistance Amplifier—J. Kline (p. 437)

A one-port, nonlinear, crossed-field microwave amplifier which uses a magnetron as a negative-resistance element is described. Such an amplifier is characterized as one using an emitting sole, reentrant beam, and reentrant RF fields. A circulator is required to achieve stable regenerative amplification. The system is analyzed as a negative-resistance amplifier and as a synchronized or forced oscillator. Experimental measurements of gain and bandwidth are illustrated. The effect of varying operating conditions upon these characteristics is indicated. Operation with a solid cathode composed of oxidized beryllium-nickel has been demonstrated.

Super-Saturated Transistor Switches—James F. Gibbons (p. 443)

The purpose of this paper is to calculate the voltage drop across a closed transistor switch in the presence of intermediate-level operating conditions. The problem is formulated to include the important one-dimensional effects associated with intermediate-level operation in such a way that the low-level theory is included as a special case. Experimental verification of the theory has been obtained over a moderately wide range of base drives.

Conservation Principles in Multivelocity Electron Flow—G. Hok (p. 452)

This paper considers the extension of Poynting's theorem to an electron gas with a continuous distribution of velocities. In particular, an extension of the concept of "kinetic potential" is attempted, since this concept has proved itself very useful in the investigation of single-velocity flow. It is found that in three dimensions the electrokinetic-flow vector cannot be expressed as the product of the convection-current density and a single scalar quantity of the dimension potential. In one-dimensional applications, however, this circumstance is immaterial. Another difficulty is encountered when a small perturbation component on a steady state is considered. The nonlinear Bolzmann transport equation gives a linear equation between the first-order perturbations, but the nonlinearity makes possible a conversion of part of the perturbation power flow to dc power flow. In other words, the power flow associated with a single-frequency perturbation is not necessarily conserved in multivelocity flow, even in the absence of an ac Poynting vector. In the case of space-charge-limited flow in an electron gun, the consequence is that the cooling of an electron stream by adiabatic expansion can proceed beyond the potential minimum and very low noise temperatures are attainable in principle if the accelerating field is maintained small over an appreciable length of the electron gun.

Uniform Turn-On in Four Layer Diodes—K. Hubner, M. Melehy and R. L. Biesele (p. 461)

The turn-on characteristics of 200-volt four-layer diodes have been investigated as a function of the rate of rise of the applied voltage close to the breakover point. A fast rate of rise of the order of kv/μsec allows bringing the voltage up to, or higher than, the designed avalanche voltage of the center junction, even if localized spots having a lower breakdown voltage are present. Such spots, possibly due to crystalline defects, surface conditions and statistical or accidental variations of impurity concentration, may lead to localized turn-on and possible burn out. There is experimental evidence that a fast rate of rise results in uniform turn-on and allows current densities up to 55,000 a/cm² without damaging the device.

A Millimicrosecond Pulse-Generator Tube—W. H. Cornetet, Jr. and J. G. Josenhans (p. 464)

This paper describes the theory and testing of a 100-Mc repetition-rate millimicrosecond pulse-generator tube utilizing klystron bunching. The optimum operation of the generator was obtained with a drift-tube transit angle Θ_0 of approximately 150°; depth of modulation $\alpha = V_1/V_0 \doteq 1$; and the reduced plasma-angle ϕ_q was less than 20°. Observations and data were determined with an Edgerton, Germeshausen and Grier traveling-wave oscilloscope and a comparison measurement technique. The generator has a measured pulse rise time of 130 $\mu\mu$sec (0–100 per cent), and a pulse duration of 300 $\mu\mu$sec at the half-voltage point. A pulse height of 20 volts on a 50-ohm coaxial line is possible.

Definition and Determination of the Series Inductance of Tunnel Diodes—W. B. Hauer (p. 470)

For the application of tunnel diodes in the microwave region it is important that the series inductance of the diodes is well below one nanohenry. Such small inductances are very difficult to measure. A method is described here by which the inductance can be calculated. The key point of the procedure is a new parameter called the effective radius of a tunnel diode. It is the radius of a cylindrical conductor which can be substituted for the diode in inductance calculations. Experimental determination of the effective radius is shown. The approach taken readily leads to a concise definition of the series inductance of tunnel diodes.

Impurity-Density Distribution in the Base Region on Drift Transistors—M. B. Das and A. R. Boothroyd (p. 475)

The nature of the impurity grading in the base of drift transistors is studied by measuring the depedance of the base transit time on collector voltage. Provided that modulation of collector-depletion layer width caused by change of collector voltage occurs only in the base material, as in the case of an alloyed collector, it is possible to deduce the base-region field parameter $m = \Delta V/(kT/q)$ from such transit-time measurements. By this means, the validity of assumed distributions of impurity density may be verified; in particular, it may be established whether the impurity grading approximates an exponential or a complementary error function (erfc) form. Results are given for a number of drift-transistor samples, most of which are believed to have undergone impurity diffusion into the base material from a constant surface concentration during fabrication. In all cases, however, interpretation of measured data indicates a base impurity-density distribution approximating exponential rather than erfc form to be present.

Closely Spaced, Aligned Grids in Vacuum Tubes—L. A. Harris (p. 481)

Applications of Cold-Cathode Parallel-Field Devices—R. E. Lundgren, C. Susskind and J. R. Woodyard (p. 489)

Investigations have been undertaken at the University of California over the past several years to extend the application of a cold-cathode parallel-field device to the fields of rectification, amplification, oscillation, and electron emission. The characteristics of the device as functions of voltage and magnetic field in the pressure range near 10^{-3} mm of Hg have been studied. Operation of the device as a short-circuit-stable variety of negative resistance is possible, and a simple theory is given to explain the negative-resistance phenomenon. Results indicate that rectifiers capable of withstanding inverse voltages of at least 30 kv and currents of 10 ma with a forward drop of 300 v are possible. A peak electron emission of 7.2 ma/w indicates that the device might serve as an electron source in long-pulse applications.

Cerenkov Radiation in Slow-Wave Guiding Structures—B. W. Hakki (p. 497)

Cerenkov radiation of a prebunched beam in a dielectric tube waveguide is here shown to be about an order of magnitude higher than that in a comparable slab of dielectric material. Furthermore, two distinct types of interaction are analyzed: multibunch, and bunch-by-bunch radiation, wherein integral transform methods are used. Multibunch Cerenkov radiation in a helix waveguide is used to show the identity with TWT interaction. It is subsequently shown that for open boundary inhomogeneous waveguides a TWT type analysis will give the same results as multibunch Cerenkov radiation based on integral transform methods. For single bunch or a bunch-by-bunch type of interaction in such structures where energy orthogonality and completeness of the eigenfunctions do not hold, an integral transform method is used. The underlying difference between bunch-by-bunch and multibunch radiation is discussed.

The utility of Cerenkov radiation in a dielectric tube waveguide for the generation of coherent electromagnetic power in the millimeter range is pointed out. Numerical results are given to show the variation of power with respect to various parameters. Experimental results from a 0.8-Mev beam are shown to agree satisfactorily with theory.

Single-Transit, Large Radius E-Type Devices—W. M. Nunn, Jr. and J. E. Rowe (p. 508)

A small-signal analysis of E-type traveling-wave devices is presented for systems in which a ribbon-shaped electron beam completes a single excursion through the interaction space of a coaxial-cylindrical structure. The incremental propagation constants and the associated forward-wave growing- and beating-wave gain, as well as the backward-wave start-oscillation conditions, are examined over a wide range of operating conditions. The results point to an interaction-principle that specifies the manner in which the rate of change of energy transfer between the electron beam and the RF wave is affected by the curvature of the path.

The origins of slipping-stream motion are examined and an approximate method of computing gain under these circumstances is presented. In addition, predictions based on a small-signal study reveal that E- and O-type devices possess comparable efficiencies.

Experimental data on a C-band E-type amplifier are presented in support of the theoretical predictions.

Parametric Amplifier with Thermoelectric Refrigeration—M. Uenohara and R. Wolfe (p. 521)

The noise figure of a variable-capacitance parametric amplifier can be greatly improved by refrigerating the diode. A thermoelectric refrigerator can be used for this purpose without losing the advantage of system simplicity. A two-stage thermoelectric refrigerator has been built into a 6-Gc parametric amplifier. With no load this refrigerator has produced a temperature difference of 101°C below room temperature. In the amplifier it cooled the diode to 213°K. The effective noise temperature of the amplifier was reduced from 170°K when the GaAs diode was at room temperature to 108°K when the diode was cooled to 213°K. The design, construction and characteristics of the amplifier and the thermoelectric refrigerator are described.

Equivalence of the Collective and the Corpuscular Theories of Noise in Junction Diodes—A. van der Ziel (p. 525)

There are at present two methods of approach to the problem of noise in *p-n* junction diodes: 1) the corpuscular theory and 2) the more fundamental collective theory. In the first the noise is expressed in terms of the fate of individual carriers, and in the second the noise is described in terms of distributed diffusion noise and recombination noise sources. The equivalence of the two methods is demonstrated by integrating the continuity equation and identifying the various groups of carriers showing up in the corpuscular theory.

Electron-Beam Defocusing Due to High-Intensity RF Fields—G. R. Brewer and J. R. Anderson (p. 528)

The electric fields associated with a "slow" electromagnetic wave propagating along an electron beam will modulate this beam; the axial component of field will produce the usual "bunching," while the radial component will cause periodic perturbations in the radius of such an electron beam. This paper presents the results of analytical and experimental studies in an attempt to explain certain features of this defocusing effect of intense RF fields on the beam in a traveling-wave tube. In particular the effect of cathode flux on the RF defocusing of a beam in Brillouin flow is treated.

Constant-Output-Frequency, Octave Tuning Range Backward-Wave Parametric Amplifier—S. Okwit and E. W. Sard (p. 540)

This paper describes a varactor diode backward-wave parametric amplifier (BWPA) which operates in a new mode that yields a constant-output frequency as the amplification frequency is electronically turned over an octave signal tuning range. A theoretical discussion on the design considerations is presented and applied to the construction of an LF model (1.5–3.3 Mc). Experimental data on the important amplifier characteristics are given and shown to correlate well with theory. The amplifier yields stable gains in excess of 20 db over a greater-than-octave tuning range, and the output frequency (which is taken at the idler) has less than a ±1.6 per cent variation over the octave band. The over-all effective receiver input-noise temperatures were measured and agreed well with theory, being about 160°K at the LF end of the band and increasing to 300°K at the HF end.

Propagation in Helical and Narrow-Band Slow Wave Structures—B. M. Bevensee (p. 549)

A variational analysis is developed for a large class of helical slow-wave structures and illustrated for a narrow tape helix. A 2×2 determinantal equation accurately describes the dispersion of the wave with predominant $n=1$ mode component, both in the ordinary and anomalous regions. A 3×3 determinantal equation would be necessary to describe well the dispersion of the ordinary TWT wave at low frequencies.

The analysis is modified to include the presence of a filamentary electron beam. It is shown how the equation for the incremental propagation constant takes the Pierce form, provided the frequency does not lie too near cutoff.

A variational analysis is also made of the propagation in a general narrow-band cavity chain interacting with a longitudinally-confined beam. It is shown how the equation for the incremental phase shift per cavity is formally related to Pierce's equation for the incremental propagation constant in a helix TWT. Provided certain necessary (but not sufficient) conditions are met, an appropriately-defined C^3 parameter exists for the cavity chain.

The resonant-slot coupled cavity chain is treated variationally and it is shown how a very accurate dispersion relation obtains. The waves in this structure interacting with a beam are also described formally, under certain conditions, by the Pierce equation for the TWT.

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Information Theory

Vol. IT-7, No. 4. October, 1961

Optimum Linear Finite-Dimensional Estimator of Signal Waveforms—Tadao Kasami (p. 206)

This paper deals with the linear estimation of noise corrupted signal waveforms, from observation over a specified finite time-interval. Two new features are delineated in this paper: 1) the estimating operator is assumed to be finite-dimensional, because economically realizable operators are finite-dimensional in general, and 2) the cost of observation and estimation is taken into account, and assumed to be dependent upon the dimension of the operator only. As a result, the optimum linear finite-dimensional operator that minimizes the risk (the sum of cost and average loss) is obtained for the case in which a quadratic loss function due to error is adopted.

A Method of Digital Signalling in the Presence of Additive Gaussian Noise—Ludwik Kurz (p. 215)

This paper considers the basic problem of transmitting digital information through a noisy channel with minimum probability of error in finite time. The transmitted signals are average-power limited, and the noise is assumed to be additive Gaussian with a power spectrum which may be nonwhite. A theory of so-called efficient codes (minimax, equal separation, and nearly equal separation) is developed. Efficient codes are formed from weighted sums of eigenfunctions generated by an integral equation with its kernel corresponding to the inverse Fourier transform of the Gaussian noise power spectrum. In addition, the theory of equidistant and nearly equidistant codes is extended to the case of nonwhite Gaussian noise. It is shown that efficient codes perform better than equidistant codes if the noise is nonwhite; *i.e.*, properly chosen waveforms are more efficient than binary coding. Performance results are given for several different codes when the interference is white Gaussian noise and when the noise power density increases with increasing frequency. The detection scheme used does not require estimation of the signal or the noise levels at the receiver and is thus independent of fading.

Some Results on the Problem of Discriminating Between Two Gaussian Processes—P. Bello (p. 224)

This paper is concerned with certain aspects of the problem of discriminating between two Gaussian processes. The emphasis is on determining approximate optimum detector structures which avoid some of the mathematical difficulties inherent in the evaluation of the exact optimum detector structure. To this end, an approach termed the "inverse operator" approach is presented which leads to approximate detector structures via the Neumann series expansion of linear operator theory. These approximate detectors are found by using a finite number of terms in an "optimum detector" expansion which results from the use of the above Neumann series expansion. A sufficient condition for the rapid convergence of the optimum detector expansion is found to be that the eigenvalues of a certain operator have magnitudes much less than unity. An upper bound is derived for the error incurred at the detector output by the use of a finite number of terms in the optimum detector expansion. Error probabilities are calculated for the case in which the detector outputs may be assumed approximately normally distributed. From an output "signal-to-noise" ratio point of view, it is shown that the performance of the optimum detector and approximate detectors will differ negligibly if the above eigenvalues have squared magnitudes much less than unity. Some upper bounds are derived for the largest eigenvalue (magnitude).

Error Correcting Codes and Their Implementation for Data Transmission Systems—J. E. Meggitt (p. 234)

Presented here is a practical automatic error-correcting system that may be applied to many data transmission problems. It is particularly suited to the correction of bursts of errors and so may be applied to the problem of the transmission of data over telephone networks.

The attractive feature of the system is its remarkable simplicity from the point of view of implementation. It is so simple that it can readily be incorporated into much existing equipment.

In the system, messages are transmitted in blocks and each block is coded separately. The codes used within the blocks are cyclic codes. This means that coders and decoders employ linear feedback shift registers to form check digits and to correct errors.

The basic ideas are presented in terms of the hardware components to which the system gives rise and analyzed afterwards in terms of mathematics so that it is easy for the engineer to see, at once, what is involved.

The theory usually applies to binary messages in which data is transmitted serially. However, an extension is included which shows how the same ideas may be applied to binary codes in which information is sent in parallel.

On the Optimum Range Resolution of Radar Signals in Noise—N. J. Nilsson (p. 245)

Optimum radar resolution is recognized to be a problem in distinguishing between different possible target configurations. Radar reception systems which perform optimum range resolution are then designed using the principles of statistical decision theory. In particular, the design of the optimum resolution system is carried out for a squared-error loss function, modified to provide extra penalties for wrong guesses about the number of targets present. Such a system is capable of simultaneously deciding the number of targets present, their spatial positions (ranges) and their relative amplitudes. The analysis also includes a discussion of an optimum device for the resolution of distributed (clutter-like) targets.

Phase Shift Pulse Codes with Good Periodic Correlation Properties—R. C. Heimiller (p. 254)

A method of generating phase shift pulse codes of arbitrarily long length with zero periodic correlation except for the peak for zero shift is presented. The codes are of length p^2 where p is any prime number, and p different phase shifts corresponding to the pth roots of unity are necessary to generate them. Since p different phase shifts are required, these codes are not as easy to generate and process as the binary codes, but this does not seem to be a serious limitation to their usefulness. Application of these codes can be made as interpulse phase modulation for range resolution in pulse Doppler radars or for a method of synchronizing a pulse code communication system.

Optimum Nonlinear Filters for Quantized Inputs—F. Tung and R. J. Schwarz (p. 257)

Optimum least-square filters belonging Zadeh's nonlinear class $\mathfrak{N}_1$ are considered. Attention is restricted to those systems whose present output is influenced only by a portion of the past input. The input signal consists of a message and noise, both of which are stationary random processes. It is assumed that the amplitude of the input time series is bounded and takes on discrete values at all times. This assumption leads to a nonlinear filter which can be realized as a quantizer or amplitude selector followed by a parallel set of linear filters. The system becomes optimum when the impulse responses of the linear filters satisfy a system of integral equations of the Wiener-Hopf type adapted to finite memory filters. By virtue of the assumptions made concerning the joint probability density functions of the message and noise processes, it is found that the Fourier transforms of the kernels of these equations are rational functions. A method is developed for the solution of this set of integral equations. This method is illustrated by an example, and the mean-square error of the nonlinear filter so obtained is compared with the best linear filter.

Correspondence (p. 265)

Correction to "On the Approach of a Filtered Pulse Train to a Stationary Gaussian Process"—Phillip Bello (p. 277)

Correction to "Maximum-Weight Group Codes for the Balanced M-ary Channel"—Carl W. Helstrom (p. 277)

Correction to "On the Factorization of Rational Matrices"—D. C. Youla (p. 277)

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PGIT Affiliates Plan (p. 289)

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Reliability and Quality Control

Vol. RQC-10, No. 3, November, 1961

Manufacture and Control of High Frequency Transistors for Consumer Products—C. B. Tague and H. C. Bartels (p. 1)

The philosophy of manufacturing transistors on an automatic line optimized for a particular high frequency type is reviewed. The method of optimizing for different types and the process controls used to monitor these procedures are analyzed.

The concept of defining high frequency characteristics through functional tests rather than by traditional parameters and some of the problems encountered with this approach are discussed.

Results of sampling tests for life and electrical characteristics are reviewed. Field data, both consumer production line and in use failure rates, are given.

Some Aspects of Satellite and Space Probe Reliability—T. W. Gross and H. C. Werner (p. 7)

It is becoming a recognized fact that many of the electronic piece parts used in present day ballistic missile and space probes are inadequate for the long-life space programs of the near future. Realization of the major differences of life and reliability requirements resulting from short and long term exposures to the space environment has led to a four point plan for the establishment of an acceptable class of parts that will meet the more rigid demands. Presented here are the plan and some of the preliminary results obtained to date.

Design by Worst-Case Analysis: A Systematic Method to Approach Specified Reliability Requirements—W. D. Ashcraft and W. Hochwald (p. 15)

Design criteria which result in a systematic method to approach specified reliability requirements of equipment are essential to the development of complex electronic systems. A practical approach to designing equipment which lends itself to theoretically valid reliability prediction has long been sought. The worst-case analysis offers an answer in terms of a simple, well-organized, analytical method. Furthermore, application of this method results in a number of factors which benefit the component design engineer, the system design engineer, the administration of the engineering organization which applies these techniques, and the manufacturer of the component parts. The subject of worst-case criteria will be introduced, and the resulting effects on hardware and the engineer will be examined. The procedure for performing the worst-case analysis will then be described, and sample circuits designed in this manner will be shown. Lastly, the relation of worst-case criteria to statistical theory will be described, and the validity of worst-case criteria for reliability prediction will be shown.

Correlation-Capability Studies—Howard L. Berke (p. 22)

In missile production, a recurring problem is the measurement of derived parameters. Unlike volts, amperes, and ohms, these parameters do not lend themselves to standard black-box test equipment. Nevertheless, whatever test equipment is used for their measurement still must be closely correlated to yield consistent and reliable measures.

The Span Plan, which has been successful largely because it uses range estimates rather than actual variances, has shown itself to be the most practical method for conducting correlation studies on Sidewinder missile test stations.

After discussing the Span Plan in some detail, the author concludes that the Correlation Capability Program, in establishing and maintaining close correlation of test stations, has been a significant contributor to Sidewinder's success.

Built-In Reliability—E. P. Laffie (p. 28)

There has been a growing tendency, lately, to substitute the word "reliability" for "quality control." There have been many definitions for reliability proposed. A simple but adequate one has been given by Knight, *et al.*: "The reliability of an electronic product is the probability that the product will give satisfactory performance for a given period of time when used in the manner and for the purpose intended."

One of the basic and fundamental concepts for achievement of true reliability is to build the product right in the first place. The aim of this paper is to point out how this can be accomplished and areas of actual application.

Biographies (p. 36)

Nuclear Science

Vol. NS-8, No. 4, October, 1961

The Nuclear Instrumentation Symposium—C. S. Lisser (p. 2)

Reactor Power Monitor Utilizing Cerenkov Radiation—James L. Lovvorn (p. 3)

Cerenkov light in the central region of a water-moderated reactor has been measured as a function of reactor power. Simple photovoltaic light detectors were used with a recording microammeter. The electrical current responded to a power increment increase with an initial rise proportional to the reactor power. This was followed by an additional increase in current of from 4 to 5% with a time constant of approximately 4 minutes.

High-Current Regulated Supply, Combining Transistors and Magnetic Amplifiers, for the Oak Ridge Isochronous Cyclotron—B. C. Behr, W. E. Lingar, and W. H. White, Jr. (p. 6)

Circuits are shown and design criteria given for a current-regulated power supply which is typical of 19 supplies of four types built for the Oak Ridge Isochronous Cyclotron. These supplies range from 6.25 to 34 kw with current ratings from 250 to 1200 amp. Current regulation and ripple of 1 part in 10^4 are obtained by separate ac and dc feedback circuits driving a bank of parallel power transistors in series with the load. The dissipation of the transistor bank is limited by a third feedback circuit, which drives a saturable reactor in the power supply to maintain a constant voltage across the transistor bank. The output can be varied from zero to maximum by setting a reference potentiometer which is supplied by a Zener reference with a low temperature coefficient. Plug-in circuits in the feedback networks allow versatility; any regulator can be used with a wide variety of magnets of varying time constants. A modular transistor bank simplifies construction and maintenance.

Low Level Alpha Counting with Solid State Detectors—L. Cathey (p. 10)

Two instruments are described that utilize the high discrimination of solid state alpha detectors against beta and gamma radiation. An alpha particle survey meter has been developed with a 6 cm² sensitive area. It will detect alpha particles in the presence of a 50 r/hr gamma ray field. A low level alpha counter has been built with a background of less than 0.001 count per minute. It is useful for measurement of alpha emission from bio-assay samples.

A Sub-Millimicroampere Current Amplifier Utilizing An Unusual Transistor Effect—F. T. May and R. A. Dandl (p. 16)

Due to a very unusual effect that occurred when certain transistors were operated at low current levels, a direct coupled, transistorized, current amplifier has been developed with sensitivity extending below the millimicroampere region of input currents. This amplifier was necessary for the measurement of certain characteristics of the experimental controlled fusion machine, DCX (Direct Current Experiment). Characteristics of the high-gain transistors are presented along with a description of the design and resulting characteristics of a current amplifier with sensitivities of $1/10^{-8}$, $1/10^{-9}$, and $1/10^{-10}$ volt/amp. A possible explanation of the high-gain effect involving the existence of an N-type inversion layer on the P-type base of the silicon transistors of interest suggests the need for more basic experiments on the effects of inversion layers on transistor action.

Gamma-Ray Astronomy in Space in the 50-KEV to 3 MEV Region—L. E. Peterson and R. L. Howard (p. 21)

The γ-ray fluxes in space which arise from nuclear and plasma interactions are expected to be of the order of 10^{-1} to 10^{-2} photons/cm²·sec. A considerable γ-ray background is produced from the interactions of cosmic-ray nucleons. A detection system is described which will allow the measurement of considerably weaker primary fluxes than previous experiments. The apparatus is to be placed on a low-altitude, low-latitude earth satellite. A scintillation counter with a shielding collimator is used in the 50–150 kev energy region. In the 0.3 to 3.0 Mev region, the high particle flux requires the use of an anticoincidence shield around the scintillation counter. A Compton coincidence telescope provides directional collimation in this energy region. Data from the detectors are entered into a temporary binary storage and are read out, together with identification and satellite orientation information, to the payload telemetry system at a one bit/sec rate. The entire apparatus weighs 30 pounds, has a complement of 400 transistors, and operates on 650 mw of power.

The University of Chicago Measurement in Space of Cosmic Rays at Energies above 0.5 MEV—J. B. Lamport and L. J. Petraitis (p. 30)

Since 1958, two groups at The University of Chicago have collaborated in designing several experiments to acquire information relating to the flux of primary cosmic rays having energies normally below the threshold imposed by the geomagnetic field. These experiments are performed with instruments carried by space vehicles. So far in this program, two types of cosmic-ray counter telescopes have been designed and built.

The first type includes a three-fold coincidence counter which has a low-energy threshold for protons at about 75 mev. Such instruments have been flown successfully aboard Explorer VI and Pioneer V. From these flights much useful information was obtained regarding the radiation belts, the Forbush decreases, the solar acceleration of protons, and the gradient of the cosmic-ray flux as a function of distance from the sun. Another instrument of this type is similar to the first but has a lower threshold, 10 mev. The third, which is based upon a different principle, is a two-element, solid-state device for measuring the flux at energies from 0.5 to 10 mev. The latter two instruments are scheduled for first flight aboard a high-eccentricity, earth satellite in late 1961 or early 1962.

This paper contains a detailed description of the first type of telescope. All the instruments of this type are alike except for the detector systems.

Instrumentation for Use with BEV Accelerators—W. A. Higinbotham (p. 34)

The object of the instrumentation is to observe the interaction of high energy protons with various nuclei and to study the characteristics of the short-lived particles which are produced. The tracks of charged particles can be directly observed in photographic film, in cloud and bubble chambers and in spark chambers. Alternatively, a number of radiation detectors may be arranged to give a signal only when a selected type of event takes place.

Some millimicrosecond logic circuits will be described which are used with scintillation and Cerenkov detectors to select and to analyze high energy events. Some recent techniques for data collection and processing will be presented.

Electronic Instruments and Techniques for Plasma Physics Research in the United Kingdom—A. B. Gillespie (p. 40)

The paper describes electronic instruments and techniques which have been developed and used in support of the U.K. plasma physics research program. These generally fall under the headings of machine design, machine control and instrumentation, and plasma diagnostics. To date, most work has been in the latter category which involes a wide variety of measurements ranging from the low frequency end of the spectrum to high energy nuclear radiations. Many of the diagnostic instruments described have been developed for use with ZETA, the large toroidal discharge experiment at the Atomic Energy Research Establishment, Harwell.

Dangers in Safety Systems—E. P. Epler (p. 51)

Reactor safety systems have failed in such a way as to destroy the reactor which they were intended to protect. Safety rods, when released, can, through malfunction, increase rather than decrease reactivity. The electronic and sensing portions of the safety system are usually arranged so that random failures may on rare occasions result in a slight impairment of protection. The total loss of protection should not in itself lead to damage to the reactor. Systematic failures have occurred however, such that the safety and control systems have become coupled either deliberately or through oversight in such a way that the loss of protection and reactor runaway have been caused by a single event.

A Survey of Plasma Instrumentation—Charles B. Wharton (p. 56)

Most of the properties of plasmas are difficult to determine. This is especially true for high temperature, dense, transient plasmas such as encountered in controlled fusion research. If high temperatures and long plasma confinement times are to be achieved the measuring instruments must not perturb the plasma. Since each transient event has its own personality and cannot be expected to be exactly like any preceding or subsequent event, all information (ideally) must be collected each time, without relying on reproducibility. Such plasma instrumentation has been given the name "plasma diagnostics."

A survey and evaluation of several diagnostic techniques is given. Applications of some of the more important ones to specific experiments are shown and a comparison of measured quantities is made.

A tabulation of several commonly used techniques is given in Tables I, II and III, together with a crude evaluation of them. Most of the methods are applicable to general plasma work but some are used only in controlled fusion experimentation.

Diagnostic Techniques of the High Compression and Neutral Injection Experiments—William F. Cummins (p. 71)

In the Multi-Stage Compression Experiment (Toy Top III) we wish to determine the energy and spatial distribution of the plasma components—ions and electrons—from the time of their initial injection into the machine through the heating and containment cycle. An ion analyzer, consisting of a magnetic momentum analyzer followed by an electrostatic analyzer, has been valuable in finding the energy distribution of the plasma ions. The density distribution of the plasma in the experiment's magnetic field has been largely determined with electrostatic extraction probes. Microwave radiometers and double probes have been utilized to find electron temperatures. Li^6 (Eu) and plastic scintillation counters assess the total number and rate of production of neutrons from D-D reactions, which may be confirmed with nuclear emulsions. Electron temperatures in the Single-Stage High Compression Experiment (Table Top II) have been determined from an analysis of the escaping electron flux.

The initial experiments on neutral atom injection into a mirror field geometry (ALICE) require a knowledge of the initial background gas density, neutral beam current, buildup of plasma density, loss rate from the confinement region, and the spatial distribution of the plasma. A magnetically shielded, nude Bayard-Alpert gauge determines the total initial gas pressure, the partial pressure of each atomic specie being found with a mass spectrograph. A copper target calorimeter indicates beam current; while neutral particle detectors on the chamber walls and screened Faraday collectors outside of the mirrors show the build-up and loss of plasma. These instruments, along with electrostatic probes, give the density of the plasma at various points in the system.

DCX-1 Instrumentation—Herman Postma (p. 77)

DCX is a steady state experiment at Oak Ridge which accumulates a plasma by means of high energy injection of diatomic ions into a magnetic mirror geometry. This paper describes the diagnostic techniques that have been employed in DCX-1 to study stability of the plasma accumulated in this device. These include studies of the radiation emitted by this plasma, the plasma density, the plasma potential, and the energy spectra of the particles ejected from the plasma. Experimental measurements are given to illustrate results obtained using these techniques.

Soft X-Ray Spectroscopy of a Hot Plasma—F. C. Jahoda (p. 81)

A Continuous Photometer for Low Concentrations of Uranium in Aluminum Nitrate-Nitric Acid Solutions—E. E. Erickson and C. M. Slansky (p. 83)

In the recovery of uranium from spent nuclear fuel elements, a waste stream from one solvent extraction cycle may be recycled as scrub solution to the prior cycle of the process. The waste solution may, as a result of this recycle, enter large columns where a critical accumulation of U-235 is prevented by concentration rather than size limitations. A continuous monitor was required to detect changes in uranium content of these recycled streams. Evaluation of the time lags and dilution effects in the sampling system showed that the sample monitored would increase by 2 grams uranium/liter before the situation could become dangerous. An in-line photometer was developed that could detect changes in uranium concentration of about 0.02 gram/liter in the 0–2 gram/liter range; it was set to alarm at a maximum uranium content of 1 gram/liter.

The photometer is basically a simple instrument, depending primarily on the use of a photovoltaic cell to compensate for the logarithmic characteristics of light absorption. The sample stream is analyzed in the wavelength region of 425 millimicrons, where uranium absorbs light, and the instrument is standardized in the 525 millimicron region, where uranium does not absorb. A Brown amplifier is utilized to adjust the instrument output to the recorder for the reference cycle to zero, so that the recorded output for the measuring cycle is a linear function of uranium concentration. Fail-safe techniques incorporated account for the bulk of the instrument hardware.

Several problems remain to be solved before use of the photometer can be extended to other applications, chief among which is interference by iron. Extension of the principles of the measurement suggests possible means of measuring individual constituents in multicomponent solutions.

Inline Applications of Gamma Monitoring and Uranium Colorimetry—M. J. Kelly, J. W. Landry, T. S. Mackey, and R. W. Stelzner (p. 89)

Techniques using gamma scintillation spectrometry and colorimetry for continuous analysis of radiochemical processing plant streams are discussed.

The use of a continuous gamma monitor on an organic processing stream is described with emphasis on the practical difficulties encountered. A study substantiated by experiment on an improved gamma monitor is presented. The use of colorimetry to determine uranium concentration in processing streams is considered. Tests and results are reported for a relatively simple colorimeter and discussion is presented concerning more elaborate techniques.

Formulation and Implementation of Criticality Control—A. H. Dexter and J. N. Wilson (p. 94)

A program to control the behavior of personnel during a criticality incident has been put into effect at Savannah River Laboratory. The objectives are to minimize exposure to radiation and its effects by prompt evacuation and by speedy identification of those individuals who received the highest doses.

The many considerations that had to be taken into account in the formulation of the basic philosophy, the specifications of the necessary alarm and dosimetry systems, and the implementation of the program are described in detail.

Failsafe AD Hi-Lo Limit Detector—E. Bianchi and P. Lenk (p. 101)

A very accurate alternating current limit detector is presented. Its design philosophy, specially suited for multi-channel applications is based on dynamic monitoring of both limits through the same channel.

Completely solid-state, this limit detector exhibits an accuracy of ±0.1% over a temperature range from 0 to 45° C. The design is inherently failsafe; and reliability evaluations were performed for a 1000 channel application.

Design philosophy, circuit diagrams and system operation are presented in detail and discussed.

A Fast Response Monitor for Airborne Uranium Dust—William O. Gentry and Garland B. Seaborn (p. 104)

A fixed filter paper alpha air monitor has been developed which will sound an alarm within 15 minutes for a uranium dust in air concentration of 6 picocuries per cubic foot. The cause and effect of background variations are discussed. Typical natural background data accumulated over a six months period are included.

A prototype instrument has been built and operated satisfactorily. The final model of the instrument which is in the design stage will be housed in a cabinet 13 inches high, 15 inches deep, and 21½ inches wide. Pictorial and schematic diagrams are included.

Effect of Thermal Neutron Irradiation on Thermocouples and Resistance Thermometers—C. W. Ross (p. 110)

Digital Computers in Nuclear Power Applications—J. Prades and Y. Panis (p. 119)

There are a number of interesting applications of digital control computers in the nuclear power field which demand highly reliable, high speed, automatic operation. These include (a) the monitoring of radioactivity levels, (b) conventional monitoring and data recording, (c) operation calculations on reactors, (d) economic calculations and plant optimization. In the case of gas-cooled reactors with a thousand or more cooling channels one problem is to detect and localize dangerous increases in radioactivity due to gas leakage in the face of background radiation which is subject to normal statistical variations of random types and variation in the mean level. This application of digital computers for monitoring radioactivity will be discussed and an existing installation will be described.

Factors Affecting Design and Sale of Transistorized Instrumentation—E. B. Hubbard (p. 127)

This paper concludes that all solid state systems are not necessarily the best for reactor control instrumentation. However, market requirements dictate that the growing preference for solid state systems outweighs all other considerations.

Nuclear Protective System Design for Reliability—V. S. Underkoffler, J. L. Cockrell and J. H. Magee (p. 130)

Three significant and almost parallel developments have taken place in recent years with respect to the development of electronic instrumentation for nuclear reactor protective systems. First, of course, has been the widespread use of solid-state components as active circuit elements. Justification for this need had its roots in the projected estimates of improved circuit mean life, reduced power requirements, smaller size, etc. The second development was of a more fundamental nature in that it recognized that the electronic protective system problem was a switching circuit and logic design problem. A natural consequence of this recognition was the desire to convert the transducer analog signals to bistate signals, and then handle the bistate signals with solid state logic devices, which in turn, control the on-off status of a power supply serving as the energy source for the safety rod actuators or similar devices. The third development arose from the desire to reduce the probability of reactor shutdown to instrument failures and introduced the "two of three" coincidence or majority decision concept. This majority decision concept could easily be accomplished if signals were in bistate form, since it represents a rather simple threshold logic problem readily solvable by a host of devices. With respect to these three developments the purpose of this paper is to provide system design concepts for reliability enhancement and to present system building blocks for a practical realization of these concepts. Although the ideas presented here are pointed toward nuclear reactor protective systems, it is felt that these same approaches can be applied to general systems design.

Survey of Space Nuclear Instrumentation—George H. Ludwig (p. 149)

Use of Electrostatic Probes in Plasma Physics—Francis F. Chen (p. 150)

The principles of classical Langmuir probe theory are outlined to indicate the type of information theoretically obtainable by this technique. In the dense, hot plasmas and high magnetic fields encountered in modern plasma physics, difficulties arise both in instrumentation and in interpretation of the measurements. Because of these difficulties, perhaps the usefulness of probes lies primarily in the measurements of local fluctuating electric fields. The types of circuits required in probe work are described in general terms.

Our Changing Views of Reactor Instrumentation—A. R. Pearson and C. G. Lennox (p. 155)

At Chalk River emphasis has been placed on triplicated instrumentation for the reliable and safe control of nuclear reactors. Experience with the three-channel systems on the NRX and NRU reactors is given, along with some comments on the problems that have arisen.

A semiconductor current amplifier, an alarm circuit and a logarithmic amplifier, are described and provide examples of instruments that meet the new design requirements.

Application of solid-state digital techniques to reactor instrumentation becomes attractive when large numbers of signals have to be examined in detail. A semiconductor system is described that is capable of scanning several hundred low-level signals in less than 0.1 sec. The switching circuits used are shown. The application of these systems to plant monitoring and control is outlined.

Instrumentation for Measurements in the Van Allen Belts—W. Wheltley (p. 165)

Space Electronics and Telemetry

Vol. SET-7, No. 4, December, 1961

Vibration, Shock, and Acoustic Noise Associated with Space Vehicles—D. G. Douglas (p. 87)

Analysis of telemetered vibration data received from Transit-Courier launchings, using Able-Star space vehicle equipment, indicates three major sources of vibration. These are missile engine acoustic noise at liftoff, aerodynamic effects as the missile reaches Mach 1, and separation shock when the second-stage vehicle is launched. These measurements were taken at the Able-Star instrument platform; thus very little vibration if any was induced by turbopumps, alternators, or snap-operating valves. Vibration decreases to extremely low levels between the three periods of major excitation; therefore, transmission of vibration directly from the rocket engine through the missile structure has negligible effect on space vehicle equipment.

Data obtained are useful in establishing equipment environmental test criteria. Specifications requiring sinusoidal combined with random vibration are indicated by some functions and shaped random, emphasizing certain frequency regions, are suggested for other functions.

Plasma-Enhanced Breakdown of Telemetry Antennas During Ballistic Missile Powered Flight—H. A. Poehler (p. 90)

Attention is directed to a previously unrecognized phenomenon, plasma-enhanced breakdown of ballistic missile antennas. Plasma-enhanced antenna breakdown is a general phenomenon, and has been observed during flight tests of a variety of ballistic missiles. Examples chosen from telemetry signal-strength records of recent ballistic missile flights are presented. The need for the consideration of this effect in the design of the staging sequence of ballistic missiles is pointed to.

Phase-Locked Loops for Electronically Scanned Antenna Arrays—M. Breese, R. Colbert, W. Rubin, and P. Sferrazza (p. 95)

This paper describes the ATHESA (Automatic THree-dimensional Electronic Scanned Array) technique for achieving very high antenna gain for deep-space communication systems. Phase coherence over a band of frequencies is established on reception automatically and continuously for either stationary or moving targets by phase-locking the signals from antenna elements separated by many wavelengths to a reference antenna element.

A New Class of FM Subcarrier Oscillators—Leonard Baker (p. 100)

This paper describes a new class of millivolt-controlled subcarrier oscillators developed especially for operation in the rugged environments of orbital programs. The oscillators cover all the standard IRIG channels. The full IRIG signal-frequency response is achieved at a modulation index equal to five.

A less-than-unity-gain, long-tailed pair, silicon-transistor, dc amplifier and a greater-than-unity-gain RC network are combined to form a sine-wave oscillator. The center frequency is deviated ±15 per cent by a ±10 millivolt control signal. Excellent linearity characteristics are achieved. A unique feature of the circuit is the inherent deviation limiting. Inductors are avoided in the oscillator design.

The passive RC network synthesis procedure is developed to realize 1) a voltage gain greater than unity, 2) the proper percentage of deviation with excellent linearity characteristics, and 3) inherent deviation limiting to contain the FM signal spectra within their assigned channels.

The center frequency is deviated by controlling the relative gain of the long-tailed pair type of differential dc amplifier. A pure sine wave is generated without additional filtering.

The balanced nature of the circuit yields excellent center frequency stability from 0°C to +85°C. Center frequency time drift provides an improvement of at least one order of magnitude over present state-of-the-art millivolt-controlled oscillators.

Measured data is presented to support the procedures used in the design. The oscillator and associated circuitry are contained in a 4½-cubic-inch package that weighs 4 ounces.

Transient Analysis of a Phase-Locked Loop Discriminator—A. K. Rue and P. A. Lux (p. 105)

This paper analyzes the step and impulse response of a phase-locked loop discriminator with an optimum-type RC filter. In previous papers the multiplier in a phase-locked loop is linearized in order to obtain the optimum filter design, and the response to a large step is not considered.

In this paper the system is treated as a second-order nonlinear system, and phase plane analysis is used to determine the transient response.

Some very interesting results were found analytically and then verified experimentally. For example, it was found that the phase error can be greater than π and the system will still lock into the carrier after an unusual transient response.

Contributors (p. 112)

Correction to "Noise Figure Measurements Using Celestial Sources"—A. M. Faries and H. A. Turner (p. 113)

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. 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 **3983**
Solution of an Integral Equation Occurring in the Study of Certain Wave-Propagation Problems in Layered Media—I. M. Longman. (*J. Acoust. Soc. Am.*, vol. 33, pp. 954–958; July, 1961.) A method of inversion of an integral equation arising in the solution of the problem of reflection or refraction of a spherical sound-wave pulse at a plane interface in a layered elastic medium.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears each June and December at the end of the Abstracts and References section.

The Index to the Abstracts and References published in the PROC. IRE from February, 1960 through January, 1961 is published by the PROC. IRE, May, 1961, Part II. It is also published by *Electronic Technology* and appears in the March, 1961, issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

534.231 **3984**
The Computation of Far-Field Radiation Patterns from Measurements made near the Source—C. W. Horton and G. S. Innis, Jr. (*J. Acoust. Soc. Am.*, vol. 33, pp. 877–880; July, 1961.) Two methods based on a) the use of Green's functions, and b) a proposal made by Pachner (1933 of 1956) are analyzed and compared.

534.232 **3985**
Sound Radiation from Prolate Spheroids—G. Chertock. (*J. Acoust. Soc. Am.*, vol. 33, pp. 871–876; July, 1961.) An analysis of the radiation from a spheroid providing expressions for pressure and velocity distributions, radiation impedance and directivity factor in each mode.

534.232–8:538.652 **3986**
Ultrasonic Transducer Design cuts Losses—H. L. Morgan. (*Electronics*, vol. 34, pp. 70–73; August 18, 1961.) Magnetostrictive tape and insulating material are wound into a cylinder. Possible methods of cooling high-power transducers are discussed.

534.24–14 **3987**
Theoretical and Experimental Study of Underwater Sound Reverberation—B. F. Cron and W. R. Schumacher. (*J. Acoust. Soc. Am.*, vol. 33, pp. 881–888; July, 1961.) Fluctuations of amplitude and phase associated with the return of a single-frequency pulse were studied. By using the correct filter, a considerable gain in signal quality could be achieved.

534.522.1 **3988**
Diffraction of Light by Ultrasonic Waves of Various Standing-Wave Ratios—B. D. Cook and E. A. Hiedemann. (*J. Acoust. Soc. Am.*, vol. 33, pp. 945–948; July, 1961.)

534.79 **3989**
Investigation for the Complete Representation of the Laws of Loudness Growth with any Complex Form of Noise Excitation—H. Niese. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 5–17; February, 1961.) Extension of investigations of the inertia effect in hearing based on earlier experimental results (*e.g.*, 11 of February).

534.844.1 **3990**
The Influence of the Diffusing Area and the Dimensions of a Reverberation Chamber on the Measured Coefficient of Sound Absorption—U. Kath and W. Kuhl. (*Acustica*, vol. 11, no. 1, pp. 50–64; 1961. In German.)

534.88:621.396.677 **3991**
Improved Directivity using Synchronous Demodulation—D. G. Tucker. (*Acustica*, vol. 11, no. 1, pp. 45–49; 1961.) In a limited application greatly improved directional response can be obtained by the use of synchronous demodulation in conjunction with a receiving array whose secondary responses are in reverse phase relative to the main beam. Compared with a multiplicative system, response suppression is better but the narrowing of the main lobe is not so great.

534.88:621.396.677.3 **3992**
Array Steering in a Layered Waveguide—C. S. Clay. (*J. Acoust. Soc. Am.*, vol. 33, pp. 865–870; July, 1961.) Time-delay steering of the array will give several maxima and minima but a mechanically steered array will give one well-defined maximum.

621.395.625.3 **3993**
Mechano-electrical Analogues for the Filtering of Electroacoustic Drive Systems—W. Wolf. (*Tech. Mitt. BRF, Berlin*, vol. 5, pp. 1–9 and 49–58; March and June, 1961.) The problem of overcoming fluctuations in mechanical tape transport mechanisms is investigated with the aid of electromechanical analogs and measurements on equivalent electrical circuits. See also 1400 of May.

ANTENNAS AND TRANSMISSION LINES

621.315.212 **3994**
Theory and Design of a Coaxial Supporting Bead made of Composite Dielectrics—G. H. Owyang. (*Canad. J. Phys.*, vol. 39, pp. 926–934; June, 1961.) A formula correlating the standing-wave ratio with parameters of the bead is derived. An example shows that ratios <1.003 are possible over a frequency range of 13:1.

621.372.2:621.372.414 **3995**
Contribution to the Problem of the Homogeneous Lecher Line with Nonlinear Capacitive Loading—G. Böhme. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 25–30; February, 1961.) An approximate solution of the line equations is derived by means of an iterative method. The application of the system to parametric amplification is considered.

621.372.21.018.78 **3996**
A New Method of Calculating Signal Distortion in High-Frequency Cables—K. Rihaczek and H. Wörner. (*Nachrtech. Z.*, vol. 14, pp. 119–124; March, 1961.) The transmission factor is evaluated in the form of a power series or, in approximation, by polynomials. Examples are given of calculations of optimum carrier frequency for low signal distortion in pulse transmissions systems.

621.372.8:537.226 3997

Cylindrical Dielectric Waveguide Modes—E. Snitzer. (*J. Opt. Soc. Amer.*, vol. 51, pp. 491–498; May, 1961.) Sets of solutions are derived for both hybrid modes, giving cutoff, field distribution and conditions far from cutoff. Field plots for the lower-order modes are shown.

621.372.82:536.48 3998

Simple Approach to Cryogenic Waveguide—D. K. Milne. (*J. Sci. Instr.*, vol. 38, p. 335; August, 1961.) The advantages of using a "quasi-elliptical" guide are discussed.

621.372.823 3999

On the Nomenclature of TE_{01} Modes in a Cylindrical Waveguide—C. T. Tai. (Proc. IRE, vol. 49, pp. 1442–1443; September, 1961.) A modification of notation produces a more logical nomenclature for all the modes in a cylindrical waveguide. The number of propagation modes for a given waveguide may be determined from the graph presented.

621.372.85 4000

Variational Principle Relating to Systems of Equations with Common Eigenvalues. Application to the Calculation of the Propagation Constant of Electromagnetic Waves in Guides with Anisotropic Media—L. Cairó and T. Kahan. (*C.R. Acad. Sci.* (*Paris*), vol. 251, pp. 1865–1867; October 31, 1960.) A variational principle of general character is established and is applied to the case of systems of equations, giving the propagation constant of gyromagnetic waveguides. See 707 of 1959 (Hauser).

621.372.851.2 4001

The Construction of Narrow-Band Microwave Diaphragm Filters in Waveguide—J. Windel. (*Nachrtech.*, vol. 11, pp. 115–121; March, 1961.) Details of the design and performance of inductive irises are given.

621.396.67 4002

Aerials—H. Page. (*Proc. IEE*, pt. B, vol. 108, pp. 473–489; September, 1961.) A review of progress. 74 references.

621.396.67 4003

Theoretical Techniques for Handling Partially Polarized Radio Waves with Special Reference to Antennas—H. C. Ko. (Proc. IRE, vol. 49, pp. 1446–1447; September, 1961.) The power available from the antenna can be expressed in terms of the Stokes parameters or the density matrix.

621.396.67 4004

Experimental Investigations on the Applicability of the Principle of Superposition in the Design of Combinations of Aerials—E. Hüttmann. (*Nachrtech.*, vol. 11, pp. 93–96; February 1961.) Consideration of the distribution of displacement currents to obtain a more precise picture of the radiation mechanism of combined antennas, with particular reference to a special type of V-antenna.

621.396.67 4005

Attenuation Effect of a Tiled Roof in the Case of Loft Aerials—H. Hügli. (*Tech. Mitt. PTT*, vol. 39, pp. 132–134; April, 1961. In German and French.) Brief note on the measurements to be carried out to determine the attenuation. Results are given of measurements made with different antennas in the range 50–950 Mc under a dry and a wet roof.

621.396.67(083.7) 4006

Categories and Types of Aerials—(*Nachrtech. Z.*, vol. 14, pp. 92–96; February, 1961.) Glossary of 75 proposed terms and definitions with, in some cases, equivalent terms in English and French.

621.396.676:551.510.535 4007

The Electrically Short Antenna as a Probe for Measuring Free Electron Densities and Collision Frequencies in an Ionized Region—King, Harrison, and Denton. (See 4134.)

621.396.677:534.88 4008

Improved Directivity using Synchronous Demodulation—Tucker. (See 3991.)

621.396.677:621.396.43 4009

Antenna-Beam Configurations in Scatter Communications—A. T. Waterman, Jr. (Proc. IRE, vol. 49, p. 1452; September, 1961.) There is a difference in the directivities of the scattering processes which operate in tropospheric scattering and in incoherent scattering by free electrons. The consequences are discussed in terms of the desirable polar diagrams of antenna systems used in scatter communications.

621.396.677.3 4010

Super-Gain Aerials—T. Heller. (*Nachrtech. Z.*, vol. 14, pp. 113–118; March, 1961.) The principle of superdirective antenna arrays is discussed and the design is given of three end-fire and side-fire arrays with 15, 21 and 31 radiators and spacings of 0.34 λ, 0.3 λ and 0.2 λ, respectively.

621.396.677.3:534.88 4011

Array Steering in a Layered Waveguide—Clay. (See 3992.)

621.396.677.7 4012

Omnidirectional *K*-Band Antenna uses Slots, Probes and Horns—G. J. Monser and E. D. Botkin. (*Electronics*, vol. 34, pp. 54–55; August 18, 1961.) The nominal vertical beamwidth is 30° centered on the horizon. The frequency range covered is 1.5 to 1 with a voltage SWR less than 3:1.

621.396.677.83 4013

Gain Limit and Tolerances of Big Reflector Antennas—A. Consortini, L. Ronchi, A. M. Scheggi and G. Toraldo di Francia. (*Alta Frequenza*, vol. 30, pp. 232–243; March, 1961. In English.) The effects of manufacturing errors on antenna gain are studied for the parabolic antenna. Assuming that the construction tolerances are greater for larger antennas, an optimum size antenna is found for the expected maximum gain.

AUTOMATIC COMPUTERS

681.142 4014

High-Speed Analogue/Digital Converter—Y. Hazoni. (*C.R. Acad. Sci.* (*Paris*), vol. 251, pp. 343–345; July 18, 1960.) Details are given of a fully transistorized unit for the conversion of pulse-amplitude data to numbers of pulses. Use is made of a tunnel-diode/transistor combination in the blocking oscillator circuit to produce oscillations at a frequency of a) 10 Mc or b) 40 Mc. The over-all dead time of the converter is reduced ten times.

681.142:535.215 4015

Optical Erasure of EL-PC and Neon-PC Storage Elements—J. A. O'Connell and B. Narken. (Proc. IRE, vol. 49, pp. 1438–1439; September, 1961.) A new method of selective storage and erasure using "optrons" is described. A single light source of constant intensity with emission within the sensitivity range of the photoconductor will latch optrons with short pulses (<30 msec) and erase them with long pulses (>200 msec).

681.142:537.227 4016

Obtaining Nondestructive Read-Out with Ferroelectric Memories—A. B. Kaufman. (*Electronics*, vol. 34, pp. 47–51; August 25, 1961.) The development and initial tests of a monomorphic, polycrystalline Ba titanate storage cell used as a three- or four-terminal device.

681.142:538.221 4017

The Magnetic Flux Counter—G. Müller. (*Nachrtech. Z.*, vol. 14, pp. 81–87; February, 1961.) The principle and application of multistable magnetic storage systems are discussed.

681.142:538.632.621.375.432 4018

Feedback Amplification Improves Hall-Effect Multipliers—R. A. Greiner. (*Electronics*, vol. 34, pp. 52–55; August 25, 1961.) A transistor feedback amplifier supplies a field coil current in phase with the input signal. A dc component is added to increase the common-mode signal rejection.

681.142:621.318.134 4019

A 0.7-Microsecond Ferrite Core Memory—W. H. Rhodes, L. A. Russell, F. E. Sakalay and R. M. Whalen. (*IBM J. Res. & Dev.*, vol. 5, pp. 174–182; July, 1961.) A storage capacity of 73,728 bits is achieved with an instruction repetition rate of 1.47 Mc.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.064:621.372.44 4020

Switching Processes in D.C. Circuits with Nonlinear Resistance and Nonlinear Inductance—W. Böning. (*Arch. Elektrotech.*, vol. 46, pp. 103–124; March 15, 1961.) Normalized differential equations are derived for the analysis of switching processes involving nonlinear resistances with semiconductor- or tube-type characteristics.

621.314.2.029.53/.62:621.372.832.6 4021

Design of Wide-Band Shielded Toroidal R.F. Transformers—R. S. Berg and B. Howland. (*Rev. Sci. Instr.*, vol. 32, pp. 864–865; July, 1961.) Details of transformers for use with magic-T or hybrid junctions in the frequency range 1–100 Mc.

621.314.22:621.374 4022

Determination of Core Size in Pulse Transformer Design—C. F. Wilds. (*Electronic Eng.*, vol. 33, pp. 566–573; September, 1961.) The construction of a design chart is described and its application is illustrated.

621.318.57:538.221 4023

Toroids of Square-Loop Ferrite Material as Switching Elements—W. Hillberg. (*Arch. elekt. Übertragung*, vol. 15, pp. 145–152; March, 1961.) A differential equation is derived which describes magnetic reversal processes in square-loop ferrites neglecting the effect of eddy currents. The application of this analysis to practical switching circuits is considered.

621.318.57:621.382.23:621.374 4024

Tunnel Diodes Increase Digital-Circuit Switching-Speeds—W. V. Harrison and R. S. Foote. (*Electronics*, vol. 34, pp. 154–156; August 11, 1961.) GaAs tunnel diodes are used with transistors in high-speed switching circuits suitable for pulse-generator operation.

621.318.57:681.142 4025

Application of Karnaugh's Mirror Symbols to Transfluxors—F. Schreiber. (*Frequenz*, vol. 15, pp. 58–61; February, 1961.) A transfluxor symbol is proposed which is based on the system used by Karnaugh (2207 of 1955) for magnetic-core switching circuits.

621.372.413:621.385.644.3 4026

An Analysis of a Cylindrical Cavity with Radial Vanes—A. Singh and R. A. Rao. (*Proc. IEE*, pt. B, vol. 108, pp. 550–555; September, 1961.) The procedure involves the use of variational methods for obtaining successively better approximations to the resonance frequencies and the field configuration.

621.372.414:621.372.8 4027

An Analysis of the Travelling-Wave Cavity—D. Karayianin and C. A. Morrison. (*Proc.*

IEE, pt. B, vol. 108, pp. 545–549; September, 1961.) The measured response of a traveling-wave cavity differs from that predicted by simple theory. An internal mismatch can explain the observed response.

621.372.52 4028

Rules for the Verification of the Stability of Normal Linear Negative-Feedback Systems—A. Lepschy and A. Ruberti. (*Alta Frequenza*, vol. 30, pp. 285–292; April, 1961.) The stability criterion and rules are formulated and the procedure of applying the rules to polar diagrams and curves of the system function is described.

621.372.54 4029

A Filter System with Favourable Pulse Characteristics—K. Uhl. (*Arch. elekt. Übertragung*, vol. 15, pp. 109–114; March, 1961.) Analysis of filter performance under pulse conditions on the basis of the distribution of poles and zeros of the transfer function. A practical filter circuit is derived and the pulse amplitude calculated.

621.372.54:534.1 4030

Mechanical Filters with Flexural Coupling—M. Börner. (*Arch. elekt. Übertragung*, vol. 15, pp. 175–180; April, 1961.) The use of flexural coupling between longitudinal resonators makes possible the construction of very small mechanical filters for operation around 500 kc. The characteristics of such a system are calculated and two practical examples are given. See also 2155 of 1959 (Börner, *et al.*).

621.372.542.2 4031

The Design of Filters with the Aid of the Catalogue of Normalized Low-Pass Filters—R. Saal. (*Frequenz*, vol. 15, pp. 111–121; April, 1961.) Filter design by the insertion-loss method is facilitated by the use of tables prepared with the aid of an electronic computer [see 2051 of 1957 (Glowatzki)]. The design of Tchebycheff-type and Cauer-parameter low-pass filters is considered in detail.

621.372.543 4032

Diagrams for the Design of Double-Tuned Band Filters with Arbitrary Coupling—Sittner and Minner. (*Frequenz*, vol. 15, pp. 93–99; March, 1961.) The design curves and equations given are based on filter parameters which can be obtained directly by measurement in the filter circuit.

621.372.543.2:534.143 4033

Insertion Loss of Electromechanical Coupling Filters—E. Trzeba. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 17–20; February, 1961.) The insertion loss is defined and design formulas are derived for minimizing this loss. See also 2497 of August.

621.372.543.2:534.143 4034

Experiences with Mechanical Band Filters—H. J. Rohde. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 31–38; February, 1961.) Problems arising in the construction of electromechanical filters are discussed and some practical types of filters are described.

621.372.543.22 4035

A Crystal Filter of Increased Bandwidth and its Synthesis—W. Herzog. (*Nachrtech. Z.*, vol. 14, pp. 153–157; March, 1961.) The method given in 1750 of June is applied to a lattice-type filter with a crystal in each bridge arm.

621.372.55:621.397.6 4036

A Differentiating Equalizer with Transistors—H. Anders and U. Reese. (*Rundfunktech. Mitt.*, vol. 5, pp. 66–75; April, 1961.) The equipment described is designed for emphasizing the upper frequencies of the video signal band without phase distortion.

621.372.6 4037

A Synthesis Procedure for an n-Port Network—D. Hazony and H. J. Nain. (Proc. IRE, vol. 49, pp. 1431–1432; September, 1961.) The method is based on the Bott-Duffin procedure for two-port networks.

621.373:621.372.44:621.382.23 4038

Bistable Systems of Differential Equations with Applications to Tunnel-Diode Circuits—J. K. Moser. (*IBM J. Res. & Dev.*, vol. 5, pp. 226–240; July, 1961.) A mathematical theory is developed for nonlinear circuits with at least two stable, steady states. The theory is applied in detail to circuits containing one or two tunnel diodes.

621.373.42 4039

New Wide-Range Oscillator Technique—L. M. Sargent. (*Brit. Commun. Electronics*, vol. 8, pp. 600–603; August, 1961.) A modification of the Wien bridge RC oscillator circuit is used in an instrument with a range of 10 cps–10 Mc.

621.373.422:621.382.23 4040

Crystal-Stabilized Tunnel-Diode Oscillators—J. J. Nagle. (*Electronics*, vol. 34, pp. 40–42; September 1, 1961.) Theory and design procedure is given for a 40-Mc oscillator with frequency stability of 7.5 parts per million per degrees Centigrade.

621.373.43:621.382.23 4041

One-Tunnel-Diode Binary—A. L. Whetstone and S. Kounosu: R. A. Kaenel. (Proc. IRE, vol. 49, p. 1445; September, 1961.) Comment on 1761 of June and author's reply.

621.373.52 4042

An Analysis of the Modes of Operation of a Simple Transistor Oscillator—J. F. Gibbons. (Proc. IRE, vol. 49, pp. 1383–1390; September, 1961.) An analysis predicting two modes of oscillation: a two-terminal transit-time mode and a three-terminal mode.

621.374:621.382.3 4043

Transistor Precision Pulse Shaper with Short Recovery Time—I. De Lotto. (*Alta Frequenza*, vol. 30, pp. 219–223; March, 1961. In English.) The circuit is designed to minimize the effects of temperature changes and individual variations in transistors. The pulse duration varied by <1 per cent for a temperature range of 25–45°C and by <2 per cent in ten similar circuits.

621.374.3 4044

A Neuristor Prototype—A. J. Cote, Jr. (Proc. IRE, vol. 49, pp. 1430–1431; September, 1961.) A description of an electronic lumped-element model of the neuristor proposed by Crane (IRE Trans. on Electronic Computers, vol. EC-9, pp. 370–371; September, 1960.)

621.374.3:621.382.233 4045

Some Circuit Applications of Avalanche Devices—G. M. Ettinger. (*Proc. IEE*, pt. B, vol. 108, pp. 563–569; September, 1961.) Applications of *pn-pn* avalanche diodes in analog computers, process control, oscillography and telemetry are discussed.

621.374.32 4046

Study of a Diode Bridge Nanosecond Resolution Coincidence Circuit—S. C. Pancholi and N. K. Saha. (*Proc. Nat. Inst. Sci., India, A*, vol. 27, pp. 155–160; March 26, 1961.) A highly stable circuit is described which can provide a resolution down to 0.7 nsec.

621.374.44 4047

Frequency Components in the Output of a Harmonic Generator Driven by the Sum of Two Sine Waves—G. Einarsson. (*Ericsson Tech.*, vol. 17, no. 1, pp. 139–148, 1961.) A frequency analysis is given for the output of a generator which emits a constant-amplitude pulse at each zero crossing of the input voltage consisting of the sum of two sine waves raised to an arbitrary dc level.

621.374.44 4048

A Random Pulse Generator with Variable Mean Rate—J. L. Douce and B. G. Leary. (*Proc. IEE*, pt. B, vol. 108, pp. 524–528; September, 1961.) Method of producing a random-pulse train with predetermined statistical properties.

621.375.018.756:621.317.7 4049

Amplifier for the Generation of Accurately Controlled Microsecond Pulses—J. B. Gunn. (*Rev. Sci. Instr.*, vol. 32, pp. 804–807; July, 1961.) A description is given of a pulse amplifier whose output voltage can be accurately controlled by the setting of a ten-turn potentiometer. The output amplitude and pulse lengths are variable up to 1000 v, 2.6 a, and 10 μsec, respectively.

621.375.024:623.451–519 4050

Designing Hybrid D.C. Amplifiers to Withstand Missile Environments—R. L. Konigsberg. (*Electronics*, vol. 34, pp. 157–159; August 11, 1961.) The amplifier comprises one tube and Si transistors, and is based on the Goldberg principle for dc drift correction.

621.375.1:621.391.822 4051

Actual Noise Measure of Linear Amplifiers—K. Kurokawa. (Proc. IRE, vol. 49, pp. 1391–1397; September, 1961.) A new noise measure is proposed which includes the noise contribution from the load. It is of particular use in comparing amplifiers whose terminations are not necessarily optimized.

621.375.233 4052

Automatic Gain Control for Superregenerative Amplifiers—J. H. Kuck. (*Electronics*, vol. 34, pp. 76–79; July 21, 1961.) The circuit gives stable operation for long periods and maintains output pulses constant over a 90-db range of input signals.

621.375.9:538.569.4 4053

'Nonadiabatic' Behaviour in a Low-Frequency Beam-Type Maser—T. Shimizu and K. Shimoda. (*J. Phys. Soc. Japan*, vol. 16, pp. 777–791; April, 1961.) Characteristics are discussed theoretically when the adiabatic focusing approximation can no longer be assumed.

621.375.9:538.569.4 4054

A Travelling-Wave Maser Using Chromium-Doped Rutile—E. S. Sabisky and H. J. Gerritsen. (Proc. IRE, vol. 49, pp. 1329–1330; August, 1961.) The active material is of small cross section and is connected to standard size X-band waveguides by dielectric tapers. Performance figures quoted show appreciable gain per unit length with a very large tunable bandwidth obtained without the use of any external slowing structure.

621.375.9:538.569.4:538.222 4055

Maser Action in Emerald—F. E. Goodwin. (*J. Appl. Phys.*, vol. 32, pp. 1624–1625; August, 1961.) The successful operation of a single-cavity reflection-type maser amplifier at 10 Gc using synthetic emerald is reported.

621.375.9:538.569.4:538.222 4056

Maser Action in Ruby by Off-Resonance Pumping—F. Arams and M. Birnbaum. (Proc. IRE, vol. 49, pp. 1426–1427; September, 1961.) Population inversion peaks have been observed using a pump frequency as much as 1400 Mc from the center frequency. The experimental results are discussed and explained.

621.375.9:538.569.4:621.396.43:551.507.362.2 4057

The Maser and its Application to Satellite Communication Systems—P. Hlawiczka. (*J. Brit. IRE*, vol. 22, pp. 265–270; September, 1961.) The properties of the traveling-wave maser with a comb structure are discussed with particular reference to the 2–10-Gc frequency range. Performance is estimated and compared with a parametric amplifier in a practical antenna system.

621.375.9:621.372.44 4058

A Tunable Reactance-Type Straight Amplifier for the U.H.F. Range—K. Bomhardt. (*Arch. elekt. Übertragung*, vol. 15, pp. 153–160; March, 1961.) An experimental variable-capacitance amplifier tunable in the range 480–750 Mc, with a pump frequency of 1900 Mc, is described. With a 10-Mc bandwidth the noise figure is better than that obtainable with a conventional earthed-grid triode amplifier. See also 95 of January (Maurer, *et al.*).

621.375.9:621.372.44:621.372.2 4059

Gain Saturation in a Travelling-Wave Parametric Amplifier—A. Jurkus and P. N. Robson. (PROC. IRE, vol. 49, pp. 1433–1434; September, 1961.) There is a periodic transfer of power along the line, between signal, idler and pump frequencies (see 1927 of 1960). The resultant reduction in gain with increase of signal strength is shown graphically.

621.375.9:621.372.44:621.382.3 4060

Parametric Amplification with Transistors—U. L. Rohde. (*Wireless World*, vol. 67, pp. 493–499; October, 1961.) A note on the operation and characteristics of a parametric amplifier using transistors operating beyond cutoff frequency.

621.375.9:621.372.44:621.385.63 4061

Microwave Amplification in Electrostatic Ring Structures—Bass. (See 4365.)

621.375.9:621.382.23 4062

On the Cascaded Tunnel-Diode Amplifier—P. M. Chirlian. (PROC. IRE, vol. 49, pp. 1434–1435; September, 1961.) A midband equivalent circuit of an artificial transmission line is described for use with tunnel diodes in cascade amplifiers (see 3044 of 1960). Only one diode is needed in each stage and the shunt resistor is an ordinary passive component.

GENERAL PHYSICS

537.222.1 4063

The Electrostatic Field in Metals, particularly at a Surface against an Insulator—F. Ollendorff. (*Arch. Elektrotech.*, vol. 46, pp. 1–26; February 15, 1961.)

537.312.8:530.145 4064

Quantum Theory and Transverse Galvanomagnetic Phenomenon—E. N. Adams and T. D. Holstein. (*J. Phys. Chem. Solids*, vol. 10, pp. 254–276; August, 1959.) A quantum theory of electrical conduction in crossed electric and magnetic fields is given for the limit of very weak scattering.

537.525.72:538.569 4065

A Gas Discharge Generated by Microwaves which Glows Free in Space at the Focus of a Closed Ellipsoid of Revolution—J. Geerk and H. Kleinwächter. (*Z. Naturforsch.*, vol. 16a, pp. 320–321; March, 1961.) Better results were obtained than in earlier experiments (2147 of July.)

537.56 4066

Electric-Field-Induced Anisotropies in an Inhomogeneous Plasma—L. Wetzel. (*Phys. Rev.*, vol. 123, pp. 722–726; August 1, 1961.) Anisotropies in electron diffusion and conductivity can be produced in the absence of a magnetic field.

537.56 4067

Nonlinear Time-Dependent Plasma Oscillations—D. Montgomery. (*Phys. Rev.*, vol. 123, pp. 1077–1078; August 15, 1961.) The Laplace transform technique used to solve the problem of the first-order motions in an unbounded, rarified, electron plasma is modified to solve the problem to arbitrarily high order.

537.56:538.56 4068

Some Boundary Value Problems Involving Plasma Media—J. R. Wait. (*J. Res. NBS*, vol. 65B, pp. 137–150; April-June, 1961.) A plasma, consisting of a neutral mixture of electrons, ions and molecules, in the presence of a constant magnetic field H_0, possesses a dielectric constant which is in the form of a tensor. Exact solutions of boundary value problems involving such media are obtained for two-dimensional configurations. Explicit results are given for the reflection coefficients of stratified plasma in planar and cylindrical geometry.

537.56:538.561 4069

Radiation in a Plasma: Part 1—Cherenkov Effect—M. H. Cohen. (*Phys. Rev.*, vol. 123, pp. 711–721; August 1, 1961.) A linear continuum theory in which the em and electroacoustic waves set up in a plasma by arbitrary charge and current sources are found from Maxwell's equations. The results are similar to those of Pines and Bohm (1021 of 1954), although the latter start from the particle viewpoint.

538.566 4070

Contribution to the Theory of Mode Conversion—F. Müller. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 20–25; February, 1961.) Maxwell's equations are formulated for three-dimensional Riemann space. (See also 1596 of May.) The propagation of a Lecher wave without mode conversion is considered.

538.566:535.42 4071

Diffraction of a Plane Electromagnetic Wave at a Circular Ferrite Cylinder—W. Eggimann. (*Bull. schweiz. elektrotech. Ver.*, vol. 52, pp. 377–381; May 20, 1961.) German version of paper published in IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-8, pp. 440–445; July, 1960.

538.566:535.43 4072

Scattering of Electromagnetic Waves from Concentric Infinite Cylinders—M. Kerker and E. Matijević. (*J. Opt. Soc. Amer.*, vol. 51, pp. 506–508; May, 1961.) "The solution for the scattering of radiant energy in the form of electromagnetic waves by concentric isotropic infinitely long circular cylinders is given for the incident energy travelling perpendicular to the cylinder axis."

538.569.4 4073

Theory of Resonance Absorption of Energy by a Rotating Solid.—J. Dreitlein and H. Kessemeier. (*Phys. Rev.*, vol. 123, pp. 835–852; August 1, 1961.)

538.569.4:621.375.9:535.61–1/2 4074

Use of a Different Type of Fluorescent Material as an Optical Maser—J. J. Markham and D. Mergerian. (*Nature*, vol. 191, pp. 586–588; August 5, 1961.) A device is suggested which is based on the Stokes' shift associated with absorption and emission bands created by point defects in solids such as KCl. Preliminary calculations indicate that such a device can be made to operate.

538.569.4:621.375.9:535.61–2 4075

Generation of Optical Harmonics—P. A. Franken, A. E. Hill, C. W. Peters and G. Weinreich. (*Phys. Rev. Lett.*, vol. 7, pp. 118–119; August 15, 1961.) Analysis is given together with a description of experimental observation of the second harmonic produced by projecting an intense beam of 6943-Å light through crystalline quartz.

538.569.4:621.375.9:535.61–2 4076

Stimulated Optical Emission in Fluorescent Solids: Part I—Theoretical Considerations—T. H. Maiman. (*Phys. Rev.*, vol. 123, pp. 1145–1150; August 15, 1961.)

538.569.4:621.375.9:535.61–2 4077

Stimulated Optical Emission in Fluorescent Solids: Part 2—Spectroscopy and Stimulated Emission in Ruby—T. H. Maiman, R. H. Hoskins, I. J. D'Haenens, C. K. Asawa and V. Evtuhov. (*Phys. Rev.*, vol. 123, pp. 1151–1157; August 15, 1961.)

538.569.4:621.375.9:535.61–2 4078

Fluorescence and Optical Maser Effects in $CaF_2:Sm^{++}$—W. Kaiser, C. G. B. Garrett and D. L. Wood. (*Phys. Rev.*, vol. 123, pp. 766–776; August 1, 1961.)

538.691 4079

Electron Trajectories in a Nonuniform Axially Symmetric Magnetic Field—D. A. Dunn and R. E. Holaday. (*J. Appl. Phys.*, vol. 32, pp. 1612–1620; August, 1961.) The equations for motions of electrons in a sinusoidally varying magnetic field have been solved. A device for converging a hollow cylindrical electron beam designed and built on the basis of the solutions, gave measured area convergences of 15 to 1.

539.2:537.311.1 4080

Energy Bands in the Presence of an External Force Field: Anomalous Velocities—E. N. Adams and E. I. Blount. (*J. Phys. Chem. Solids*, vol. 10, pp. 286–303; August, 1959.) The theory of field-modified energy bands is extended to include the effect of weak scattering forces. The modified current operator is found to contain terms giving anomalous currents. The theory of these currents is developed and their relation to the spontaneous Hall effect in ferromagnets is considered.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164:523.165 4081

Spectrum of Galactic Radio Emission—F. G. Smith. (*Nature*, vol. 191, p. 1381; September 30, 1961.) A discrepancy between theoretical and experimental values of α, the radio spectral index, is discussed. New observations at 408 Mc are consistent with $\alpha = 0.37$, the value previously obtained.

523.164:551.594.5 4082

Radio-Star Scintillations and the Auroral Zone—P. A. Forsyth and K. V. Paulson. (*Canad. J. Phys.*, vol. 39, pp. 502–509; April, 1961.) Scintillations at 53 Mc over a four-year period are analyzed. The dependence on the altitude of the source was much stronger in 1955 than in 1958. The scintillations probably arise in regions of the atmosphere which migrate southwards when solar activity is high.

523.164.3 4083

Radio Spectrum of Jupiter—A. G. Smith. (*Science*, vol. 134, pp. 587–595; September 1, 1961.) Review of the results of radio observations from 1955 to the present. 35 references.

523.164.32 4084

The Solar Noise Burst of November 12, 1960—A. E. Covington, G. A. Harvey and L. R. McNarry. (*Canad. J. Phys.*, vol. 39, pp. 635–636; April, 1961. Plate). A description of the noise received at 2800 and 48 Mc.

523.164.32 4085

Spectral Characteristics of Type II Solar Radio Bursts—P. A. Sturrock. (*Nature*, vol.

192, p. 58; October, 7, 1961.) An explanation of a) the appearance of radiation only at fundamental and second harmonic frequencies and b) the frequent splitting observed in these lines is given. See also Roberts (3816 of 1960).

523.164.32 **4086**
Decimetre Radio Bursts Concurrent with Solar Type IV Radiation—T. Krishnan and R. F. Mullaly. (*Nature*, vol. 192, pp. 58–59; October 7, 1961.) Measurements taken during a Type IV burst show that radiation at 1420 Mc can originate in a part of the sun's atmosphere widely different from the source of 45–60-Mc radiation.

523.164.32:523.75 **4087**
Coincidence of the Explosive Phase of Solar Flares with 10.7 cm Solar Noise Bursts—A. E. Covington and G. A. Harvey. (*Nature*, vol. 192, pp. 152–153; October, 1961.) Comparison of commencement times of 10.7-cm solar noise bursts with those of the observed explosive phase of solar flares shows good correlation.

523.164.4 **4088**
Amplitude and Angular Scintillations of the Radio Source Cygnus-A Observed at Boulder, Colorado—R. S. Lawrence, J. L. Jesperson and R. C. Lamb. (*J. Res. NBS*, vol. 65D, pp. 333–350; July/August, 1961.) Comparison with records from a suitably-placed ionosonde shows correlation between scintillations and spread-*F*, but not with E_s. The angular dependence of scintillations shows that the ionospheric irregularities are anisotropic. During the day slow, irregular variations of angle occur, and are attributed to large-scale lens-like irregularities.

523.165 **4089**
Intensity Variations in Outer Van Allen Radiation Belt—S. E. Forbush, D. Venkatesan and C. E. McIlwain. (*J. Geophys. Res.*, vol. 66, pp. 2275–2287; August, 1961.) Count rates observed with Explorer VII during stormy periods (October–December, 1959) indicate a decrease outside 3.5 earth radii during the main phases. The lower cutoff values on the counters are 18 mev for protons, 1 mev for electrons. The proportion of soft radiation increases during periods of increased intensity.

523.165 **4090**
Some Effects of Diamagnetic Ring Currents on Van Allen Radiation—A. J. Dessler and R. Karplus. (*J. Geophys. Res.*, vol. 66, pp. 2289–2295; August, 1961.) A qualitative theoretical model is proposed to explain the depletion of the outer Van Allen belt during the main phase of a storm. The theory incorporates a magnetic tail on the dark side of earth, a ring current centered at 4–6 earth radii and local betatron acceleration and deceleration of Van Allen and solar particles.

523.165 **4091**
Energy Spectrum and Angular Distributions of Electrons Trapped in the Geomagnetic Field—J. B. Cladis, L. F. Chase, Jr., W. L. Imhof and D. J. Knecht. (*J. Geophys. Res.*, vol. 66, pp. 2297–2312; August, 1961.) Count rates taken during a sounding rocket flight to 1045 km are used to derive distributions of pitch angles at various heights and the energy spectrum. The latter was found to be constant above 600 km. Electron flux for energies >50 kev at apogee was $(4.2\pm0.8)\times10^6$ cm^{-2} sec^{-1}.

523.165 **4092**
The Observed Outer-Belt Electron Distribution and the Neutron Decay Hypothesis—W. N. Hess, J. Killeen, C. Y. Fan, P. Meyer and J. A. Simpson. (*J. Geophys. Res.*, vol. 66, pp. 2313–2314; August, 1961.) The electron flux distribution along a line of force of range 21,000 km in the equatorial plane is compared with a curve based on the neutron albedo theory of electron production. There is sufficient disparity between the two curves to suggest that other sources of electrons must exist.

523.165 **4093**
Unusual Increase in the Cosmic-Ray Intensity on November 20, 1960—M. A. Pomerantz and S. P. Duggal. (*J. Franklin Inst.*, vol. 271, pp. 327–335; April. 1961.) Observations at arctic and antarctic stations show a statistically significant anisotropy.

523.3:621.396.96 **4094**
Radar Scattering Cross-Section—Applied to Moon Return—Hayre. (See 4155.)

523.5:621.396.96 **4095**
The Latitude Dependence of Radar Meteor Shower Observations—C. S. L. Keay and C. Ellyett. (*J. Geophys. Res.*, vol. 66, pp. 2337–2343; August, 1961.) A simplified method of assessing the relative significance of meteor showers is described which leads to good agreement between surveys in both hemispheres.

523.5:621.396.96 **4096**
Head Echoes from Meteor Trails—J. S. Greenhow. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 64–74; September, 1961.) Head echoes are often reflected from regions up to 100 km in length which may persist for many seconds after the passage of the meteor. The consequences of this fact are discussed.

523.72:523.165 **4097**
Solar X Rays—R. W. Kreplin. (*Ann. Géophys.*, vol. 17, pp. 151–161; April-June, 1961. In English.) Measurements of the intensity of solar X rays made during the past eleven years using rockets are reviewed and a useful table showing intensities as a function of wavelength and solar activity is given.

523.72:551.510.535 **4098**
Preliminary Data on Solar Extreme Ultraviolet Radiation in the Upper Atmosphere—H. E. Hinteregger. (*J. Geophys. Res.*, vol. 66, pp. 2367–2380; August, 1961.) A rocket-borne monochromator was flown in August, 1960. The intensity of the solar spectrum from 260–1300 Å was scanned in 10 s at a series of heights up to 235 km. Tables of the intensities, extrapolated to a height above the earth's atmosphere, are given together with absorption cross sections and other useful information. For completeness, similar tables are given for wavelengths outside the range covered by the August measurements.

523.745 **4099**
Secular Variations of Short-Lived Sunspots—T. S. Ringnes. (*Nature*, vol. 192, pp. 151–152; October 14, 1961.

550.38 **4100**
Geomagnetism in Relation to Aeronomy—E. H. Vestine. (*Ann. Géophys.*, vol. 17, pp. 181–194; April-June, 1961; In English.) A brief review dealing with quiet-day and storm variations and associated current systems, the motion of particles trapped in the radiation belts, and phenomena in the outer atmosphere. 57 references.

550.385.4 **4101**
Geomagnetic Variations between November 12 and November 16, 1960—E. R. Niblett. (*Canad. J. Phys.*, vol. 39, pp. 619–622; April, 1961.)

550.385.4 **4102**
On the Reflection and Refraction of Hydromagnetic Waves in Ionized Gas—T. Namikawa. (*J. Geomag. Geoelec.*, vol. 12, no. 3, pp. 117–128; 1961.) Reflection and refraction of hydromagnetic waves at a plane boundary between two semi-infinite homogeneous media of anisotropic electrical conductivity are considered for the case of a steady uniform magnetic field perpendicular to the plane. Difficulties in obtaining solutions when the field is at other angles are demonstrated.

550.385.4:551.594 **4103**
Electric Currents associated with Geomagnetic Disturbances in the Auroral Zone—F. J. Scrase. (*Nature*, vol. 191, pp. 963–966; September, 1961.) Currents of the order of 10^6 a have been observed at Lerwick during bay-type disturbances. A west-to-east maximum occurred between 1700 and 1800 GMT, zero current at 2100 and an east-to-west peak was measured at about midnight. The west-to-east current was at the level of the *F* layer and the east-to-west current near E-layer level.

550.386.6 **4104**
The Diurnal Variation of *K* Indices of Geomagnetic Activity on Disturbed Days in 1940–1948—S. B. Nichsolson and O. R. Wulf. (*J. Geophys. Res.*, vol. 66, pp. 2399–2404; August, 1961.)

551.507.362.2 **4105**
Analysis of Satellite Data—D. G. Parkyn. (*Nature*, vol. 191, pp. 896–897; August 26, 1961.) From an analysis of the orbit of satellite 1958 δ_2 atmospheric density variations have been isolated which are tentatively ascribed to third-harmonic tides.

551.507.362.2 **4106**
Errors in Orbital Predictions for Artificial Satellites of Earth—K. Moe. (*Nature*, vol. 192, p. 151; October 14, 1961.)

551.507.362.2 **4107**
Atmospheric Densities at Heights of 180–700 km—D. G. King-Hele and D. M. C. Walker. (*Ann. Géophys.*, vol. 17, pp. 162–171; April-June, 1961. In English.) Atmospheric densities determined from changes in the orbits of 22 earth satellites are discussed.

551.507.362.2 **4108**
Determination of Atmospheric Densities from Satellite Observations—C. A. Whitney. (*Ann. Géophys.*, vol. 17, pp. 237–244; April-June, 1961. In English.) A discussion of the errors arising from various factors involved in the determination of upper-atmosphere densities from satellite drag measurements.

551.507.362.2:061.3 **4109**
Proceedings of the Fourth AFBMD/STL Symposium: Advances in Ballistic Missile and Space Technology—(*Planet-Space Sci.*, vol. 7, pp. 1–462; July, 1961.) The text is given of 45 papers on various aspects of rocket and satellite techniques including instrumentation, space guidance and computers. For papers on space communications, see *e.g.*, 4298 below.

551.507.362.2:523.164 **4110**
Proposed Satellite Techniques for Performing a High-Resolution Survey of the Radio Sky at Medium Wavelengths—R. C. Jennison. (*J. Brit. IRE*, vol. 22, pp. 205–208; September, 1961.) The reception pattern of a short dipole above the *F* layer is modified by refraction in the ionized medium. The properties of this pattern are discussed and satellite experiments using this phenomenon are suggested.

551.507.362.2:523.165 **4111**
Satellite-Borne Instrumentation for Observing Flux of Heavy Primary Cosmic Radiation—P. Schwed, M. A. Pomerantz, H. Hanson and H. Benjamin. (*J. Franklin Inst.*, vol. 271, pp. 275–291; April, 1961.)

551.507.362.2:523.165 **4112**
Cosmic-Ray Measurements in the U. K.

Scout I Satellite—H. Elliot, J. J. Quenby, D. W. Mayne and A. C. Durney. (*J. Brit. IRE*, vol. 22, pp. 251–256; September, 1961.) The main purpose is to measure accurately the primary cosmic-ray energy spectrum and the changes in it due to modulation by the interplanetary magnetic field. G-M counters and a Cherenkov counter will be used.

551.507.362.2:621.38:539.12.04 **4113**

Radiation and other Environmental Effects on Satellites—R. Innes. (*J. Brit. IRE*, vol. 22, pp. 241–250; September, 1961.) Transistor failure in a satellite spending an appreciable part of its life in the inner Van Allen belt seems likely after several years. Solar cells cannot be effectively protected from hard radiation.

551.510.535 **4114**

Lyman-α Radiation—H. Friedman. (*Ann. Géophys.*, vol. 17, pp. 245–248; April-June, 1961. In English.) The role of Lyman α in the ionization of the *D* region is discussed. Nocturnal observations of Lyman α which lead to the concept of the hydrogen geocorona are reviewed.

551.510.535 **4115**

Theory of Electrostatic Fields in the Ionosphere at Equatorial Latitudes—J. R. Spreiter and B. R. Briggs. (*J. Geophys. Res.*, vol. 66, pp. 2345–2354; August, 1961.) An analysis is made of the elongated electrostatic fields required to provide the *E*- and *F*-layer coupling in the dynamo theory, for equatorial conditions. The attenuation of the electrostatic field is found to be large, implying very small coupling, for fields with horizontal wavelengths of a few km; most of this attenuation occurs at heights near the *E* region. (For an analysis in non-equatorial regions, see 3393 of October.)

551.510.535 **4116**

Apparent Reflection Heights and Reflection Properties of the Ionosphere above Tsumeb, South-West Africa, determined by means of Pulses in the Frequency Band from 350 to 5600 kc/s—W. Elling. (*Arch. elekt. Übertragung*, vol. 15, pp. 115–124; March, 1961.) The results of nighttime ionospheric measurements made during the period August, 1957 to March, 1958 are analyzed to determine the electron density distribution below the *F* layer. At least three ledges to the electron distribution were regularly observed at heights between 110 and 170 km and in the lower *E* region. The occurrence of *E*. layers was also noted.

551.510.535 **4117**

Comparison of Absorption Coefficients of the Ionosphere from Pulse Amplitude Measurements and Field-Strength Recordings—H. Schwentek and G. Umlauft. (*Arch. elekt. Übertragung*, vol. 15, pp. 200–204; April, 1961.) Ionospheric absorption was measured by two methods: a) vertical-incidence soundings at 1.73 Mc; b) field-strength measurements over an oblique-incidence path 295 km in length, at 2.61 Mc. These frequencies are such that, in both methods, the heights of reflection are equal. The absorption coefficients determined using the two methods have a correlation coefficient of 0.84 for both daily and monthly mean values, and are related to each other by the square of the effective frequency $(f+f_L)$.

551.510.535 **4118**

The Ionospheric Disturbance of November 12 and 16, 1960—W. S. Campbell and P. L. Hubert. (*Canad. J. Phys.*, vol. 39, pp. 614–616; April 1961.) Changes in ionospheric absorption as shown by *f* min are discussed.

551.510.535 **4119**

Nocturnal and Seasonal Variations of Equatorial Spread-*F*—M. S. V. G. Rao and B. R. Rao. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 12–22; September, 1961.) A classification of spread-echo patterns based on the distribution of relative amplitude within the echo is described. The variations in spread-*F* index at Waltair (7.4°N geomag.) are compared with those at Kodaikanal (0.6°N geomag.)

551.510.535:523.164 **4120**

Measurement of Ionospheric Drift by Radio-Star Observations—B. Dueño. (*J. Geophys. Res.*, vol. 66, pp. 2355–2365; August, 1961.) The incidence of scintillations is greatest in the hours before sunrise at Puerto Rico; it is not correlated with spread-*F* or E_s. The drift results show that the ionosphere irregularities causing scintillation are elongated along the lines of force of the geomagnetic field, and that drift, rather than random changes, is the main source of scintillation.

551.510.535:523.164 **4121**

A Method of Estimating Ionospheric Absorption Heights from Recovery Times—S. Ziauddin. (*Nature*, vol. 191, pp. 1084–1085; September 9, 1961.) Rapid recovery from absorption occurs when the ionizing source is suddenly cut off. From a plot of ionospheric relaxation time against height, an estimate can be made of the height of the absorbing layer which is in good agreement with measurements made by the three-frequency riometer method.

551.510.535:523.164 **4122**

Ionospheric Absorption on November 12, 1960—E. L. Vogan and T. R. Hartz. (*Canad. J. Phys.*, vol. 39, pp. 630–635; April, 1961.) The results obtained using riometers in Canada are described.

551.510.535:523.745 **4123**

Ionospheric Absorption and Solar Activity in Intermediate Latitudes—M. Anastassiades, D. Ilias and C. Caroumbalos. (*Ann. Géophys.*, vol. 17, pp. 256–260; April-June, 1961.) Measurements of the absorption of cosmic noise were made at Athens on 27.6 and 58 Mc. Analysis of the data allowed the components of the total absorption due to the *F* and *D* layers and the extra component to be estimated.

551.510.535:523.75 **4124**

Ionospheric Current Systems of Ten I.G.Y. Solar-Flare Effects—D. Van Sabben. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 32–42; September, 1961.) The current systems consist of two vortices, one on each side of the magnetic equator. The northern vortex is centered west of the southern one and is more intense in local summer than the southern one. The asymmetry in the vortices can be described in terms of an ionospheric current system.

551.510.535:550.385.2 **4125**

Ionospheric Critical Frequencies and Magnetic Parameters of the Day before—H. W. Bergh and J. W. King. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 74–75; September, 1961.) The correlation between f_0F_2 at 0300 and 0500 LMT at Capetown and local magnetic activity *K* was calculated. The correlation is greatest when the magnetic indexes for the previous day are used. This is true both when 24-h and 6-h means of *K* are used in the correlations.

551.510.535:550.385.37 **4126**

Connexion between Micropulsations and the Ionosphere—C. M. Lock and P. J. Stevens. (*Nature*, vol. 191, pp. 584–585; August 5, 1961.) Measurements made at Christchurch, England, and Ascension Island, 6750 km distant, show that P_t pulses occur simultaneously at both stations but the same coherence is not observed for P_c pulsations.

551.510.535:551.507.362.1 **4127**

Atmospheric Ion Compositions Measured above Wallops Island, Virginia—H. A. Taylor, Jr., and H. C. Brinton. (*J. Geophys. Res.*, vol. 66, pp. 2587–2588; August, 1961.) Preliminary data for heights between 100 and 250 km, obtained from two rocket flights, are given and compared with results of previous determinations.

551.510.535:551.507.362.2 **4128**

Ionospheric Information Deduced from the Doppler Shifts of Harmonic Frequencies from Earth Satellites—O. K. Garriott and A. W. Nichol. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 50–63; September, 1961.) Measurements made near the end of the life of Sputnik III have indicated the presence of appreciable ionization at a height of 90 km in the daytime and have revealed the existence of ionospheric irregularities below 175 km.

551.510.535:551.507.362.2 **4129**

Height of Nighttime *F*-Layer Irregularities at the Equator—G. S. Kent and J. R. Koster. (*Nature*, vol. 191, pp. 1083–1084; September 9, 1961.) Satellite transmissions propagated through the irregularities produce a diffraction pattern from whose velocity the effective height of the diffracting screen may be determined. Irregularities are elongated with axial ratio at least 7:1 along the earth's magnetic field and lie between 50 and 100 km above the base of the *F* layer.

551.510.535:621.3.087.4 **4130**

The Determination of the Absorption of the Ionosphere from Pulse Amplitude Observations—G. Umlauft. (*Arch. tech. Messen*, No. 302, pp. 53–56; March, 1961.) Equipment for measuring ionospheric absorption by means of vertical-incidence soundings is described. It operates at 1.73 or 2.23 Mc with pulse width variable between 50 and 500 μsec. Details are given of the polarimeter receiving circuit. See also 4117 above.

551.510.535:621.391.812.63 **4131**

On the Validity of some Approximations to the Appleton-Hartree Formula—Davies and King. (See 4277.)

551.510.535:621.391.812.63 **4132**

Investigation of Ionospheric Absorption on 5.65 Mc/s at Waltair—Ramana and Rao. (See 4278.)

551.510.535:621.391.812.63 **4133**

The Fading of Radio Waves Reflected Vertically from the Ionosphere during Magnetic Storms—G. F. Fooks. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 43–49; September, 1961.) Correlation analysis is applied to the spread, rapidly-fading reflection observed at Cambridge (52°N) during magnetic storms. The drift velocity is usually large and towards the West, and the direction of elongation of the diffraction pattern shows that the ionospheric irregularities are aligned with the earth's magnetic field.

551.510.535:621.396.676 **4134**

The Electrically Short Antenna as a Probe for Measuring Free Electron Densities and Collision Frequencies in an Ionized Region—R. King, C. W. Harrison, Jr., and D. H. Denton, Jr. (*J. Res. NBS*, vol. 65D, pp. 371–384; July/August, 1961.) The integral-equation method is used to obtain expressions for the admittance of an electrically short center-driven dipole or base-driven monopole when immersed in a conducting dielectric. The conductivity and permittivity of the medium can be calculated from measurements of the admittance antenna immersed in the medium and in air.

551.510.535:681.142 **4135**

An Analogue Solution of the Continuity

Equation of the Ionospheric F Region—B. H. Briggs and H. Rishbeth. (*Proc. Phys. Soc.* (*London*), vol. 78, pp. 409–422; September 1, 1961.) An electromechanical analog computer is used to solve an idealized *F*-layer continuity equation. The results show how the layer above the F_2 peak is mainly controlled by vertical diffusion of ionization, and below the peak by the production and loss processes. The machine can also simulate the effect of a solar eclipse.

551.510.535(98):523.75 4136

Effects of Corpuscular Emissions on the Polar Ionosphere Following Solar Flares—G. E. Hill. (*J. Geophys. Res.*, vol. 66, pp. 2329–2335; August, 1961.) During the events of July 24–25, and September 18–22, 1957, ionospheric absorption was more intense in the sunlit portion of the polar cap. After the September event, there was strong auroral absorption, a geomagnetic storm and a decrease in F_2-layer ionization; none of these occurred after the July event.

551.510.535(99):621.391.812.63 4137

Ionospheric Reflections below 50 Kilometres during Polar-Cap Absorption Events—J. B. Gregory. (*J. Geophys. Res.*, vol. 66, pp. 2575–2577; August, 1961.) Vertical-incidence $h'(t)$ observations on 2·3 Mc at Scott Base (latitude 78°S) showed partial reflections from heights below 50 km during two polar-cap absorption events. The results are correlated with measurements made in space vehicles during the same periods.

551.510.536:551.594.6 4138

Electron Densities in the Outer Ionosphere Deduced from Nose Whistlers—R. L. Smith. (*J. Geophys. Res.*, vol. 66, pp. 2578–2579; August, 1961.) Comment on 1519 of May (Pope).

551.510.536:551.594.6 4139

Some Comments on Obtaining Electron Densities in the Exosphere using Nose Whistlers—J. H. Pope. (*J. Geophys., Res.*, vol. 66, pp. 2580–2581; August, 1961.) Discussion of a paper by Smith (see 4138 above).

551.593.5 4140

Variations of [OI] 5577-Å Emission in the Upper Atmosphere—F. E. Roach. (*Ann. Géophys.*, vol. 17, pp. 172–180; April-June, 1961. In English.) A review of experimental data, with consideration of excitation mechanisms and relations with other airglow emissions.

551.593.5 4141

A Simultaneous Study of the λλ 5577, 5893 and 6300 Emissions of the Night Airglow at Poona—M. W. Chiplonkar and V. V. Agashe. (*Ann. Géophys.*, vol. 17, pp. 231–236; April-June, 1961. In English.) Results of measurements made in early 1960.

551.594 4142

Hydrogen Emissions during the Period November 9–16, 1960—R. Montalbetti and D. J. McEwen. (*Canad. J. Phys.*, vol. 39, pp. 617–619; April, 1961.)

551.594:621.396.9 4143

Power Density Requirements for Airglow Excitation by Gyro-waves—V. A. Bailey. (*J. Res. NBS*, vol. 65D, pp. 321–322; July/August, 1961.) An array of 80 dipoles radiating a gyrowave with a mean power of 500 kw would produce an easily observable enhancement of the nocturnal airglow overhead within a solid angle of 32 square degrees.

551.594.5 4144

Temperatures Deduced from Aurora and Airglow Spectra—D. M. Hunten. (*Ann. Géophys.*, vol. 17, pp. 249–255; April-June, 1961. In English.) Several experimental results are combined to give a temperature profile between 70 and 250 km.

551.594.5 4145

Observations of Auroral Ionization during the Period November 11–16, 1960—H. V. Serson and B. C. Blevis. (*Canad. J. Phys.*, vol. 39, pp. 622–625; April, 1961.) Auroral luminosity and radar echo amplitude at 488 and 944 Mc were observed.

551.594.5:523.164 4146

Three-Frequency Measurements of Auroral Absorption—S. Ziauddin and P. A. Forsyth. (*J. Geophys. Res.*, vol. 66, pp. 2315–2319; August, 1961.) The ratio of the measured absorption of cosmic noise during auroral events was measured on two pairs of frequencies near 30 Mc. Assuming that the electron collision frequency as a function of height is known, the height at which the absorption occurs can be estimated. Heights of 50–70 km were usual for intense auroral events, but lower heights were observed for cosmic-ray events.

551.594.5:621.396.96 4147

Auroral Observations by Radar in Terre Adelie (Nov. 1957–Jan. 1958.)—K. Bullough. (*C.R. Acad. Sci.* (*Paris*), vol. 251, pp. 2222–2223; November 14, 1960.) Observations made during the southern summer show a correlation between the geomagnetic latitude of the echoes and the K_p value. For full report see *Ann. Géophys.*, vol. 17, pp. 195–230; April-June, 1961.

551.594.5:621.396.96 4148

Auroral Radar Observations at 48 Mc/s during the Period of the 12 November, 1960, Solar Event—A. G. McNamara. (*Canad. J. Phys.*, vol. 39, pp. 625–627; April, 1961.)

551.594.6 4149

Extension of Nose Whistler Analysis—R. L. Smith and D. L. Carpenter. (*J. Geophys. Res.*, vol. 66, pp. 2582–2586; August, 1961.) A method of analysis is described which gives the nose frequency and the time delay to the nose when the nose is not directly observable.

LOCATION AND AIDS TO NAVIGATION

534.88 4150

Consideration of the Problem of Optimum Frequency in Underwater Sonar Systems—H. Thiede. (*Acustica*, vol. 11, no. 1, pp. 22–25; 1961. In German.) The optimum frequencies of sonar systems for obtaining a given range are determined taking account of the frequency-dependent parameters of sound propagation in sea water. The relation between range and minimum areas of radiating surface is also calculated.

621.396.9:551.507.362.2 4151

Instrumentation and Telemetry of Transit Navigational Satellites—J. W. Hamblen and J. B. Oakes. (*Electronics*, vol. 34, pp. 148–153; August 11, 1961.) The system is based on the accurate positional information obtainable from the Doppler shift of the signal from a satellite. The experimental satellite described has four transmitters operating between 54 and 324 Mc, driven from a single-frequency source accurate to within a few parts in 10^{10}. Operational, constructional and circuit details are given.

621.396.9:629.19.053 4152

Inertial Systems in Space Vehicles—M. A. V. Matthews. (*J. Brit. IRE*, vol. 22, pp. 231–239; September, 1961.) A discussion of the advantages and disadvantages when used in conjunction with radio methods.

621.396.933.2:621.396.67 4153

The Application of the Interferometer to H.F. Direction Finding—C. W. McLeish and N. Burtnyk. (*Proc. IEE*, pt. B, vol. 108, pp. 495–499; September, 1961.) A considerable reduction in bearing error, compared with a narrow-aperture system, is obtained in the presence of interference. A standard deviation of 1.2° was obtained with a 400-ft interferometer on skywave transmission at 5–25 Mc.

621.396.933.2:621.396.67 4154

Antenna Coupling Error in Direction Finders—C. W. Harrison, Jr. (*J. Res. NBS*, vol. 65D, pp. 363–369; July/August, 1961.) "Antenna coupling or scattering error in an interferometer-angle-measuring system consisting of two identical base-loaded quarter-wave vertical antennas over a perfectly conducting earth is investigated. A curve is supplied for the error in angle as a function of antenna spacing for an incident vertically-polarized electric field arriving in the plane of the antennas, when the ratio of antenna length to radius is 75 and the base loads are resistors of 50 ohms."

621.396.96:523.3 4155

Radar Scattering Cross-Section—Applied to Moon Return—H. S. Hayre. (Proc. IRE, vol. 49, p. 1433; September, 1961.) A discussion of the effect of rough terrain on the scattered echo. Results indicate that the effective surface as seen by radar is not as smooth as was assumed by some previous authors.

621.396.96:621.374.5 4156

The Delay-Lock Discriminator—an Optimum Tracking Device—J. J. Spilker, Jr., and D. T. Magill. (Proc. IRE, vol. 49, pp. 1403–1416; September, 1961.) A statistically optimum device for the measurement of the delay between two correlated waveforms. A nonlinear feedback system including a variable delay line is used. The theory and operation are described.

621.396.96:621.391.82 4157

Search-Radar Analogue Simulator Reproduces Jamming, Scintillation—L. G. Fischer and G. Frenkel. (*Electronics*, vol. 34, pp. 58–60; August 25, 1961.) Simulated signals are injected at IF into a radar set.

621.396.962.3 4158

An Interpretation of 'Paired Echo Theory' for Time-Domain Distortion in Pulsed Systems, and, an Extension to the Radar 'Uncertainty Function'—J. V. Difranco and W. L. Rubin. (Proc. IRE, vol. 49, pp. 1432–1433; September, 1961.) A discussion of the application of Fourier analysis in determinations of distortion in pulse systems and degradation of radar performance.

621.396.962.33:621.376.23 4159

A Sequential Detection System for the Processing of Radar Returns—A. A. Galvin. (Proc. IRE, vol. 49, pp. 1417–1423; September, 1961.) A two-step filter detection system is described which offers an appreciable reduction in size and complexity of equipment without reduction in performance.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.583:621.3.032.14 4160

Ternary System Th-Ce-Al—J. H. N. van Vucht. (*Philips Res. Repts.*, vol. 16, pp. 1–40; February, 1961.) An investigation of a non-evaporating getter material.

535.215:546.48'221 4161

The Real Structure of CdS Single Crystals—W. Möhling and H. Thiel. (*Ann. Phys.* (*Lpz.*) vol. 7, pp. 381–386; March 29, 1961.) Crystal growth and defects were investigated.

535.215:546.48'221 **4162**

Absorption Spectra of Transition Ions in CdS Crystals—R. Pappalardo and R. E. Dietz. (*Phys. Rev.*, vol. 123, pp. 1188–1203; August 15, 1961.)

535.215:546.48'221 **4163**

Effects of Vacancies on Pulse Photoconduction in CdS Single Crystals—J. Shirafuji and Y. Inuishi. (*J. Phys. Soc. Japan*, vol. 16, p. 832; April, 1961.) Cd vacancies shorten electron lifetimes at low temperatures. S vacancies enhance the quantum efficiency and increase the lifetime.

535.215:546.48'221:534.283–8 **4164**

Photosensitive-Ultrasonic Properties of Cadmium Sulphide—H. D. Nine and R. Truell. (*Phys. Rev.*, vol. 123, pp. 799–803; August 1, 1961.) Ultrasonic attenuation is found to vary with light irradiation, some crystals showing a decrease and others an increase. The latter effect is explicable by piezoelectric coupling between conduction electrons and stress waves.

535.215:546.682'221 **4165**

Photoconductivity in Indium Sulphide Powders and Crystals—R. H. Bube and W. H. McCarroll. (*J. Phys. Chem., Solids*, vol. 10, pp. 333–335; August, 1959.)

535.376:546.47'221 **4166**

The Influence of Temperature on Field Enhancement and Field Quenching of Cathodoluminescence—H. Gobrecht, H. E. Gumlich and J. zum Bruch. (*Z. Phys.*, vol. 162, pp. 164–179; March, 17, 1961.) Further results of investigations on ZnS-Mn. See also 590 of February (Gobrecht and Gumlich).

535.376:546.47'221 **4167**

Infrared Emission and Electroluminescence in Zinc Sulphide Phosphors—D. W. G. Ballentyne. (*Proc. Phys. Soc.* (*London*), vol. 78, pp. 348–353; September 1, 1961.) Experimental results show that infrared emission does not occur in materials with high Cu concentration, and was not detected during electroluminescence.

535.376:546.48'47'221 **4168**

Electrophotoluminescence of a Zinc-Cadmium-Sulphide Phosphor as a Function of Excitation Wavelength and Width of Field Pulse—H. Gutjahr and F. Matossi. (*Z. Phys.*, vol. 162, pp. 105–113; March 3, 1961.)

537.227:538.569.4 **4169**

Proton Magnetic Resonance in Ferroelectric $(CH_3NH_3)AlSO_2\cdot12H_2O$—R. Hoshino. (*J. Phys. Soc. Japan*, vol. 16, pp. 835–836; April, 1961.)

537.227:621.318.57 **4170**

Increase in Dielectric Constant during Switching in Lithium Selenite and Triglycine Sulphate—E. Fatuzzo (*J. Appl. Phys.*, vol. 32, pp. 1571–1579; August, 1961.) Changes in dielectric constant were measured during switching up to frequencies of 100 Mc. A model to explain the experimental results is given and discussed.

537.228.1:549.514.51 **4171**

Thickness-Shear-Mode Quartz Cut with Small Second- and Third-Order Temperature Coefficients of Frequency (RT-Cut)—R. Bechmann. (PROC. IRE, vol. 49, pp. 1454; September, 1961.)

537.311.31 **4172**

Theory of Electron-Phonon Interaction in Metals—H. Stolz. (*Ann. Phys.* (*Lpz.*), vol. 7, pp. 353–367; March 29, 1961.) Calculation of matrix elements for phonon absorption or emission on the basis of assumptions also discussed by Whitfield (2146 of July).

537.311.31 **4173**

Electron-Phonon Interaction—H. Puff. (*Ann. Phys.* (*Lpz.*), vol. 7, pp. 368–370; March 29, 1961.) See also 4172 above.

537.311.31 **4174**

Electrical Resistivity at Low Temperatures—J. S. Dugdale. (*Science*, vol. 134, pp. 77–86; July 14, 1961.) Investigations of the pressure dependence of the electrical resistance of metals at low temperatures are discussed; results are interpreted in terms of Fermi surface displacement and Fermi energy changes.

537.311.31:538.569.4 **4175**

Influence of Ferromagnetic Resonance on the Electrical Resistance of a Metal—A. Bassompierre. (*C.R. Acad. Sci.* (*Paris*), vol. 251, pp. 2141–2143; November 14, 1960.) An examination of the increase in resistance produced by the collision of conduction electrons with misaligned-spin faults caused by ferromagnetic resonance.

537.311.33 **4176**

The Phonon-Assisted Auger Effect in Semiconductors—D. M. Eagles. (*Proc. Phys. Soc.*, vol. 78, pp. 204–216; August 1, 1961.) Some calculations on the effect of phonons on Auger recombination and impact ionization in semiconductors with simple bands are presented.

537.311.33 **4177**

The Possibility of Negative-Resistance Effects in Semiconductors—B. K. Ridley and T. B. Watkins. (*Proc. Phys. Soc.*, vol. 78, pp. 293–304; August 1, 1961.) If carriers in a high-mobility subband are heated with an electric field, they can transfer to a higher-energy low-mobility subband. The necessary conditions and possible side effects are discussed.

537.311.33 **4178**

A Relation between the Binding Energy and the Band-Gap Energy in Semiconductors of Diamond or Zinc-Blende Structure—P. Manca. (*J. Phys. Chem. Solids*, vol. 20, pp. 268–273; August, 1961.) For the binary semiconductors of tetrahedral coordination considered, a linear relation of the type, $E_g=a(E_s-b)$ has been determined empirically where E_g is the gap energy, E_s is the single-bond energy and a and b are characteristic constants of the series A^{IV} A^{IV}, A^{III} B^{V} A^{II} B^{VI}.

537.311.33:546.23:537.226 **4179**

Dielectric Relaxation Phenomena in Selenium—W. Ludwig and F. Eckart. (*Z. Phys.*, vol. 162, pp. 127–141; March 17, 1961.) Measurements of the variation of dielectric loss in the range 0.5 – 1000 kc at temperatures from +40 to –160°C, on Br-doped hexagonal Se.

537.311.33:[546.28+546.289 **4180**

Theory of Electrical Conductivity of Ge and Si: Part 1—The Electron-Phonon Matrix Elements—W. Klose. (*Ann. Phys.* (*Lpz.*), vol. 7, pp. 233–242; March 16, 1961.) Intravalley and intervalley electron scattering at acoustic and optical phonons is determined using an approximation valid for low temperatures.

537.311.33:546.28 **4181**

Properties of Grain Boundaries in Silicon—Y. Matukura. (*J. Phys. Soc. Japan*, vol. 16, pp. 842–843; April, 1961.) The grain boundary layer in Si is n type.

537.311.33:546.28 **4182**

Diffusion along Small-Angle Grain Boundaries in Silicon—H. J. Gueisser, K. Hubner and W. Shockley. (*Phys. Rev.*, vol. 123, pp. 1245–1254; August 15, 1961.)

537.311.33:546.28:541.135 **4183**

Saturation Currents in Silicon Electrodes—G. Greger. (*Z. Naturforsch.*, vol. 16a, pp. 284–288; March, 1961.) Investigation on p- and n-type electrodes in various electrolytes. See also 3930 of 1960 (Seipt).

537.311.33:546.281'26 **4184**

Lattice Absorption Bands in SiC—L. Patrick and W. J. Choyke. (*Phys. Rev.*, vol. 123, pp. 813–815; August 1, 1961.)

537.311.33:546.289 **4185**

Surface Kinetics and Physics Investigation of the Reaction between Single-Crystal Germanium and Iodine—W. J. Heinecke and S. Ing, Jr. (*J. Appl. Phys.*, vol. 32, pp. 1498–1504; August, 1961.) An activation energy of 18.5 kcal/mol was found for each of the three crystallographic planes studied. Frequency factors differed with orientation.

537.311.33:546.289 **4186**

The Adsorption of Chlorine on Clean Germanium Surfaces—M. Green and K. H. Maxwell. (*J. Phys. Chem. Solids*, vol. 20, pp. 274–280; August, 1961.)

537.311.33:546.289 **4187**

Cyclotron-Resonance Measurements of the Energy-Band Parameters of Germanium—B. W. Levinger and D. R. Frankl. (*J. Phys. Chem. Solids*, vol. 20, pp. 281–288; August, 1961.)

537.311.33:546.289 **4188**

Conductivity of Grain Boundaries in Ge—Y. Matukura and S. Tanaka. (*J. Phys. Soc. Japan*, vol. 16 pp. 833–834; April, 1961.) The temperature variations of resistivity and Hall coefficient are measured.

537.311.33:546.289 **4189**

Electrical Properties of Heavily Doped n-Type Germanium—Y. Furukawa. (*J. Phys. Soc. Japan*, vol. 16, pp. 687–694; April, 1961.) Experimental studies were made on electrical resistivity, Hall coefficient and magnetoresistance of As-, Sb- and (As+Sb)-doped Ge in the electron concentration range $10^{18}-3\times10^{19}$ cm^{-3}.

537.311.33:546.289 **4190**

Regrowth of Germanium from Molten Indium on Lead—M. Tomono. (*J. Phys. Soc. Japan*, vol. 16, pp. 711–719; April 1961.) A single crystal of n-type Ge was alloyed with In and the resulting regrowth of p-type Ge was observed for various cooling rates.

537.311.33:546.289 **4191**

Low-Temperature Electrical Breakdown in Germanium—J. Yamashita. (*J. Phys. Soc. Japan*, vol. 16, pp. 720–732; April, 1961.) The breakdown field estimated theoretically is in agreement with experiment when the concentration of donors and acceptors is fairly small.

537.311.33:546.57'86'241 **4192**

The Thermal Conductivity of Silver Antimony Telluride—E. F. Hockings. (*J. Phys. Chem. Solids*, vol. 10, pp. 341–342; August, 1959.)

537.311.33:546.68'86 **4193**

Temperature Variation of Optical Energy Gap for GaSb—InSb Alloys—J. C. Woolley and J. A. Evans. (*Proc. Phys. Soc.*, vol. 78, pp. 354–360; September 1, 1961.) Absorption measurements show that the energy gap varies linearly with temperature in the range 100–800°K.

537.311.33:546.682'18 **4194**

Electrical Properties of p-Type InP—M. Glicksman and K. Weiser. (*J. Phys. Chem.*

Solids, vol. 10, pp. 337–340; August, 1959.) A plot of the temperature dependence of the measured Hall mobilities of two Cd-doped samples is given, and the properties of two crystals are tabulated.

537.311.33:546.682'86 4195

The Diffusion Coefficient of Zinc in Indium Antimonide—K. F. Hulme and J. E. Kemp. (*J. Phys. Chem. Solids*, vol. 10, pp. 335–337; August, 1959.) A value of $1.6\times10^6 \exp[(-53{,}000\pm6000)/RT]$ cm^2/sec was obtained using the *pn* junction method.

537.311.33:546.682'86 4196

Interpretation of the Anisotropy of the Resistance Change in *n*-Type Insb—H. Rupprecht. (*Z. Naturforsch.*, vol. 16a, pp. 395–401; April, 1961.) Further investigations of resistivity changes in InSb as a function of growth conditions and doping material. See also 1241 of April (Rupprecht, *et al.*).

537.311.33:546.683'86:539.12.04 4197

Recovery of Electron Radiation Damage in *n*-Type InSb—F. M. Eisen. (*Phys. Rev.*, vol. 123, pp. 736–744; August 1, 1961.)

537.311.33:[546.682'86+546.682'19]:539.23 4198

Properties of Vapour-Deposited InSb and InAs Films—K. G. Günther and H. Freller. (*Z. Naturforsch.*, vol. 16a, pp. 279–283; March, 1961.) Further report on results obtained with films prepared by the "three-temperature methods." See also 3349 of 1959 (Günther).

537.311.33:546.812.221 4199

Measurement and Analysis of the Infrared Reflection Saectrum of Semiconducting SnS—C. Haas and M. M. G. Corbey. (*J. Phys. Chem. Solids*, vol. 20, pp. 197–203; August, 1961.)

537.311.33:546.823–31 4200

Electrical Conduction in *p*-Type Titanium Sesquioxide—J. Yahia and H. P. R. Frederikse. (*Phys. Rev.*, vol. 123, pp. 1257–1261; August 15, 1961.) Measurements indicate that below 450°K the crystals behave like semiconductors while above it the conductivity is apparently metallic.

537.311.33:621.382.23 4201

The Band Structure of Strongly Doped Semiconductors—G. Winstel and W. Heywang. (*Z. Naturforsch.*, vol. 16a, pp. 440–441; April, 1961.) Theoretical consideration of the discrepancies arising between the conventional band model of doped semiconductors and the measured characteristics of tunnel diodes. See also 4329 below.

537.312.62 4202

A Study of Superconducting Niobium by Electron Tunnelling—P. Townsend and J. Sutton. (*Proc. Phys. Soc.*, vol. 78, pp. 309–311; August 1, 1961.)

537.322.1 4203

Peltier Coefficient at High Current Levels—J. R. Reitz. (*J. Appl. Phys.*, vol. 32, pp. 1623–1624; August, 1961.) The advantages of a metal weld at the junction of a thermocouple are considered. At high current densities the effect of using a two-metal weld is discussed.

538.221 4204

Experimental Test of the Spin-Wave Theory of a Ferromagnet—A. C. Gossard, V. Jaccarino and J. P. Remeika. (*Phys. Rev. Lett.*, vol. 7, pp. 122–124; August 15, 1961.)

538.221 4205

The Internal Stress Energy of Bloch Walls—M. Hieke, E. Max and H. G. Müller. (*Z. angew. Phys.*, vol. 13, pp. 125–130; March, 1961.) The internal stress energy is determined on the basis of experimental investigations of Bloch wall shape and compared with the eddy-current energy component in magnetization reversal processes.

538.221 4206

Fine Structure of Weiss Domains on {110} Surfaces of Si-Fe Single Crystals as a Function of Magnetic Field Strength and Field Direction—R. Kohlhaas and G. Ballensiefen. (*Z. angew. Phys.*, vol. 13, pp. 131–134; March, 1961.)

538.221 4207

Calculation and Measurement of the Internal Demagnetization Factor—G. Vogler. (*Ann. Phys. (Lpz.)*, vol. 7, pp. 268–279; March 16, 1961.) Magnetic energy and internal demagnetization factors are calculated for various types of distribution of small ferromagnetic particles in a given volume. Results of measurements on rubber-iron mixtures are compared with theoretical findings.

538.221 4208

The Diffusion After-Effect of Permeability of Iron-Nickel Alloys—D. Gerstner and E. Kneller. (*Z. Metallkde*, vol. 52, pp. 426–436; June, 1961.) Investigations were made on the change of permeability in quenched Fe-Ni alloys with 50 to 85 per cent Ni content by weight, at temperatures between 300°C and the Curie point.

538.221:537.288.4 4209

Magnification of the Magneto-optical Kerr Rotation by Interference—J. Kranz and A. Schauer. (*Optik (Stuttgart)*, vol. 18, pp. 186–200; April, 1961.) The rotation can be magnified by means of suitable vapor-deposited films, and this effect is calculated. The method has been used for the observation of Weiss domains. See also 1281 of 1959 (Kranz & Drechsel).

538.221:538.23 4210

Investigations of Rotational Hysteresis—V. Zehler. (*Z. angew. Phys.*, vol. 13, pp. 139–141, March, 1961.)

538.221:538.632 4211

Measurements of Hall Effect in Binary Mixed-Crystal Series of Iron—W. Jellinghaus and M. P. de Andrés. (*Z. angew. Phys.*, vol. 13, pp. 121–122; March, 1961.) See also 3481 of October.

538.221:538.632 4212

Extraordinary-Hall-Effect Measurements on Ni, some Ni Alloys, and Ferrites—J. M. Lavine. (*Phys. Rev.*, vol. 123, pp. 1273–1277; August 15, 1961.)

538.221:538.652 4213

The Change in Length of Mixtures of Ferromagnetic and Carrier Materials in a Magnetic Field—G. Vogler. (*Ann. Phys. (Lpz.)*, vol. 7, pp. 387–396; March 29, 1961.) Magnetostrictive effects are calculated and investigated by measurements on a suspension of iron powder in rubber. See also 4207 above.

538.221:538.652 4214

Volume Magnetostriction in Nickel-Iron Alloys—Y. Nakamura, T. Nakajima and H. Takaki. (*J. Phys. Soc. Japan*, vol. 16, pp. 840–841; April, 1961.) Magnetostriction was measured as a function of Ni/Fe ratio in ordered and disordered specimens.

538.221:538.652:534.232 4215

Improved Nickel-Base Alloys for Magnetostrictive Transducers—C. A. Clark. (*J. Acoust. Soc. Am.*, vol. 33, pp. 930–933; July, 1961.)

538.221:539.23 4216

Remanent State in One-Dimensional Micromagnetics—A. Aharoni. (*Phys. Rev.*, vol. 123, pp. 732–736; August 1, 1961.)

538.221:539.23 4217

The Theory of Magnetization Reversal in Thin Films—W. Andrä. (*Z. angew. Phys.*, vol. 13, pp. 141–143; March, 1961.) Inhomogeneous reversal processes are considered on the basis of domain theory.

538.221:539.23 4218

The Magnetic Domain Structure of 'Single-Crystal' Thin Films of Iron—D. Unangst. (*Ann. Phys. (Lpz.)*, vol. 7, pp. 280–301; March 16, 1961.) Bitter pattern technique is used to investigate the domain structure of films of thickness 500–10,000 Å during magnetic reversal and after demagnetization.

538.221:539.23 4219

The Magnetic Properties of Very Thin Vapour-Deposited Iron Films in a Vacuum and their Change after the Admission of Air—H. Hoffmann. (*Z. angew. Phys.*, vol. 13, pp. 149–151; March, 1961.)

538.221:539.23 4220

The Curie Temperature of Very Thin Nickel Films—W. Hellenthal. (*Z. angew. Phys.*, vol. 13, pp. 147–149; March, 1961.) Discrepancies in the results of Curie-temperature measurements on thin films by various authors are accounted for by reference to a model based on an investigation of slow magnetization processes (2311 of July) and calculation of the temperature dependence of spontaneous magnetization.

538.221:539.23 4221

The Influence of Local Fluctuations of Anisotropy on the Magnetic Characteristics of Thin Ni-Fe Films—S. Middelhoek. (*Z. angew. Phys.*, vol. 13, pp. 151–154; March, 1961.)

538.211:539.23 4222

The Magnetization Reversal of Thin Nickel-Iron Films in the Difficult Direction—Pt. 1: Electron-Microscope Observations of the Origin of Domains—E. Fuchs, pp. 157–160; Pt. 2: Hysteresis and Domain Behaviour—E. Feldtkeller. pp. 161–164; (*Z. angew. Phys.*, vol. 13; March, 1961.)

538.221:539.23 4223

The Temperature Dependence of the Magnetic Properties of Thin Permalloy Films—A. Segmüller. (*Z. angew. Phys.*, vol. 13, pp. 154–157; March, 1961.)

538.221:539.23:538.569.4 4224

Ferromagnetic Resonance Absorption of Single-Crystal Thin Films of Nickel—M. Kuriyama, H. Yamanouchi and S. Hosoya. (*J. Phys. Soc. Japan*, vol. 16, pp. 701–706; April, 1961.) The saturation magnetization, the crystalline anisotropy and the Landé splitting factor were measured for film thicknesses between 100 and 1000 Å.

538.221:539.23:538.61 4225

Measurement of Magnetization at the Surface of Iron-Silicon Single Crystals with the aid of the Magnified Magneto-optical Kerr Effect—J. Kranz and B. Passon. (*Z. angew. Phys.*, vol. 13, pp. 122–124; March, 1961.) See also 3879 of November.

538.221:539.23:621.385.833 4226

Electron-Optical Investigation of Domain Boundaries in Thin Ferromagnetic Films—L. Reimer. (*Z. angew. Phys.*, vol. 13, pp. 143–147; March, 1961.) Investigation by means of an electron microscope of Ni and Fe films using the technique described by Fuller and Hale (2085 of 1960).

538.221:539.38 4227

Deformation of Domain Structures by Stresses—C. Greiner. (*Z. angew. Phys.*, vol. 13, pp. 134–136; March, 1961.) Investigation of the effects of tension and compression on Bloch wall displacement in Ni-Fe (50/50) wire.

538.221:621.3.042.2 4228

The Dynamic Behaviour of Bloch Walls in Grain-Oriented Laminations—P. K. Hermann. (*Z angew. Phys.*, vol. 13, pp. 136–139; March, 1961.) An interpretation of the cause of hysteresis effects observed particularly in thin grain-oriented laminations is given.

538.221:621.318.124:538.569.4 4229

Ferrimagnetic Resonance of Single-Crystal Barium Ferrite in the Millimetre Wave Region—F. F. Y. Wang, K. Ishii and J. B. Y. Tsui. (*J. Appl. Phys.*, vol. 32, pp. 1621–1622; August, 1961.) Results of measurements at frequencies about 50 Gc agree with the theory developed by Weiss and Anderson (3326 of 1955).

538.221:621.318.134 4230

The Estimation of Domain-Wall Velocity in a 'Square-Loop' Ferrite, and some Observations on the Reversal Process—J. E. Knowles. (*Proc. Phys. Soc.*, vol. 78, pp. 233–238; August 1, 1961.)

538.221:621.318.134 4231

The Angle Dependence of the Anisotropy Induced in Iron-Nickel Ferrite Single Crystals by a Magnetic Field—R. Wagner. (*Ann. Phys.* (*Lpz.*), vol. 7, pp. 302–311; March 16, 1961). Results of measurements are compared with the theories of other authors.

538.221:621.318.134 4232

Magnetic Relaxation in Magnetite and in its Solid Solution with Manganese and Nickel Ferrites—A. Bragiński and T. Merceron. (*Nature*, vol. 191, pp. 898–899; August 26, 1961.) Relaxation was examined in the systems $Mn_xFe_{3-x}O_{4+y}$ and $Ni_xFe_{3-x}O_{4+y}$ by measuring their permeability disaccommodation. In solid solutions of magnetite with other cubic ferrites, the formation of local configurations of anisotropic space charge is explained by the presence of octahedral sites having trigonal symmetry. Peaks in the disaccommodation spectrum are explained by migration of these vacancies.

538.221:621.318.134 4233

The Magnetic Spectrum of Ni-Zn Ferrites with Perminvar Loop—F. Voigt. (*Naturwiss.*, vol. 48, pp. 215–216; April, 1961.) Report of measurements of the frequency dependence of complex permeability on two types of Ni-Zn ferrite, one containing CoO, with or without annealing in a transverse magnetic field.

538.221:621.318.134 4234

Magnetic Anomaly in $FeTiO_3$—αFe_2O_3 System by Mossbauer Effect—S. L. Ruby and G. Shirane. (*Phys. Rev.*, vol. 123, pp. 1239–1240; August 15, 1961.) The anomaly revealed by the Mossbauer measurements is that the ferrimagnetic phase consists of ferrimagnetic clusters surrounded by paramagnetic media.

538.221:621.318.134:538.569.4 4235

Gyromagnetic Resonance of Ferrites and Garnets in the Frequency Range 400–1200 Mc/s—J. Deutsch and H. G. Maier. (*Nachrtech. Z.*, vol. 14, pp. 142–146; March, 1961.) Experimental investigation of nonreciprocal attenuation near gyromagnetic resonance. (See also report in PROC. IRE, vol. 49, pp. 1443–1445 September, 1961.)

538.221:621.318.134:538.569.4 4236

Pressure Dependence of the Microwave Resonance Properties of some Spinel and Garnet Ferrites—I. P. Kaminow and R. V. Jones. (*Phys. Rev.*, vol. 123, pp. 1122–1129; August 15, 1961.)

538.221:621.318.134:538.569.4 4237

Nonlinear Damping Losses in YIG—R. L. Conger. (*J. Appl. Phys.*, vol. 32, pp. 1525–1527; August, 1961.) A theoretical investigation shows that nonlinear damping causes the microwave ferromagnetic resonance susceptibility to decrease with increasing power.

538.221:621.385.833 4238

The Origin of Image Contrast in the Display of Ferromagnetic Domains by means of an Illuminating Electron Microscope—J. Kranz and H. Bialas. (*Optik*, (*Stuttgart*), vol. 18, pp. 178–185; April, 1961.) A formula is given which expresses image contrast as a function of a) the gradient of magnetic induction over the specimen and b) the radius of the image on the screen. Experimental results confirm the conclusions reached.

538.222:538.569.4:621.375.9 4239

On the Theory of Cross-Relaxation in Maser Materials—M. Hirono. (*J. Phys. Soc. Japan*, vol. 16, pp. 766–776; April, 1961.) A theoretical study in which the cross-relaxation is calculated by the method of moments taking into account the effect of the crystalline electric field.

539.23:548.5:539.12.04 4240

Study of Electron Bombardment of Thin Films—G. B. Gilbert, T. O. Poehler and C. F. Miller. (*J. Appl. Phys.*, vol. 32, pp. 1597–1600; August, 1961.) Results obtained using a 35-kev electron beam indicate substantial crystal growth in films of In, Bi and Ge.

539.23:621.382.032.27 4241

Lead Wire Attachment Technique for Thin-Film Studies—R. I. George and R. B. McQuistan. (*Rev. Sci. Instr.*, vol. 32, pp. 855–856; July, 1961.) Au leads are used on a synthetic sapphire substrate.

621.315.5/.6:536.48:538.214 4242

Magnetic Susceptibility of Materials Commonly Used in the Construction of Cryogenic Apparatus—G. L. Salinger and J. C. Wheatley. (*Rev. Sci. Instr.*, vol. 32, pp. 872–874; July, 1961.)

621.315.5/.6].004.6 4243

The Long-Term Behaviour of Materials and Components in Electrical Engineering—J. Tretter. (*Frequenz*, vol. 15 pp. 39–47; February, 1961.) A theoretical interpretation of after-effect, aging and recovery processes in semiconductor, ferroelectric, ferromagnetic and other materials is attempted. Certain laws appear to be followed in the establishment of new equilibrium conditions in materials and components subsequent to the application of energy in the form of heat, light, electricity or mechanical stress.

MATHEMATICS

512.83:513.882 4244

Computational Problems concerning the Hilbert Matrix—J. Todd. (*J. Res. NBS*, vol. 65B, pp. 19–22; January–March, 1961.) The influence of theoretical mathematics and of computational experiment on each other is illustrated by a discussion of recent work.

517.522.2 4245

A Power Series Expansion of the Master Equation—N. G. van Kampen. (*Canad. J. Phys.*, vol. 39, pp. 551–567; April, 1961.)

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) 4246

The Standard Frequency of the Federal Post Office—J. Bastelberger. (*Nachrtech. Z.*, vol. 14, pp. 72–81; February, 1961.) A standard frequency of 1 kc with an accuracy within ±1.5 parts in 10^9 and hourly time pulses are available throughout the whole of West Germany. Details are given of the standard frequency equipment, the comparison system and the distribution network.

621.3.018.41 (083.74):621.396.61 4247

Effect of External Conditions on the Phase of Radio Signals—Watt-Carter and Corke. (See 4327.)

621.317.3:537.311.33 4248

Apparatus for the Measurement of Galvanomagnetic Effects in High-Resistance Semiconductors—G. Fischer, D. Greig and E. Mooser. (*Rev. Sci. Instr.*, vol. 32, pp. 842–846; July, 1961.) The resistivity of samples can be measured if the resistance is in the range $10^{-1}-10^{12}$ Ω. The Hall coefficient can also be determined provided the charge-carrier mobility of the sample exceeds 1 cm²/v·sec.

621.317.3:537.311.33 4249

Microwave Measurement of Mobility: Analysis of Apparatus—S. H. Liu, Y. Nishina and R. H. Good, Jr. (*Rev. Sci. Instr.*, vol. 32, pp. 784–789; July, 1961.) A semiconductor sample is mounted in a bimodal cavity with an applied static magnetic field and the power transfer produced by the Faraday rotation is measured. A formula is derived for the power-transfer ratio as a function of magnetic field, conductivity and mobility of the sample, and the properties of the equipment.

621.317.3:537.311.33 4250

Microwave Measurement of Hall Mobility: Experimental Method—Y. Nishina and G. C. Danielson. (*Rev. Sci. Instr.*, vol. 32, pp. 790–793; July, 1961.) The measurements were made at 9 Gc, over the temperature range 30–300°K, and the analysis given by Liu, *et al.* (4249 above) was verified by observations on an *n*-type sample.

621.317.3:537.311.33 4251

Microwave Determination of Semiconductor Carrier Lifetimes—H. A. Atwater: H. Jacobs. (PROC. IRE, vol. 49, pp. 1440–1442; September, 1961.) Atwater suggests that a microwave power transient will not in general follow the exponential decay of the injected carriers [see 3774 of 1949 (Ramsa *et al.*)]. Jacobs points out a flaw in the criticism and describes an alternative approach which gives good agreement between theoretical and experimental values [2730 of August (Jacobs *et al.*).

621.317.33:537.311.33 4252

An A.C. Silicon Resistivity Meter—C. C. Allen and W. R. Runyan. (*Rev. Sci. Instr.*, vol. 32, pp. 824–828; July, 1961.) A description of a four-point probe resistivity meter, which is direct reading and covers a resistivity range of 0.01–300 Ω cm for full-scale deflection.

621.317.34.029.64:621.372.824 4253

Precision Instruments for Coaxial Line Measurements up to 4 Gc/s—D. Woods. (*Proc. IEE*, pt. B, vol. 108, pp. 490–492; September, 1961. Discussion, pp. 493–494.) A program to improve the standardization and measurement of admittance, voltage, power and attenuation is described.

621.317.373 4254

Phase Measurement at Audio and High Frequencies—O. Macek. (*Arch. tech. Messen*, no. 301, pp. 35–38; February, 1961.) The methods reviewed include those using cathode-ray oscilloscopes, ring modulators, and pulse-controlled multivibrators. 31 references.

621.317.44:538.632 **4255**

Measurements in Inhomogeneous Magnetic Fields by means of the Hall Probe—P. Menzel (*Arch. tech. Messen*, no. 303, pp. 83–86; April, 1961.) The characteristics and limits of accuracy of Hall-effect devices used for the measurement of magnetic induction are discussed.

621.317.7:621.372.2.012.11 **4256**

Smith-Diagram Display Unit uses Microwave Phase-Directional Coupler—S. Tsuchiya. (*Electronics*, vol. 34, pp. 80–84; July 21, 1961.) Microwave impedances can be measured and displayed directly on a cathode-ray tube.

621.317.733 **4257**

Choosing Transformer Ratio-Arm Bridges—W. H. P. Leslie. (*Proc. IEE*, pt. B, vol. 108, pp. 539–545; September, 1961.) Five bridges are considered in detail and their suitability for the following applications is discussed: a) similar impedances with appreciable capacity to earth from each terminal, b) comparison of low impedances when lead impedance is not negligible and c) measurement by substitution. Karo's circuit (554 of 1959) shows no practical advantage and is based on incorrect analysis.

621.317.74:621–526 **4258**

An Automatic Electronic Nyquist Plotter—H. J. Fraser and W. V. P. Reece (*Proc. IEE*, pt. B, vol. 108, pp. 535–538; September, 1961.) An instrument to plot automatically the transfer function of electrical networks or servo systems over the frequency range 0.2–200 cps.

621.317.74:621.396.679.4 **4259**

Measurements on a Transmission-Line Ring-Hybrid—A. J. Hymans. (*Electronic Eng.*, vol. 33, pp. 578–581; September, 1961.) The design of accessories for measurements in the range 20–30 Mc is described including a matched load for 2-kw dissipation.

621.317.761:538.569.4 **4260**

A Magnetically Tunable Microwave Frequency Meter—G. H. Thiess. (PROC. IRE, vol. 49, pp. 1439–1440; September, 1961.) A note on a method of frequency measurement based on ferromagnetic resonance [see 1496 of 1956 (Vartanian and Melchor)]. An accuracy within 0.3 per cent can be achieved using single-crystal spheres of Y-Fe garnet.

621.317.794.089.6 **4261**

A Bolometer Mount Efficiency Measurement Technique—G. F. Engen (*J. Res. NBS*, vol. 65C, pp. 113–124; April–June, 1961.) A variation of the impedance measurement method is described which gives increased accuracy and simplifies the operational procedure. It is independent of connector discontinuities and is applicable to matched and unmatched mounts.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621–52:681.142 **4262**

Statistical Design of Linear Discrete-Data Control Systems via the Modified *z*-Transform Method—J. T. Tou. (*J. Franklin Inst.*, vol. 271, pp. 249–262; April, 1961.)

621.317.361.029.42:612 **4263**

Frequency Measurement of a Strongly Disturbed and Damped Oscillation—R. Giersiepen. (*Frequenz*, vol. 15, pp. 105–111; April, 1961.) The problem considered is that of measuring the natural frequency of the blood circulating system. The use of an electrical equivalent of the circulatory system enables the interference function to be simulated and subtracted from the input function.

621.317.39:531.71:537.312.8 **4264**

Design of Magnetoresistance Elements—D. H. Roberts. (*J. Sci. Instr.*, vol. 38, pp. 332–333; August, 1961.) Suggestions are made for improving the performance of an instrument for measuring linear displacement.

621.362:621.387 **4265**

Optimization of Emission-Limited Thermionic Generators—A. Schoek. (*J. Appl. Phys.*, vol. 32, pp. 1564–1570; August, 1961.) Equations are derived describing the performance of space-charge-neutralized thermionic convertors. An expression is derived for maximum conversion efficiency in terms of the system parameters.

621.383.5:621.3.087.9 **4266**

A Fast Numeral Reading Machine—G. Fougere, J. Preston and R. Farnsworth. (*Electronic Eng.*, vol. 33, pp. 562–565; September, 1961). The machine optically reads three digit numbers printed on plastic tape at a rate of 150 characters per second.

621.385.833 **4267**

Improved Electron Filter Lens—J. A. Simpson and L. Marton. (*Rev. Sci. Instr.*, vol. 32, pp. 802–803; July, 1961.)

621.385.833 **4268**

First-Order Examination of an Optical System Formed from Four Quadrupolar Electrostatic Lenses—D. Dhuicq. (*C. R. Acad. Sci.* (*Paris*), vol. 251, pp. 1989–1991; November 7, 1960.)

621.387.4:621.383.292 **4269**

Semiconductors yield Improved Electron Multiplier—F. A. White, J. C. Sheffield and W. D. Davis. (*Nucleonics*, vol. 19, pp. 58–60; August, 1961.) The magnetic electron multiplier described, which is designed for use as an ion detector, incorporates a parallel-plate dynode system consisting of Au-doped Si.

621.387.4:621.383.292 **4270**

Resistance-Strip Magnetic Electron Multiplier—G. W. Goodrich and W. C. Wiley. (*Rev. Sci. Instr.*, vol. 32, pp. 846–849; July, 1961). The multiplier uses a continuous surface of high-resistance semiconductor materials in place of a conventional multielement dynode structure.

621.398:621–52 **4271**

Analogue and Digital Methods and their Application in Telemetry—K. Bauer. (*Z. Ver. Dtech. Ing.*, vol. 103, no. 24, pp. 1209–1217; August 21, 1961.) Examples of circuits and equipment are given and their operation is described with particular regard to the presentation in digital form of process control measurements.

621.561.57:537.322.15 **4272**

Frigistors—Thermocouples of Semiconductor Material for use as Cooling Elements and Heat Pumps—W. Dörr. (*Elektron. Rundschau*, vol. 15, pp. 107–111; March, 1961.) Various patterns of commercial-type Bi-Te junctions using the Peltier effect are described.

PROPAGATION OF WAVES

621.391.812.62 **4273**

Effect of Multiple Atmospheric Inversions on Tropospheric Radio Propagation—F. H. Northover. (*J. Res. NBS*, vol. 65D, pp. 385–392; July/August, 1961.) An earlier theory (1409 of 1952) for a single elevated inversion layer is extended to cover the case of multiple layers. Under certain conditions, several weak high-level inversions can produce a similar effect to a single strong inversion.

621.391.812.62 **4274**

Smooth-Earth Diffraction Calculations for Horizontal Polarization—L. E. Vogler. (*J. Res. NBS*, vol. 65D, pp. 397–399; July/August, 1961.) A simplified method of calculating attenuation in the "far diffraction region" is described, and an improved criterion is given for determining the ranges of validity of the method.

621.391.812.63 **4275**

Effects of Equatorial Spread-*F* Irregularities on C. W. Transmissions—M. S. V. G. Rao and B. R. Rao. (*Canad. J. Phys.*, vol. 39, pp. 596–603; April, 1961.) A CW signal on 11.7 Mc was recorded to study the fading rate during spread-*F* conditions and the horizontal extent of spread-*F* patches affecting the 1300-km path from Colombo to Waltair. The fading rate increased from 10.7 cycles per minute (no spread-*F*) to about 40 cycles per minute (severe spread-*F*) with more rapid flutter also present. Patches of more than 650-km extent in the N-S direction apparently occur over this path.

621.391.812.63 **4276**

Observations of Unusual Low-Frequency Propagation made on 12th November 1960—J. S. Belrose and D. B. Ross. (*Canad. J. Phys.*, vol. 39, pp. 609–614; April, 1961.) Amplitude and phase measurements were made near 80 kc in northern Canada.

621.391.812.63:551.510.535 **4277**

On the Validity of some Approximations to the Appleton-Hartree Formula—K. Davies and G. A. M. King. (*J. Res. NBS*, vol. 65D, pp. 323–332; July/August, 1961.) Various approximations for the refractive index are compared with values computed from the complete formula, for various latitudes and a frequency of 2.0 Mc. Some approximations become very poor only a short distance from where they are exact. Formulas are given that yield sufficiently accurate values for all geomagnetic latitudes. Certain approximations to the group refractive indexes are also considered.

621.391.812.63:551.510.535 **4278**

Investigation of Ionospheric Absorption on 5.65 Mc/s at Waltair—K. V. V. Ramana and B. R. Rao. (*J. Atmos. Terr. Phys.*, vol. 22, pp. 1–11; September, 1961.) Measurements were made during the I.G.Y. and I.G.C. periods and the diurnal and seasonal variations of absorption were studied. The simple relation log $\rho \propto f^{-2} \cos^n \chi$ is only approximate and the observations can be better fitted to an expression depending on Jaeger's function (3252 of 1947).

621.391.812.63:551.510.535 **4279**

The Fading of Radio Waves Reflected Vertically from the Ionosphere during Magnetic Storms—Fooks. (See 4133.)

621.391.812.63.029.45:621.396.67 **4280**

On the Possibility of Rejecting Certain Modes in V.L.F. Propagation—J. R. Wait. (PROC. IRE, vol. 49, pp. 1429–1430; September, 1961.) An analysis of the second-order mode and its dependence on the structure of the transmitting antenna.

621.391.814.2.029.4/.5 **4281**

On the Theory of Mixed-Path Ground-Wave Propagation on a Spherical Earth—J. R. Wait. (*J. Res. NBS*, vol. 65D, pp. 401–410; July/August, 1961.) "The problem formulated concerns the mutual impedance between two vertical dipole antennas *A* and *B* located near the surface of a spherical smooth earth. The path between *A* and *B* is made up of a number of homogeneous segments where the surface impedance is constant. Various formulas are developed, for two- and three-section paths, which are suitable for computation. Certain limiting cases are discussed and where possible a physical interpretation of the results is given. Comparisons with previous work are made."

621.391.814.2.029.53 4282

A Few Observations of the Perturbations in the Phase of the Low-Frequency Ground Wave—J. M. Ross and J. E. Kirch. (*J. Res. NBS*, vol. 65D, pp. 393–396; July/August, 1961.) "The effects of anomaly-producing terrain features on the phase of the low-frequency ground wave were measured in a nearly idealized environment to confirm theoretical predictions. Results for an isolated butte gave good correlation with the theoretical model. In most cases the mean value of the phase perturbation approached zero."

RECEPTION

621.376.23:523.164 4283

Digital Methods for the Extraction of Phase and Amplitude Information from a Modulated Signal—R. S. Lawrence, J. L. Jesperson and R. C. Lamb. (*J. Res. NBS*, vol. 65D, pp. 351–356; July/August, 1961.) A zero-crossing method, a filter method and a statistical approach intended for the analysis of radio-star scintillations are described. They may be applied to other types of modulated signals.

621.391.81 4284

Correlation Measurements on Frequency Diversity in the Short-Wave Range—J. Grosskopf, G. Heinzelmann and K. Vogt. (*Nachrtech. Z.*, vol. 14, pp. 124–128; March, 1961.) Report on investigations of frequency-diversity reception over the path Tokyo-Frankfurt (Main) using SSB transmitters on 19.79 and 14.73 Mc, the frequency diversity in each case being obtained by AF modulation of the transmitter. A reduction of the frequency spacing normally used in frequency-diversity systems appears to be possible. For investigations of space-diversity reception see 953 of 1959 (Grosskopf, *et al.*).

621.391.812.63 4285

Some Experiments on the Reception of Ionospheric Waves—M. Mancianti. (*Alta Frequenza*, vol. 30, pp. 293–295; April, 1961.) Report on reception tests at 872 kc to investigate phase distortion due to ionospheric propagation of AM signals. Typical oscillograms are reproduced. For English version see *Alta Frequenza*, vol. 30, pp. 543–544; July, 1961.)

621.391.82:551.507.362 4286

Analysis and Prediction of Radio Signal Interference Effects due to Ionized Layer around a Re-entry Vehicle—W. C. Taylor. (*Planet-Space Sci.*, vol. 7, pp. 286–300; July, 1961.) The problem of transmitting em waves through the thermally ionized layer of gases in the shock layer round a hypersonic missile is considered.

621.391.823:621.317.3(083.74) 4287

IRE Standards on Radio Interference: Methods of Measurement of Conducted Interference Output to the Power Line from F.M. and Television Broadcast Receivers in the Range of 300 kc to 25 Mc, 1961—(PROC. IRE, vol. 49, pp. 1398–1402; September, 1961.) Standard 61 IRE 27.S1.

621.396.62:534.76 4288

Multiplex Adaptors for Compatible F.M. Stereo Reception—L. Solomon. (*Electronics*, vol. 34, pp. 45–47; August 18, 1961.) A summary of methods and design principles.

621.396.62:621.396.71 4289

New Short-Wave Large-Station Receiving Equipment for Telegraphy and Telephony—D. Leypold, P. Schucht and H. Wich. (*Frequenz*, vol. 15, pp. 45–55; February, 1961.) Equipment for SSB and telegraphy diversity operation is described.

621.396.62:621.396.71:621.394.62 4290

Telegraphy Equipment in the New Short-Wave Receiver for Large Stations—H. Meissner. (*Frequenz*, vol. 15, pp. 56–58; February, 1961.) See 4289 above.

621.396.621.54:621.391.82 4291

The Determination of the Image Response in a Superheterodyne with regard to Noise-Factor Measurement—M. Piatelli. (PROC. IRE, vol. 49, p. 1435; September, 1961.)

621.396.666:621.391.812.63 4292

Elimination of Ionospheric Refraction Effects—J. M. Brito-Infante. (PROC. IRE, vol. 49, pp. 1451–1452; September, 1961.) If radio signals from a mobile transmitter are received on several different frequencies, the frequency shift due to ionospheric refraction can be eliminated. The analytical procedure for using this principle in practice is given.

STATIONS AND COMMUNICATION SYSTEMS

621.376.3 4293

Power in a Frequency-Modulated Wave—R. G. Medhurst (*Electronic Tech.*, vol. 38, pp. 378–379; October, 1961.) The time variation of the mean power in a sampling period is calculated for the general case where there is no specific relation between modulation frequency and carrier frequency or sampling period.

621.39(94) 4294

Broadband Telecommunication Systems—A. H. Kaye, F. D. Morris, J. F. Sinnatt, P. H. Richards, J. R. Walklate and J. C. Wilson. (*J. Inst. Engrs.* (*Aust.*), vol. 33, pp. 213–234; June, 1961.) A collection of six papers on operative and proposed Australian communication systems.

621.391 4295

Reconstruction Error and Delay for Amplitude-Sampled White Noise—D. S. Ruchkin. (PROC. IRE, vol. 49, p. 1436; September, 1961.) An expression is derived for the mean-square reconstruction error.

621.391:621.376.5 4296

Instantaneous-Value Companding in Time-Modulated Transmission Systems—W. Glaser. (*Nachrtech.*, vol. 11, pp. 106–114; March, 1961.) The application of a companding method to pulse-modulated systems, and the resultant advantages, are discussed.

621.396.43:523.3 4297

Long-Distance Communication via the Moon—P. A. Webster. (*J. Brit. IRE*, vol. 22, pp. 257–264; September, 1961.) Optimum frequency, type of modulation, antenna size and other requirements for economic operation of a circuit via the moon are considered. Experiments at 210 Mc using a 1-kw transmitter and the Jodrell Bank radio telescope are described.

621.396.43:523.3 4298

A Study of U.H.F. Space Communications through an Aurora using the Moon as a Reflector—R. P. Ingalls, J. C. James and M. L. Stone. (*Planet-Space Sci.*, vol. 7, pp. 272–285; July, 1961.) Signals at 440 Mc from a transmitter at College, Alaska, in the auroral zone were reflected from the moon and received at two stations in the midlatitude zone. No decrease of received power was noticed when the propagated signal passed through an auroral disturbance but the observed fading rate increased and the observed polarization angle fluctuated rapidly.

621.396.43:523.5 4299

Propagation of Meteor-Burst Signals during the Polar Disturbance of November 12–16, 1960—L. A. Maynard. (*Canad. J. Phys.*, vol. 39, pp. 628–630; April, 1961.) It is concluded that the fall in signal strength was due to absorption below the meteoric region at heights of 80–100 km.

621.396.43:551.507.362.2 4300

Radio Communication by means of Satellites—C. C. Cutler. (*Planet-Space Sci.*, vol. 7, pp. 254–271; July, 1961.) An examination is made of the difficulties of long-distance radio communication by satellite, comparing the simplicity of a passive repeater with a very low reflected signal to the complex power and orientation problems of a solar-powered transmitter-receiver with much narrower bandwidth.

621.396.43:551.507.362.2 4301

Satellite-Supported Communication at 21 Megacycles—R. Soifer. (PROC. IRE, vol. 49, pp. 1455–1456; September, 1961.) A group of transmitters and receivers operated near New York in 1959–1960. Signal bursts which occurred during six-minute periods before, during, and after satellite passes were recorded. 29 bursts occurred during the central periods. The statistical probability that the bursts are associated with the passes is very high.

621.396.43:551.507.362.2 4302

The Advantages of Attitude Stabilization and Station Keeping in Communications Satellite Orbits—W. F. Hilton and B. Stewart. (*J. Brit. IRE*, vol. 22, pp. 193–202; September, 1961.) Satellites using high-gain antennas requiring stabilizing equipment in the payload and unstabilized satellites using higher transmitter power are compared. Spin stabilization is examined and random and station-keeping satellites are discussed.

621.396.43:551.507.362.2 4303

A Proposal for an Active Communication Satellite System based on Inclined Elliptic Orbits—B. Buss and J. R. Millburn. (*J. Brit. IRE*, vol. 22, pp. 209–225; September, 1961.) Four satellites in a 63° elliptical orbit would give a good service to the northern hemisphere. The design for a 1000-telephone-channel satellite having a radiated power of 12 w at the optimum frequency of 2 Gc is given.

621.396.43:551.507.362.2 4304

Some International Aspects of Satellite Communication System Planning—C. F. Booth. (*J. Brit. IRE*, vol. 22, pp. 227–229; September, 1961.) The allocation of frequency bands, methods of minimizing interference with other services and the type of modulation and base-band characteristics require international agreement.

621.396.43:551.507.362.2:621.375.9:538.569.4 4305

The Maser and its Application to Satellite Communication Systems—Hlawiczka. (See 4057.)

621.396.43:621.396.677 4306

Antenna-Beam Configurations in Scatter Communications—Waterman. (See 4009.)

621.396.43:629.19 4307

Long-Range Radio Communication by Satellite Microwave Dipoles—(*Nature*, vol. 191, pp. 1237–1238; September 23, 1961.) A note on the Space Science Board's general recommendations on the West Ford Project.

621.396.65 4308

The Development of the R.T.F. Radio-Link System—F. Anguéra. (*E.B.U. Rev.*, no. 66A, pp. 61–64; April, 1961.) Fundamental design principles are discussed covering technical and economic criteria.

621.396.712:621.316.9 4309

The Lightning Protection of Broadcasting Stations in Norwegian Mountains—H. Otterbech. (*E.B.U. Rev.*, no. 66A, pp. 58–60; April, 1961.) Protective measures are summarized and statistical data of station breakdowns in relation to thunderstorm occurrence are given.

621.396.74:681.84.087.7 4310

Review of the Main Proposals for the Broadcast Transmission of Stereophony—K. Wilhelm. (*Nachrtech. Z.*, vol. 14, pp. 129–141; March, 1961.) International proposals are summarized and compared. 20 references.

621.396.946 4311

The Possibility of Reception of Communications from Extraneous Planetary Systems—M. v. Ardenne and G. Böhme. (*Hochfrequenz. und Elektroak.*, vol. 70, pp. 2–5; February, 1961.) The feasibility and transmitter power requirements of a communication system with a planet at a distance of ten light years are estimated on the basis of information theory and galactic noise data. See, *e.g.*, 2900 of 1960 (Bracewell).

SUBSIDIARY APPARATUS

621-526:522.6 4312

Servo System Design for Balloon-Borne Star Trackers—W. J. Wichman and M. M. Birnbaum. (*Electronics*, vol. 34, pp. 43–46; September 1, 1961.) The telescope-pointing equipment can operate over a dynamic light range of 1,000,000:1, the average pointing error being within 5 seconds of arc.

621.3.087.4:621.395.625.3 4313

Magneto-optical Readout of Magnetic Recordings—T. Lentz and J. Miyata. (*Electronics*, vol. 34, pp. 36–39; September 1, 1961.) This technique is based on the rotation of the plane of polarization of a light beam reflected from a magnetized surface.

621.311.69:629.19 4314

Power Supplies for Space Vehicles—K. E. V. Willis. (*J. Brit. IRE*, vol. 22, pp. 179–188; September, 1961.) Available unconventional power sources are reviewed in the light of possible future requirements.

621.314.63 4315

Silicon Power Rectifiers—A. J. Blundell, A. E. Garside, R. G. Hibberd and I. Williams. (*Proc. IEE*, pt. A, vol. 108, pp. 273–288; August, 1961. Discussion, pp. 288–294.) The processes involved in preparing single-crystal Si are outlined and details of design, production techniques and characteristics of Si rectifiers are given.

TELEVISION AND PHOTOTELEGRAPHY

621.397.13:389.6 4316

Standard Conversion of Television Signals with Different Vertical Scanning Frequency—E. Sennhenn. (*Radio Mentor*, vol. 27, pp. 184–188; March, 1961.) Problems arising in the reproduction of recorded television signals of differing field frequency are discussed. An automatic control system is described for compensating the fluctuations produced by the conversion process.

621.397.132 4317

Circuit Technique of Demodulators for Colour Television Signals according to the N.T.S.C. Method—F. Jaeschke. (*Arch. elekt. Übertragung*, vol. 15, pp. 187–199; April, 1961.) Design problems peculiar to color television engineering are discussed and several types of demodulator circuits are considered. For the design of a modulator see, 2762 of 1959.

621.397.331.21 4318

Investigations on Image-Iconoscope Camera Tubes by means of Special Test Equipment—A. Kaufmann. (*Rundfunktech. Mitt.*, vol. 5, pp. 27–34; February, 1961.) The test equipment is described and a test specification meeting practical studio requirements is proposed. Tests were made on different types of tubes to determine the dependence of tube performance on parameters such as temperature, beam current and collector voltage.

621.397.331.22:621.397.61 4319

The Simplification of the Operation of Image Orthicons—T. Mayer and G. E. Partington. (*Rundfunktech. Mitt.*, vol. 5, pp. 1–7; February, 1961.) Description of the principal features of the $4\frac{1}{2}$-in image orthicon and its applications. See also 1658 of May (Partington).

621.397.331.222 4320

Operation of Vidicons in Unusual Environmental Conditions—G. A. Robinson. (*J. Soc. Mot. Pic. Telev. Engrs.*, vol. 70, pp. 264–266; April, 1961.) The effects of temperature, high pressure and nuclear radiation on vidicon tubes were examined.

621.397.334:621.396.621 4321

Progress Report on the Development of a Photoelectric Beam-Index Colour-Television Tube and System—R. Graham, J. W. H. Justice and J. K. Oxenham. (*Proc. IEE*, pt. B, vol. 108, pp. 511–520; September, 1961. Discussion, pp. 520–523.) The principle of operation of the Zebra system is described and methods of overcoming some of the basic problems of indexing are outlined. Comparisons are drawn between index and other color television displays. See also 1044 of March (Macwhirter).

621.397.6:535.376 4322

The Application of Electroluminescence in Television Techniques—L. Jung. (*Tech. Mitt. BRF, Berlin*, vol. 5, pp. 25–36; March, 1961.) Present-day developments in the design of image intensifiers, image display and storage devices are reviewed.

621.397.61:621.376.2 4323

The Modulation of Television Transmitters in the Intermediate-Frequency Range—P. Klopf. (*Rundfunktech. Mitt.*, vol. 5, pp. 79–87; April, 1961.) The advantages of using low-level modulation at intermediate frequency are reviewed, with particular reference to results obtained with C.C.I.R. modified N.T.S.C. color television signals. See also 1035 of 1960.

621.397.7-182.3 4324

Mobile Eurovision Installation of Österreichischer Rundfunk [Austrian Broadcasting System]—F. Brunner. (*Rundfunktech. Mitt.*, vol. 5, pp. 23–26; February, 1961.) The installation described comprises all the equipment necessary for Eurovision transmissions apart from that carried by the television outside-broadcast vehicle.

621.397.7-182.3 4325

The Television Outside-Broadcast Vehicles for Southern Switzerland—H. Probst. (*Tech. Mitt. PTT*, vol. 39, pp. 71–87; March, 1961. In German and Italian.) Detailed description of the mobile studio installation which is housed in two vehicles.

621.397.74 4326

Video Cross-Bar Distributor—H. H. Freytag. (*Rundfunktech. Mitt.*, vol. 5, pp. 61–65; April, 1961.) Description of a distribution installation for television networks which is remotely controlled and incorporates transistor amplifiers.

TRANSMISSION

621.396.61:621.3.018.41(083.74) 4327

Effect of External Conditions on the Phase of Radio Signals—D. E. Watt-Carter and R. L. Corke. (*Nature*, vol. 191, p. 1286; September 23, 1961.) Phase errors in the 16-kc transmission from Rugby are due to a) antenna detuning caused by changing weather and antenna sway, and b) mechanical fluctuations of the tuning inductors and transient thermal and electrical effects in the amplifiers when the transmitter is keyed. Only those due to a) would be likely to be observed at a distance.

TUBES AND THERMIONICS

621.382(083.74) 4328

International State of Standardization of Semiconductor Components—M. Falter. (*Nachrtech.*, vol. 11, pp. 50–55; February, 1961.) Standards of terms and symbols are compared and standardized measuring circuits are given.

621.382.23 4329

Interpretation of the Tunnel-Diode Characteristic from its Variation with Temperature—G. Winstel. (*Z. Naturforsch.*, vol. 16a, pp. 438–440; April, 1961.) The characteristics of Ge and GaAs tunnel diodes were investigated in the temperature range 100–400° K.

621.382.23 4330

Diode Reverse Characteristics at Low Temperatures—A. S. Clorfeine. (PROC. IRE, vol. 49, pp. 1425–1426; September, 1961.) The characteristics of inductance and negative resistance have been observed in a junction diode at temperatures below 78°K. Possible explanations are advanced.

621.382.23 4331

Reverse Characteristics of Low-Lifetime Germanium Diodes—A. Blicher and I. H. Kalish. (PROC. IRE, vol. 49, p. 1427; September, 1961.) Measurements are described which confirm the theory that carrier generation in the depletion region is responsible for the nonsaturable reverse characteristics of Ge *pn* junctions.

621.382.23 4332

The Tunnel Diode: Physical Basis, Manufacture and Applications—H. J. Hartmann, M. Michelitsch and W. Steinhäuser. (*Arch. elekt. Übertragung*, vol. 15, pp. 125–144; March, 1961.) A summary giving physical principles and circuitry. 30 references.

621.382.23 4333

Temperature Dependence of the Peak Current of Germanium Tunnel Diodes—A. Blicher, R. M. Minton and R. Glicksman. (PROC. IRE, vol. 49, pp. 1428–1429; September, 1961.) The temperature dependence was investigated as a function of carrier concentration.

621.382.23 4334

The Effect of Nonsymmetrical Doping on Tunnel Diodes—F. D. Shepherd and A. C. Yang. (PROC. IRE, vol. 49, pp. 1435–1436; September, 1961.) Esaki's model (1784 of 1958) is used to explain the I/V characteristics of the tunnel diode. The properties are shown to be quite dependent upon the degeneracy ratio and the cutoff voltage.

621.382.23:621.372.44 4335

Negative-Resistance Diode handles High Power—A. P. Schmid, Jr. (*Electronics*, vol. 34, pp. 44–46; August 25, 1961.) Theory and applications of a bonded diode with current-controlled negative resistance are discussed.

621.382.23:621.372.44 4336

The Negative Resistances in Junction Diodes—I. Hefni. (PROC. IRE, vol. 49, pp. 1427–1428; September, 1961.) Negative-resistance effects are described which could explain the anomalous *V/I* characteristics of variable-capacitance diodes used in parametric devices [see, *e.g.*, 2804 of August (McDade)].

621.382.233:621.318.57 4337

The Silicon Four-Layer Diode and its Applications as an Electronic Switch—H. Keller and G. Wieczorek. (*Frequenz*, vol. 15, pp. 33–3[illegible]; February, 1961.) The principle of operation and several practical switching circuits are described.

621.382.233:621.318.57 4338

Minimum Time for Turn-Off in Four-Layer Diodes—M. A. Melehy. (PROC. IRE, vol. 49, p. 1424; September, 1961.) An approximate simple form for the minimum time τ_t required to turn off a four-layer diode is derived and a method of plotting τ_t against the device constants is demonstrated.

621.382.3 4339

'Spread' in Transistor Samples—R. Paul. (*Nachrtech.*, vol. 11, pp. 56–65; February, 1961.) Investigation of the causes of deviation in the characteristics of individual mass-produced transistors.

621.382.3 4340

Problems of Lifetime and 'Spread' of Transistors—G. Fassbender. (*Nachrtech.*, vol. 11, pp. 66–71; February, 1961.) Ways of improving transistor quality and reliability are discussed with reference to curves giving the spread in the characteristics of manufactured components.

621.382.3 4341

The Recombination of Injected Charge Carriers in Transistors—H. U. Harten. (*Elektron. Rundschau*, vol. 15, pp. 111–116; March, 1961.) The significance of the effect and means of reducing it are discussed.

621.382.3 4342

Some Advantages of Silicon Transistors in Circuit Design—M. K. McPhun. (*Proc. IEE*, pt. B, vol. 108, pp. 570–575; September, 1961.) The performance of Si transistors is compared with that of Ge devices from the viewpoint of the circuit designer.

621.382.3:621.375.4 4343

Stabilization of the D.C. Operating Point of Transistors—A. E. Bachmann. (*Tech. Mitt. PTT*, vol. 39, pp. 88–99; March, 1961.) Account is taken of the effect of emitter-base voltage changes on stability. Tables are included which give the stability factors of two- and three-stage direct-coupled transistor amplifier circuits.

621.382.3:621.375.9:621.372.44 4344

Parametric Amplification with Transistors—Rohde. (See 4060.)

621.382.3.004.6:519.24 4345

The Representation of the Results of Life Tests in the Manufacture of Transistors—C. Villalaz. (*Arch. elekt. Übertragung*, vol. 15, pp. [illegible]1–186; April, 1961.) The problem of storing and evaluating the data obtained from reliability studies is considered. The limitation and subdivision of the information range on the basis of statistical principles are considered, and a system of symbols is given for the various types of life-test curve obtained. The evaluation of life-test results as discussed could be carried out automatically.

621.382.32 4346

D.C. Characteristic of the Unipolar Transistor taking account of Loss Impedances—A. H. Frei. (*Arch. elekt. Übertragung*, vol. 15, pp. 161–165; April, 1961.) A formula is derived for the dc characteristics of the unipolar field-effect transistor (tecnetron) assuming smooth distribution of the space-charge field in the active channel. Good agreement between theoretical and measured results is obtained.

621.382.32:539.12.04 4347

Electron-Radiation Damage in Unipolar Transistor Devices—B. A. Kulp, J. P. Jones and A. F. Vetter. (PROC. IRE, vol. 49, pp. 1437–1438; September, 1961.) The characteristics after various degrees of electron bombardment are shown. Some recovery could be obtained by annealing.

621.382.333 4348

The Thermal Behaviour of Transistors with Nonstationary Collector Dissipation—H. Reimann. (*Nachrtech. Z.*, vol. 14, pp. 69–72; February, 1961.) Theoretical investigation of conditions in a transistor with fluctuating load, on the basis of a rough approximation.

621.382.333 4349

Noise Investigations on Silicon Transistors—B. Kurz. (*Bull. schweiz. elektrotech. Ver.*, vol. 52, pp. 369–377; May 20, 1961.) Results of measurements on Si alloy-junction transistors agree satisfactorily with the theoretical noise factors obtained..

621.382.333.015.5 4350

Voltage Breakdown of Transistors—S. C. Ryder-Smith. (*Electronic Tech.*, vol. 38, pp. 348–356; October, 1961.) The mechanisms of avalanche and punch-through breakdown are explained and the effects of external circuit elements on the breakdown voltages are discussed. The application of voltage ratings to practical circuits is considered with a view to protecting the transistors without being unduly restrictive.

621.382.333.33:621.318.57 4351

Temperature Effects on GaAs Switching Transistors—E. D. Haidemenakis, J. A. Mydosh, N. Almeleh, R. Bharat and E. L. Schork. (PROC. IRE, vol. 49, p. 1448; September, 1961.) The current transfer ratio increases slightly at high temperatures and increases ten times at low temperatures.

621.383.292 4352

The Responsivity Profile of Photomultipliers—H. C. Ingrao and J. M. Pasachoff. (*Rev. Sci. Instr.*, vol. 32, pp. 866–867; July, 1961.)

621.385.032.213.23 4353

Improvement in Oxide-Cathode Stability by Controlled Electrolysis—A. Sandor. (*Le Vide*, vol. 16, pp. 79–90; May/June, 1961. In French and English.)

621.385.032.213.23 4354

Effect of Ultraviolet Irradiation on the Electrical Conductivity of Cathodes Coated with Barium Oxide—C. H. B. Mee. (*Nature*, vol. 191, pp. 1383–1384; September 30, 1961.) Ultraviolet irradiation increased the room-temperature dark conductivity of a BaO-coated cathode by a factor of about 2 at 370°K after which the effect decreased with the increase of temperature and disappeared when 550°K was exceeded. After cooling, the cathode was shown to have returned to its original state.

621.385.032.213.62 4355

The Lifetime of a Tungsten Heater Filament under Different Operating Conditions—B. Berkeš. (*Arch. Elektrotech.*, vol. 46, pp. 59–63; February 15, 1961.) The investigations are concerned with the heater filaments of saturated diodes used in current or voltage stabilizers. The operating conditions considered are a) constant heater voltage, b) constant heater current, c) constant filament temperature, and d) constant anode current. Formulas are given for calculating the stability of voltage or current regulators controlled by such diodes.

621.385.032.269.1:621.391.822 4356

The Finite-Beam Space-Charge-Limited Diode as a Noise Transducer—R. A. McFarlane. (*Canad. J. Phys.*, vol. 39, pp. 568–578; April, 1961.) A method is presented for calculating the noise properties of a cylindrical electron beam from a space-charge limited diode electron gun. The multivelocity character of the beam is considered and correction is made for finite beam diameter. The ratio of anode noise current to full-shot noise is found to depend on $\omega^{1/2}/V_{0a}$, where ω is the frequency and V_{0a} the anode voltage.

621.385.032.269.1:621.391.822 4357

Microwave Noise in Accelerated Electron Streams—R. A. McFarlane. (*Canad. J. Phys.*, vol. 39, pp. 579–587; April, 1961.) Noise current fluctuations on the beam from a space-charge-limited diode electron gun were measured at 1400, 4250 and 9520 Mc. The measurements are compared with a theory which takes account of the finite-beam diameter and the velocity distribution of the electrons.

621.385.3.029.6 4358

New Method of Calculating the Admittance of Disk-Seal Triodes under Small-Signal Conditions—I. Vágó and M. Uzsoky. (*Period Polyt. Bp.*, *Elec. Engrg*, vol. 5, no. 2, pp. 107–127; 1961. In German.) The solution is obtained by a variational method and given in operational form applicable to any type of voltage function.

621.385.5.032.265:621.376.23 4359

Synchronous Detector using the 7360 Beam-Deflection Tube—A. Sobel. (*Rev. Sci. Instr.*, vol. 32, pp. 867–868; July, 1961.) The advantages of the detector include cancellation of drift due to changes in cathode emission, and the elimination of balancing transformers or phase inverters.

621.385.6:621.372.8 4360

On the Theory of a Microwave Electron Device with Interaction in the Transverse Direction—L. N. Loshakov. (*Radiotekhnika i Elektronika*, vol. 5, pp. 1448–1457; September, 1960.) Results of a preliminary theoretical investigation of an electron-beam waveguide system in which the beam moves in a direction perpendicular to the power flow.

621.385.62.072.6 4361

An Arrangement for the Frequency Stabilization of a Klystron Oscillator—E. Pietzsch. (*Tech. Mitt. BRF, Berlin*, vol. 5, pp. 37–41; March, 1961.) A mechanical wobbulator system and discriminator amplifier for stabilization at 4 Gc are described.

621.385.623.5:621.376.3 4362

Frequency Modulation of a Reflex Klystron with Minimum Incidental Amplitude Modulation—W. R. Day, Jr. (PROC. IRE, vol. 49, pp. 1449–1450; September, 1961.)

621.385.624 4363

Finite-Gap Klystrons—L. Solymar. (*Electronic Tech.*, vol. 38, pp. 373–377; October, 1961.) The amplitudes of the fundamental and harmonic components of current are calculated using a large-signal ballistic theory. For large signals, the dependence on gap width is altered but gap admittance is invariant.

621.385.63 4364

Contribution to the Theory of Minimum Noise Figure of Low-Noise Travelling-Wave Tubes—K. B. Niclas (*Arch. elekt. Übertra-*

gung, vol. 15, pp. 101–107; February, 1961.) Conditions for obtaining the minimum noise figure are formulated; they are of a more general character than those given by Bloom and Peter (3402 of 1954). A method of measuring noise parameters is described with an example of its application at 3 Gc.

621.385.63:621.375.9:621.372.44 4365

Microwave Amplification in Electrostatic Ring Structures—J. C. Bass. (Proc. IRE, vol. 49, pp. 1424–1425; September, 1961.) Amplification by means of axially symmetric electrostatic rings with an axial magnetic field is described and experimental results are given.

621.385.63:621.396.65 4366

Travelling-Wave Amplifiers—T. Berceli. (*Period. Polyt., Bp., Elec. Engrg*, vol. 5, no. 1, pp. 75–91; In German.) Traveling-wave amplifier equipment for the final stage of a radio-link installation is described. A design method for the helix coupling system is given which takes account of the effect of screening.

621.385.632 4367

Experimental Study of a Backward-Wave Tube with a Nonuniform Delay System—L. P. Lisovskiĭ. (*Radiotekhnika i Elektronika*, vol. 5, pp. 1442–1447; September, 1960.) An experimental investigation of a backward-wave tube with delay line in the form of a double comb, the teeth of which decrease uniformly in height along the axis of the tube. A variation of 2.25 per cent in tooth height over one wavelength resulted in a 10-fold increase in restarting current required.

621.385.69:621.372.2 4368

Feedback-Type Distributed-Amplifier Tube—T. Kojima. (*Rev. Elec. Comm. Lab., Japan*, vol. 9, pp. 150–164; March/April, 1961.) The theory and design details are given for a tube in which the anode and grid form a double helix. Part of the amplified signal from the anode is fed back to the grid to compensate for line loss at high frequencies. The bandwidth of an experimental tube was 0–300 Mc. See also 3208 of September.

621.387:621.362 4369

Optimization of Emission-Limited Thermionic Generators—Schoek. (See 4265.)

621.387.322:621.316.722 4370

Comparison of Argon, Krypton and Xenon as Admixtures in Neon Glow-Discharge Reference Tubes—F. A. Benson and G. P. Burdett. (*Proc. IEE*, pt. B, vol. 108, pp. 501–507; Discussion, pp. 507–510; September, 1961.) Certain desired tube characteristics can frequently be obtained by correct choice of gas mixture.

621.387.322.2 4371

The Stability of Reference Tubes using Monocrystalline Cathodes—F. A. Benson and B. Rigg. (*Electronic Eng*, vol. 33, pp. 524–525; August, 1961.) The Mo crystal slices used are cut to expose the (1, 0, 0) face and the tubes are filled with Ne+0.3 per cent A at 45-mm-Hg pressure. Performance is superior to that of polycrystalline types.

MISCELLANEOUS

001.891:621.396 4372

Radio Research 1960: The Report of the Radio Research Board and the Report of the Director of Radio Research—[(Book Review) Publishers: H. M. Stationery Office, London, 23 pp., 2s.6d.; 1961.] (*Nature*, vol. 191, pp. 1255–1256; September 23, 1961.)

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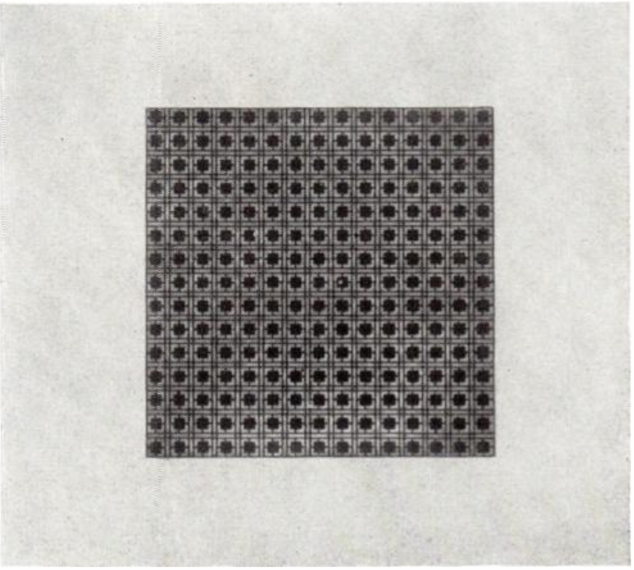

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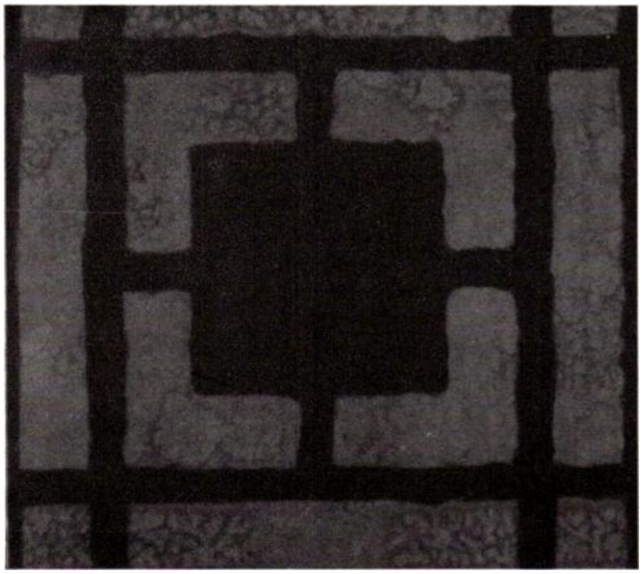

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Manufacturing researchers examine manufacturing processes before products are developed. Recently, when development of micromodule circuitry started, manufacturing researchers studied possible interconnection techniques. This analysis suggested an arrangement of laminated circuit planes which, combined with a standard matrix, permitted automated production of complex, miniaturized interconnection packages. This interconnection technique makes and breaks standard conductor patterns to form the circuits required on individual pieces of hardware. Specialists in fields as diverse as physical chemistry, metallurgy, and hydraulic servo-systems all contributed to its development.

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(Continued on page 85A)

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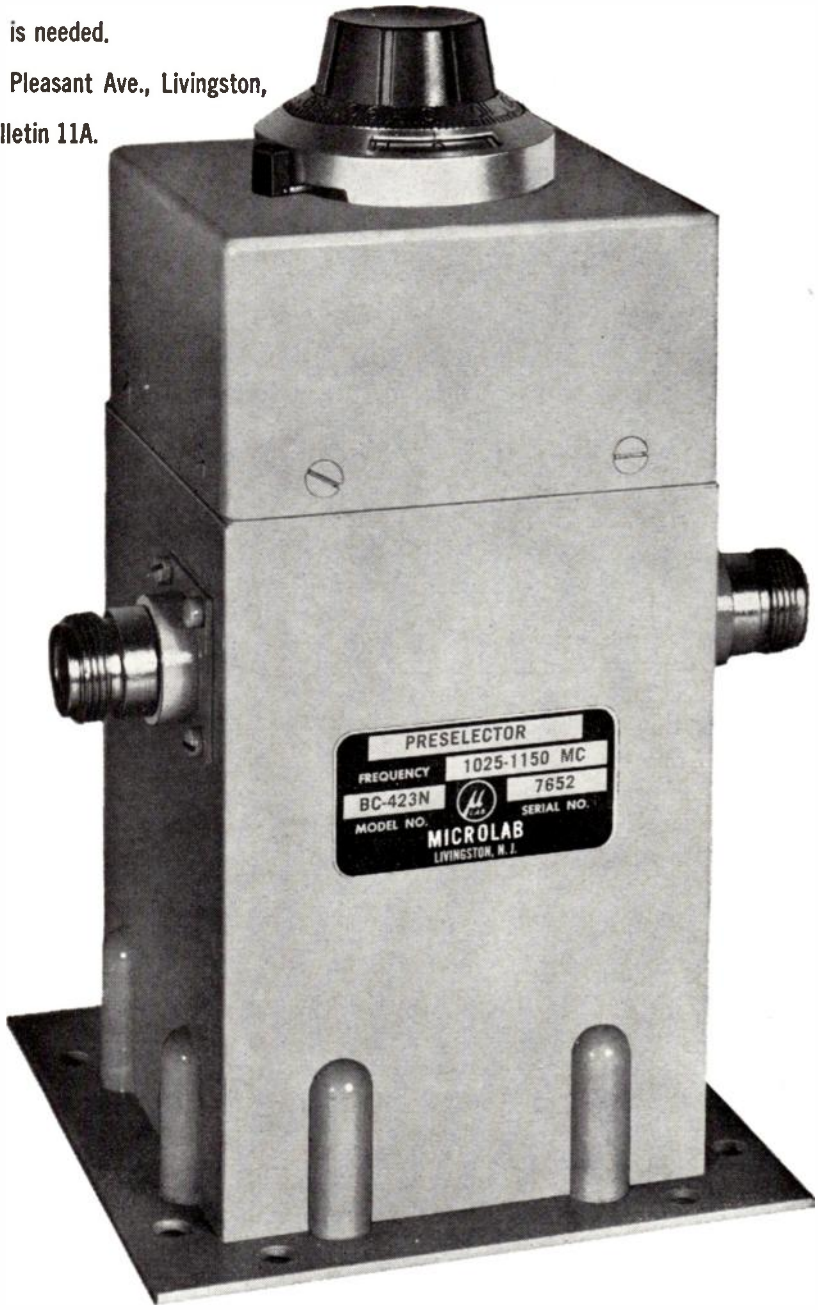

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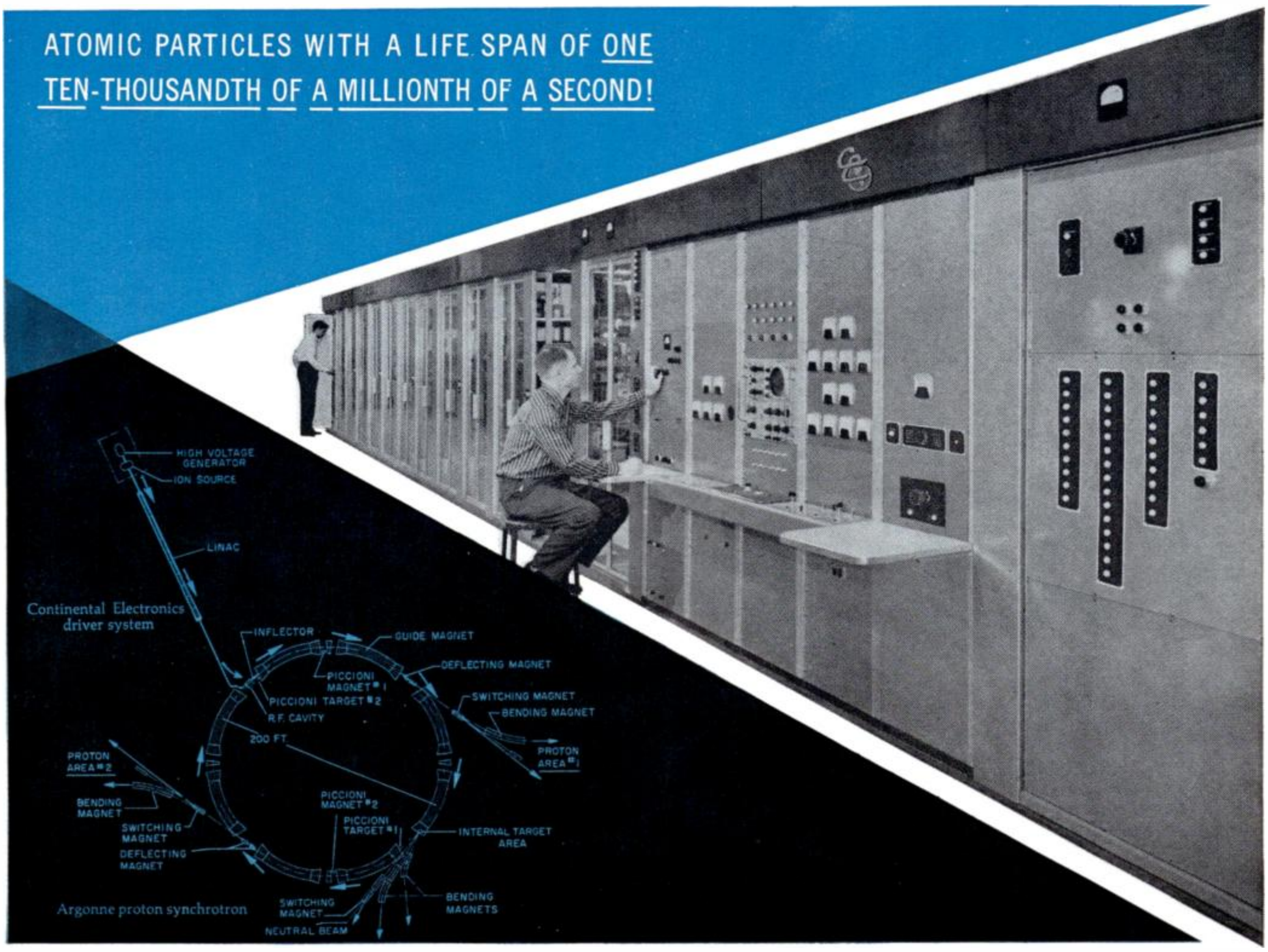

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Professional Group Meetings

AEROSPACE AND NAVIGATIONAL ELECTRONICS

Oklahoma City—October 9

"Telemetry Equipment and Transistorized Electronic Organs—Made by Dorsett," D. Clotz, Dorsett Electronics Inc., Norman, Okla.

ANTENNAS AND PROPAGATION

Boston—October 25

"Twelve-Horn Monopulse Antenna," L. J. Ricardi, Lincoln Lab., Lexington, Mass.

"Hyperbolic Direction Finding with Artificial Sferics," Dr. E. A. Lewis, AFCRL, Bedford, Mass.

Los Angeles—September 21

"Spatial Information Available from Monopulse Radar," D. B. Anderson, No. American Aviation (Autonetics) Anaheim, Calif.

San Francisco—October 11

"The Benelux Cross Antenna Project," C. L. Seeger, Stanford Radio Astronomy Inst., Stanford Univ., Stanford, Calif.

Washington, D. C.—October 17

"Fabric (inflatable) Antennas and Their Applications," C. C. Jones and A. T. Hawthorne, Westinghouse Electric Corp., Baltimore, Md.

ANTENNAS AND PROPAGATION MICROWAVE THEORY AND TECHNIQUES

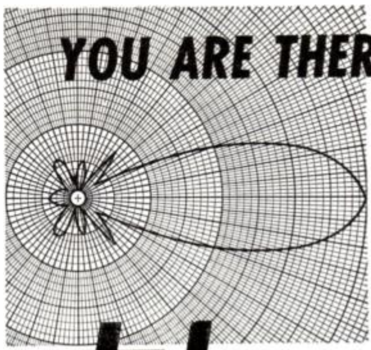
Baltimore—November 8

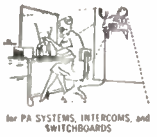
"Recent Research in Frequency Independent Antennas at the University of Illinois," J. Dyson, Univ. of Ill.

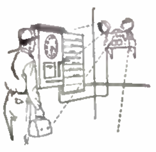
Columbus—October 18

"Signal Processing Antennas," Dr. R. C. Hansen, Aerospace, Inc., Los Angeles Calif.

(*Continued on page 86A*)







Professional Group Meetings

(Continued from page 85A)

AUDIO

Boston—September 28

"FM Stereo Transmission," R. L. Kaye, WCRB-FM, 750 South St., Waltham, Mass.

Chicago—October 13

"Stereo Geometry," P. W. Klipsch, Klipsch & Associates, Hope, Ark.

San Diego—September 6

"Limitations to Radio Communications in Noise," J. C. Webster, Naval Electronics Lab., San Diego, Calif.

San Francisco—November 1

"A New Battery-operated Transistorized Audio Oscillator and a New Transistorized Wave Analyzer," D. Cochran, Hewlett Packard Co., Palo Alto, Calif.

San Francisco—October 17

"Acoustical Considerations in the Design of Sound Studies,"

"A Very Directional Condenser Microphone" M. Rettinger, RCA Eng'g. Dept., Hollywood, Calif.

San Francisco—September 13

"Multiplex Simplified," R. Greisinger, Sargent-Rayment Co., Oakland, Calif.

AUTOMATIC CONTROL

Baltimore—October 10

"Process Control," N. B. Nichols, Taylor Instrument Co., Rochester, N. Y.

Chicago—October 13

"The VOLSCAN Automatic Voice Relay," T. Dunsheath, Cook Research Labs., Morton Grove, Ill.

Long Island—May 13

"Field Trip—Inspection of Totalizator and Pari-Mutuel Systems," Courtesy American Totalizator Co., & N. Y. Racing Assn. (Aqueduct Race Track)

Los Angeles—October 10

"Model Feedback Applied to Flexible Booster Control," W. K. Waymeyer, Douglas Aircraft, Santa Monica.

Pittsburgh—October 11

"Optimizing Control of Physical Systems," Dr. I. Lefkowitz, Case Inst. of Technology, Cleveland, Ohio

Schenectady—April 25

"Navigation Theory for Interplanetary Flight," Dr. R. Battin, MIT, Cambridge, Mass.

Schenectady—February 14

"Chemical Plants and Processes; a Challenge for Control Engineering," W. Kipiniak, MIT, Cambridge, Mass.

AUTOMATIC CONTROL INSTRUMENTATION

Long Island—March 7

"The Automatic Control and Instrumentation of Hovering Vehicles," E. J. Knight, Natl. Research Associates, College Park, Md.

BIO-MEDICAL ELECTRONICS

Baltimore—October 30

"Some Electrical Properties of Nerve Fibers of Engineering Interest," Dr. L. J. Mullins, Univ. of Maryland, School of Medicine, Baltimore

Chicago—October 13

"Measuring and Monitoring in Heart Surgery," Dr. P. V. Moulder, Jr., Univ. of Chicago, Billings Hospital, Chicago, Ill.

Columbus—September 21

This was an organizational meeting for the Chapter's Steering Committee. The committee outlined the year's forthcoming activities and announced the group's program for the next three months.

Los Angeles—October 19

"Physiological Monitoring—Aerospace and the Modern Hospital," Dr. J. T. Burroughs, VA; C. C. Miller, Jr., ITT Federal Labs., R. Harper, ITT Federal Labs.

Memphis—October 20

"Computers in Medicine," Dr. D. Wall, IBM assigned to Natl. Institutes of Health, Washington, D. C.

Omaha-Lincoln—October 26–27

"The Federal Interest in Biomedical Engineering,"—in absentia, B. Jones, read by R. L. Bowman, Natl. Institutes of Health, Bethesda, Md.

San Francisco—October 17

"Computer Simulation and Mathematical Analysis of Non-Linear Biological Systems," J. G. Defares, M.D., University of Leyden, Holland.

Albuquerque-Los Alamos—November 9

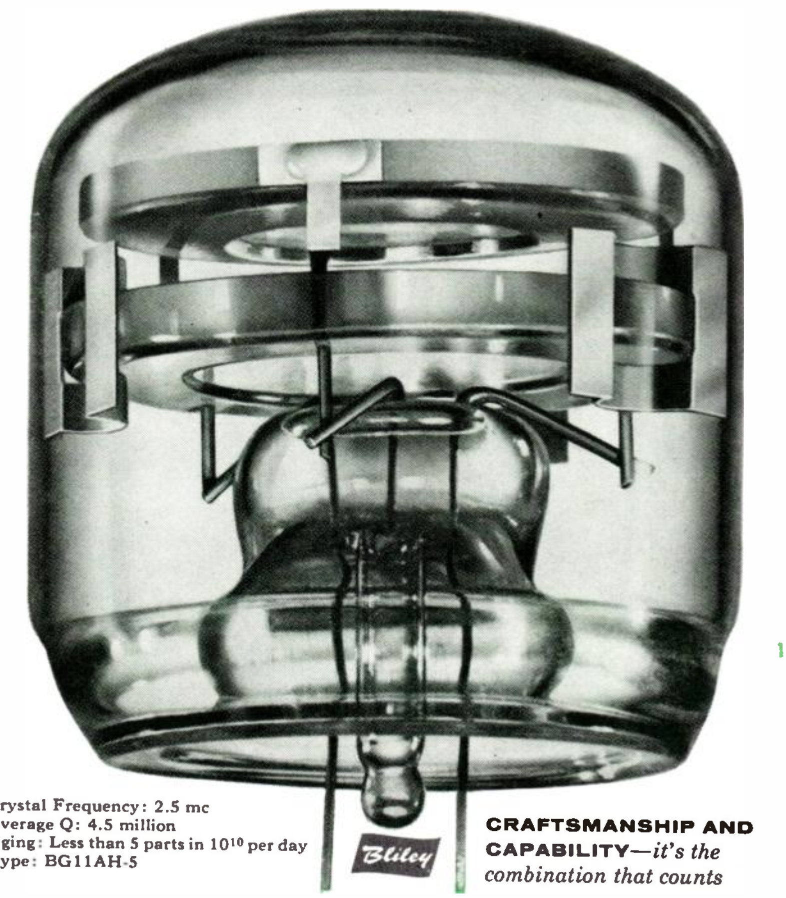
"Foundations of Circuit Theory," Dr. S. Karni, Univ. of New Mexico

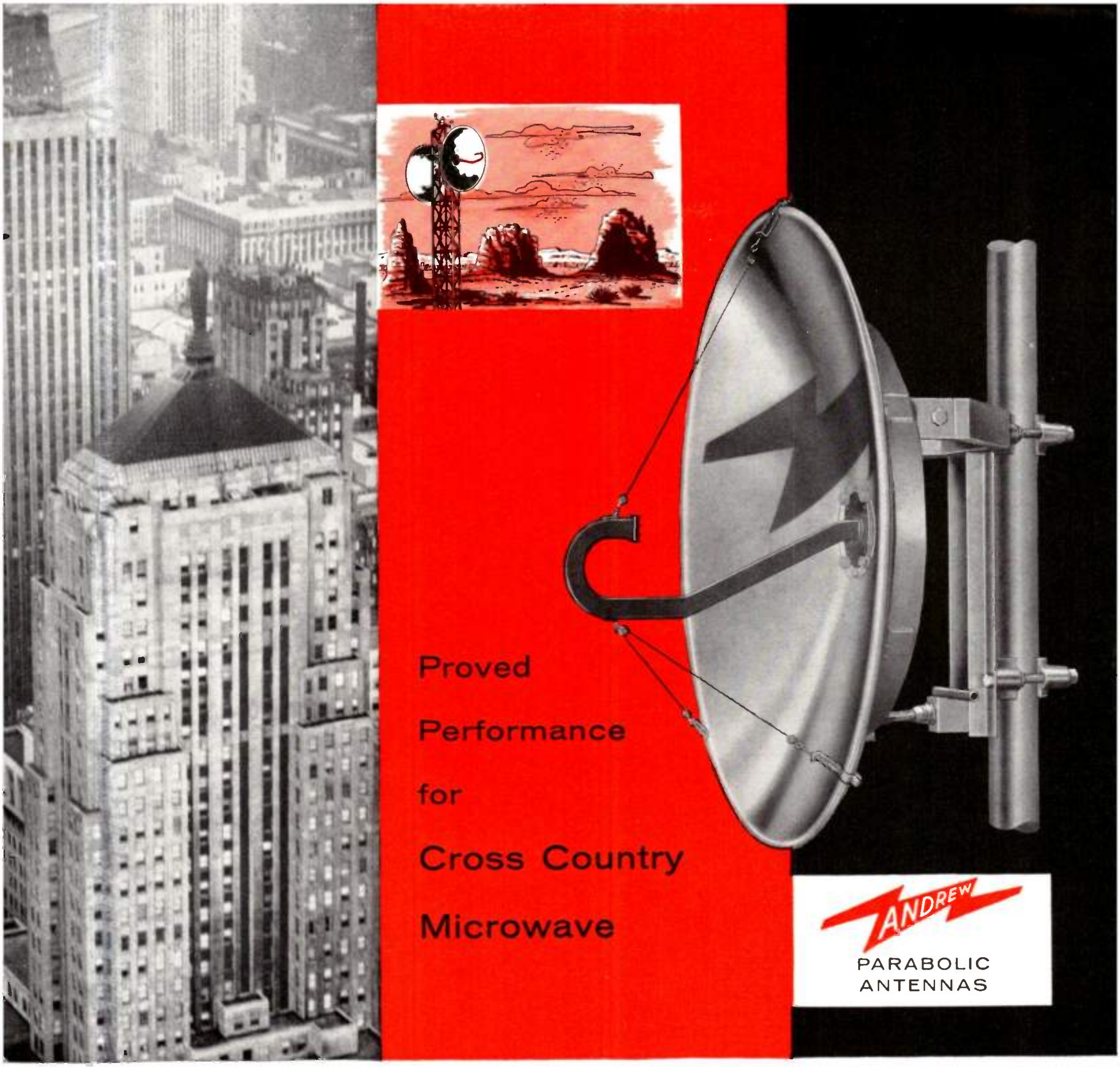
CIRCUIT THEORY

Syracuse—September 21

"Impressions of the Prague Conference on Circuit Theory," Dr. N. Balabanian, Syracuse Univ., Syracuse, N. Y.

(Continued on page 100A)





This busy metropolitan area is the termination of over 1000 miles of microwave systems, providing reliable communications across town and country for the Western Union Telegraph Company. ANDREW's experience in research, development and manufacturing is the reason why the dependable performance of an ANDREW P8-37, eight-foot parabolic antenna was selected for this installation.

All ANDREW parabolic antennas conform to the newly proposed EIA-FCC standards governing radiation patterns and side lobes, and they are *guaranteed* to give specified pattern and VSWR in your microwave system.

From over sixty standard parabolic antennas, you can choose the type and size that will give optimum system performance with absolute mechanical and electrical reliability.

Microwave engineers have found ANDREW a valuable partner in planning communications systems. A computer is available upon request for calculating antenna system performance. The recently published Catalog M presents complete engineering specifications in a wide selection of parabolic antennas serving the microwave frequencies. Write for catalog or call local ANDREW application engineer.

NEW ANDREW CATALOGS
for Communication Engineers

Catalog M — Microwave Antennas
Catalog R — Rigid Transmission Lines
Catalog H — Heliax — Coaxial Cables
Catalog T — Telemetry Antennas

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Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

CREATIVE ENGINEER

Nation's leading manufacturer of high speed solid state counting, timing and frequency measuring instrumentation seeks imaginative young engineer to design and develop digital and pulse instruments and systems. Outstanding opportunity for technical growth and advancement.

One to two years' experience in transistor application of digital and pulse techniques plus BSEE or equivalent are required. Send resume to: Personnel Manager, Computer Measurements Company, 12970 Bradley Avenue, San Fernando, California.

PRODUCTION MANAGER—ELECTRONIC EQUIPMENT

Require a man with at least 10 years experience in direct management of male and female assemblers, wiremen and testers, working on both military and commercial electronic equipment.

Must be familiar with military specifications for workmanship and know cost cutting techniques. Should be experienced in estimating labor hours and maintaining budgeted hours via strict production control.

This man will run the entire wiring and assembly operation (up to 100 people) and will be a part of the top management team. Salary open. Contact George W. Fellendorf, President, Planetronics, Inc., Easton, Pennsylvania.

TECHNICAL WRITER

Technical Writer to work with scientific, engineering and public relations personnel. BS in engineering or physics. Nuclear experience desirable. Send resume to Mgr. of Employment, General Atomic Div., General Dynamics Corp., P.O. Box 608, San Diego 12, Calif.

ASSISTANT PROFESSOR

Ph.D.s required with interests in circuits, electromagnetic theory or controls, in expanding Graduate and Undergraduate School. Applications from abroad invited. Apply to Chairman, Department of Electrical Engineering, Villanova University, Villanova, Pa.

ASSISTANT or ASSOCIATE PROFESSOR

Assistant or Associate Professor with Ph.D. in Electrical Engineering or Physics. Interest in Electronics, Microwaves, Solid State, Circuit Theory, or Computers. Active research and graduate program with opportunities for outside consulting. May start February 1 or September 1, 1962. Send resume to W. P. Smith, Chairman Electrical Engineering Department, University of Kansas, Lawrence, Kansas.

ELECTRONICS ENGINEER

Critical vacancy in the position of Electronics Engineer, GS-11, $7560 per annum. Contact Industrial Relations Department, U. S. Naval Air Station, Quonset Point, R. I.

(Continued on page 90A)

USAF/AWCS-412-L Consisting of a closely coordinated network of data acquisition stations, data processing and display centers and weapon bases, **Air Weapons Control System 412-L** provides the tools for effective and flexible air space management, continent-wide or in single point defense. Vital detection and tracking information is supplied automatically to human decision-makers within seconds. Effective direction of both manned and unmanned weapons, including return of manned aircraft to base, is a system function.

412-L STAFF EXPANDING

as program moves ahead in advanced development phase

If you are one of the fast-growing group of scientists and engineers who recognizes the unequaled career advantages and satisfactions to be gained through work on very large-scale, complex systems, we urge you to scan this abbreviated list of current openings on 412-L. If your background, experience or interests are represented, write us today. We promise you a careful review of your qualifications and a prompt reply.

COMMUNICATION SYSTEMS

Training in information theory and switching theory; broad experience in both RF and wire communication systems design, especially automatic subscriber and dial systems; to determine and specify characteristics for intra- and inter-site command and control complexes.

OPERATIONS ANALYSIS

Involves synthesis/analysis of computer oriented weapons control systems (ground/air environments) from an operational standpoint. Estimates system (personnel & equipment) capabilities; derives and evaluates system procedures using analyses and computer simulation as tools. BS/MS – minimum 3 years experience.

EQUIPMENT EVALUATION

Solving man-machine problems, evaluating alternative components, displays, or techniques, devise simulators. Advanced degree in experimental psychology required.

An Equal Opportunity Employer

SYSTEM EQUIPMENT ANALYSIS

Performs analysis on system-equipments and their functions; derives criteria and parameters to meet operational requirements. Work in areas of signal processing, detection, tracking and digital communications. Requires BS/MS/EE/Physics/Math—knowledge of simulation techniques, feedback theory, probability and information coding desirable. 3 years experience.

WEAPONS INTEGRATION

Determines and specifies optimum weapon utilization. Specifies weapon interfaces to equipment designers to insure system compatibility. Support / provide analytical inputs relative to weapons capabilities; analytical studies to assure optimum weapon employment within specific computer capabilities. BS/MS. 2 years manned/unmanned weapons system experience.

INFORMATION PROCESSING & DISPLAY SYSTEMS

Training in information and statistical theory; broad experience in design and integration of tactical and strategic data processing systems; to define system parameters, identify interface problems, perform math analyses of control loops as applied to real-time processing and display systems.

TECHNICAL WRITERS

Creation of systems manuals, engineering reports, proposals; required is at least 2 years of technical education with minimum of 5 years in engineering writing. Knowledge of graphic arts techniques would be helpful.

PROGRAMMING ANALYSIS

Degree required. Math background; 2 years experience on medium/large-scale digital computers solving real-time military operational problems. Activity ranges from problem analysis to program development, including flow diagramming, coding and debugging in the utility control, simulation and operation areas.

ANTENNA & MICROWAVE COMPONENTS

BSEE, advanced degree preferred; experience with high power phenomenon, antenna development, RF plumbing, modulators, transmitters and receivers.

APPLIED MATHEMATICS

Performs mathematical analyses on all phases of ground electronic control systems. Utilizes probability theory, numerical analysis, complex variables; knowledge and utilization of analog/digital computers desirable. BS/MS Math required; 2 years experience on applied math problems in engineering field.

Write in confidence to Mr. P. W. Christos, Div. 53-MA
DEFENSE SYSTEMS DEPARTMENT (A Department of the Defense Electronics Division),
General Electric, Northern Lights Office Building—Syracuse, New York

GENERAL ELECTRIC

Positions Open

(Continued from page 88A)

ENGINEER

$115 week. Approx. College grad. with a major in either electrical or mechanical engineering. Work involves investigating major appliances and building products in laboratory of national magazine. Prefer some experience but will consider recent college graduate. Veteran or Draft Deferred. Send typed resume to, Hearst Magazines, Personnel Office, 309 West 56 Street, New York 19, N. Y.

TELEVISION BROADCAST TECHNICIAN

Opportunities for video technician with group television stations. Write G. G. Jacobs, Corinthian Broadcasting Corp., 302 South Frankfort, Tulsa, Okla.

BIO-MEDICAL RESEARCH

The Department of Physiology and Pharmacology seeks a young man interested in biomedical research. Prefer a recent graduate with extensive background in circuit theory to assist in developing instrumentation for bio-medical research. Interested persons should communicate with: Albert F. Kelso, Ph.D., Dept. of Physiology and Pharmacology, c/o Chicago College of Osteopathy, 1122 East 53 Street, Chicago 15, Ill.

BROADCAST ENGINEERS

The United States Information Agency seeks Electronics Engineers with Communications or Broadcasting Systems experience for its Voice of America. The positions are located in Washington, D. C. and abroad. Domestic assignments are mainly concerned with the design and planning of broadcasting systems and transmitting plants up to 1000 kilowatts.

Foreign positions are of a construction management type.

Broadcasting and Communications Systems cover frequency ranges in the medium and high frequency bands.

Foreign Service appointments will be made at the entrance level of $10,785 per annum and range to $12,025 plus overseas allowances. Appointment levels for domestic service: $8955, $10,635, $12,210, $13,730 and range to $15,030 per annum commensurate with background and experience.

Relocation expenses paid in accordance with government regulations.

Applicants *must* be AMERICAN citizens.

Please send background resume or application for Federal Employment (SF-57) to: Employment Branch, U. S. Information Agency, Washington 25, D. C.

SENIOR ELECTRONICS ENGINEER

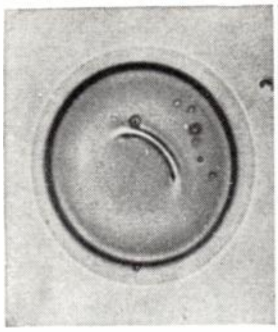
Excellent opportunity exists for professional advancement with small research organization. Seeking B.S.E.E. or M.S.E.E. with 4 to 10 years experience including design work in R. F. and microwave fields in the 30 to 3000 Mc regions. Interest in Project Management responsibilities desired. Contact Richard A. Johnson, Vice President, Systems Research Laboratories, Inc., 500 Woods Drive, Dayton 32, Ohio. Phone 244-4051.

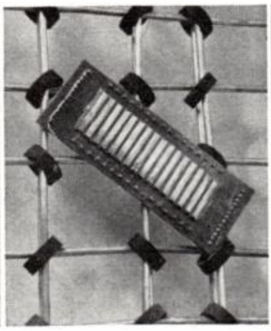
MAGNETIC CORE COMPONENTS

Electrical Engineer with design experience involving components for pulse circuitry with emphasis on filters, R. F. coils and pulse transformers. A background involving the uses of fer-

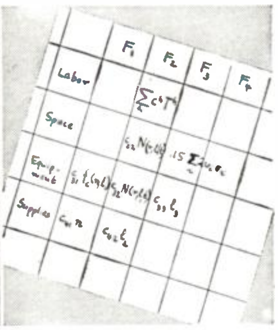
(Continued on page 94A)

NCR ...Research


MICRO-ENCAPSULATION

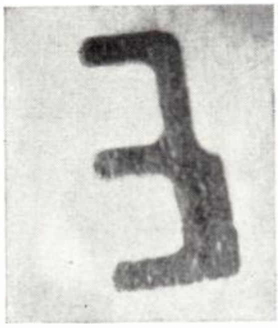

MAGNETICS

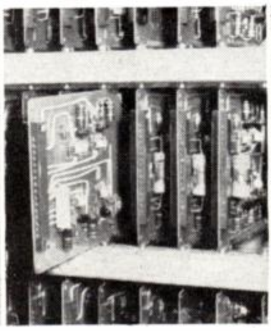

THIN FILMS



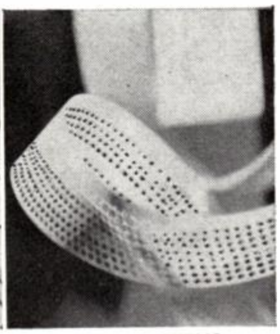
OPERATIONS

...Development


OPTICS

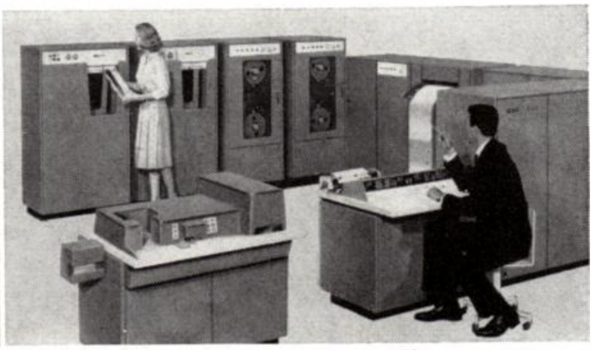

LOGIC & CIRCUITRY

MECHANICS


DIGITAL SYSTEMS

...New Products


THE NCR 450 CONTROL AND PROOF MACHINE


THE NCR 315 COMPUTER

NCR offers a wide range of opportunities for experienced scientists and engineers interested in commercially oriented projects or advanced military electronic development. Current work encompasses a broad field of activity including semiconductor research, low temperature physics, micro-electronics, photochromics, advanced electronic systems development, high frequency communications, and operations research studies related to complex business and financial systems. Previous efforts of NCR research and development have resulted in new products such as: NCR (no carbon required) Paper, the highly successful Electronic Data Processing Systems, and other items for the electronic and business machine industries. Rapid growth has caused openings at all levels of experience and education and provide opportunity for technical or administrative growth.

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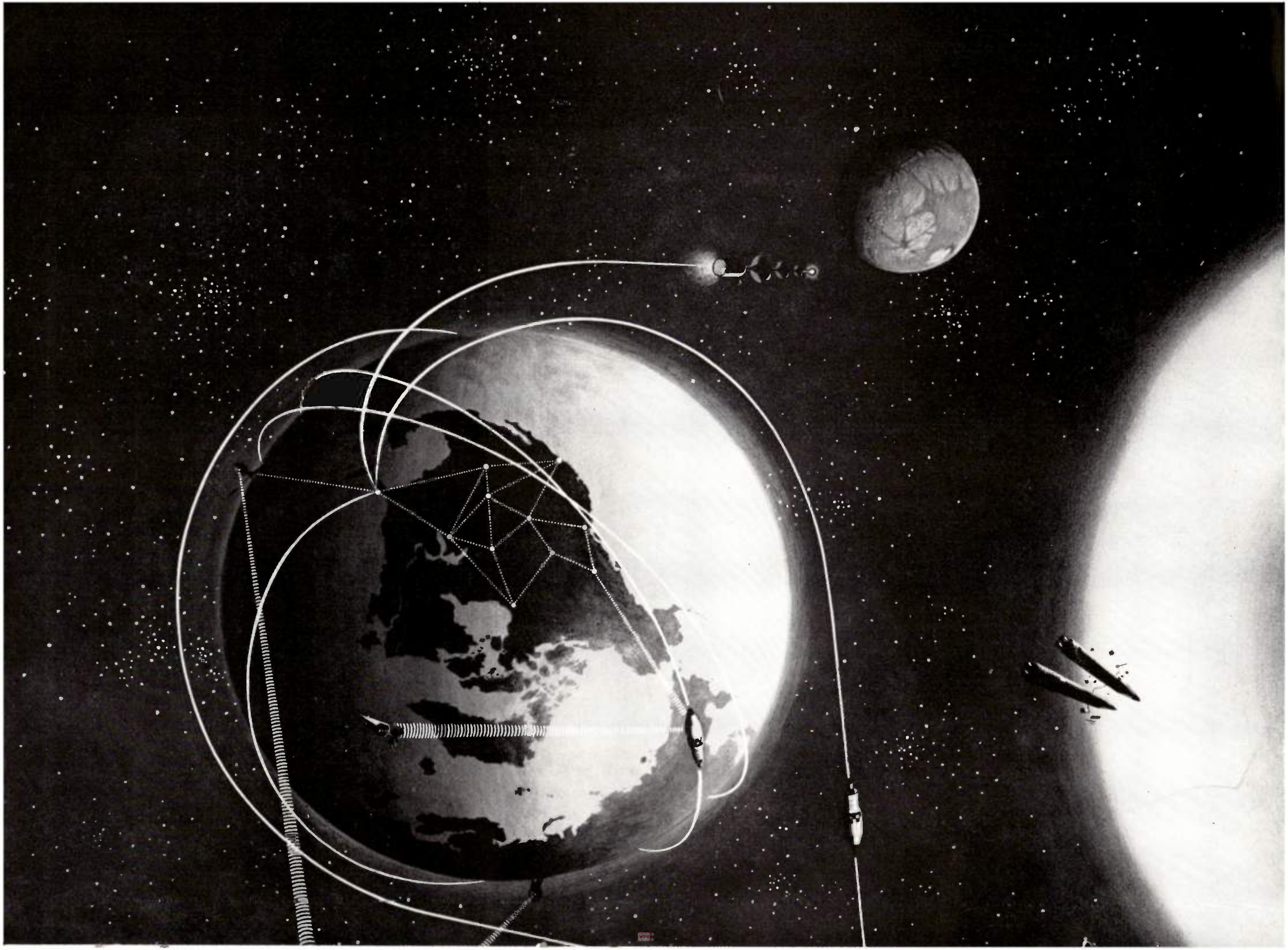

SATELLITE AND SPACE SYSTEMS SPECIALISTS

Here at Lockheed Missiles and Space Company, satellites and spacecraft are a specialty. From research to the reaches of space, the whole scope of space technology is being carefully investigated.

All aspects of research, systems analysis, design, development and operation are handled by outstanding engineers and scientists at Lockheed's Research Laboratories in Palo Alto, and in the development headquarters in nearby Sunnyvale, California.

Typical of Lockheed's complete capability is the DISCOVERER series. This, with its recoverable capsule, is used to gather research material. Other military and commercial satellites and spacecraft under study, under development, or in operation, include:

An infrared missile defense alarm system • Sophisticated orbiting biomedical capsules • Lunar probes • Interplanetary exploration programs
A space rendezvous system • Nuclear and other advanced propulsion systems
Communications satellite systems

It is clear that the projects at Lockheed Missiles and Space Company are challenging. Moreover, its location on the beautiful San Francisco Peninsula adds gracious living and perfect climate to the many rewarding opportunities available to creative engineers and scientists.

For further information, please write: Research and Development Staff, Dept. M-32C, 599 Mathilda Avenue, Sunnyvale, California.
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A GROUP DIVISION OF LOCKHEED AIRCRAFT CORPORATION

Systems Manager for the Navy POLARIS FBM and the Air Force AGENA Satellite in the DISCOVERER and MIDAS programs. Other current programs include SAINT, ADVENT and such NASA projects as OGO, OAO, ECHO, and NIMBUS.

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Positions Open

(Continued from page 90A)

rite and powdered iron cores is highly desirable. Send complete resume and salary requirements with first letter to Paul E. Dicker, Chief Engineer, Aladdin Electronics, Nashville 10, Tennessee.

ELECTRONIC MUSICAL INSTRUMENT ENGINEER

Graduate Electrical Engineer to join research group on electronic musical instruments. One to three years electronic experience required, with some musical background desirable. Write Albert Meyer, Engineering & Research Dept., The Baldwin Piano Co., 1801 Gilbert Ave., Cincinnati 2, Ohio.

EDUCATION AND RESEARCH IN BIO-MEDICAL ENGINEERING

An educational program designed to prepare life and physical scientists, doctors of medicine, and engineers for research in life systems through the current period of transition of biological sciences from the natural history stage of inquiry by inductive empiricism to that of inquiry by deductively formulated theory. An intensive four-term (12-month) preparation in methods of logical thought, mathematics, physics, and electrical theory for doctors of medicine and biologists, and one-term (3-month) preparation in life sciences for engineers, makes possible joint course work in the final three terms (nine months). Thesis research on biological systems is conducted at the Presbyterian Hospital by teams drawn from both disciplines. The program leads to a graduate degree at Drexel. Limited stipend and scholarship funds are available. Assistance with fellowship application to National Institutes of Health is offered. For information, James W. Dow, M. D., Dept. of Bio-medical Engineering, Drexel Institute of Technology, 15 North 32 St., Philadelphia 4, Pa.

ANTENNA ENGINEERS

Experienced Antenna Engineers, Physicists, B.S., M.S., Ph.D. Citizens of U. S. A. Very challenging theoretical, applied research and engineering problems on ground, missile, satellite and deep space probe antenna systems. Write H. W. Haas, Physical Science Laboratory, New Mexico State University, University Park, New Mexico.

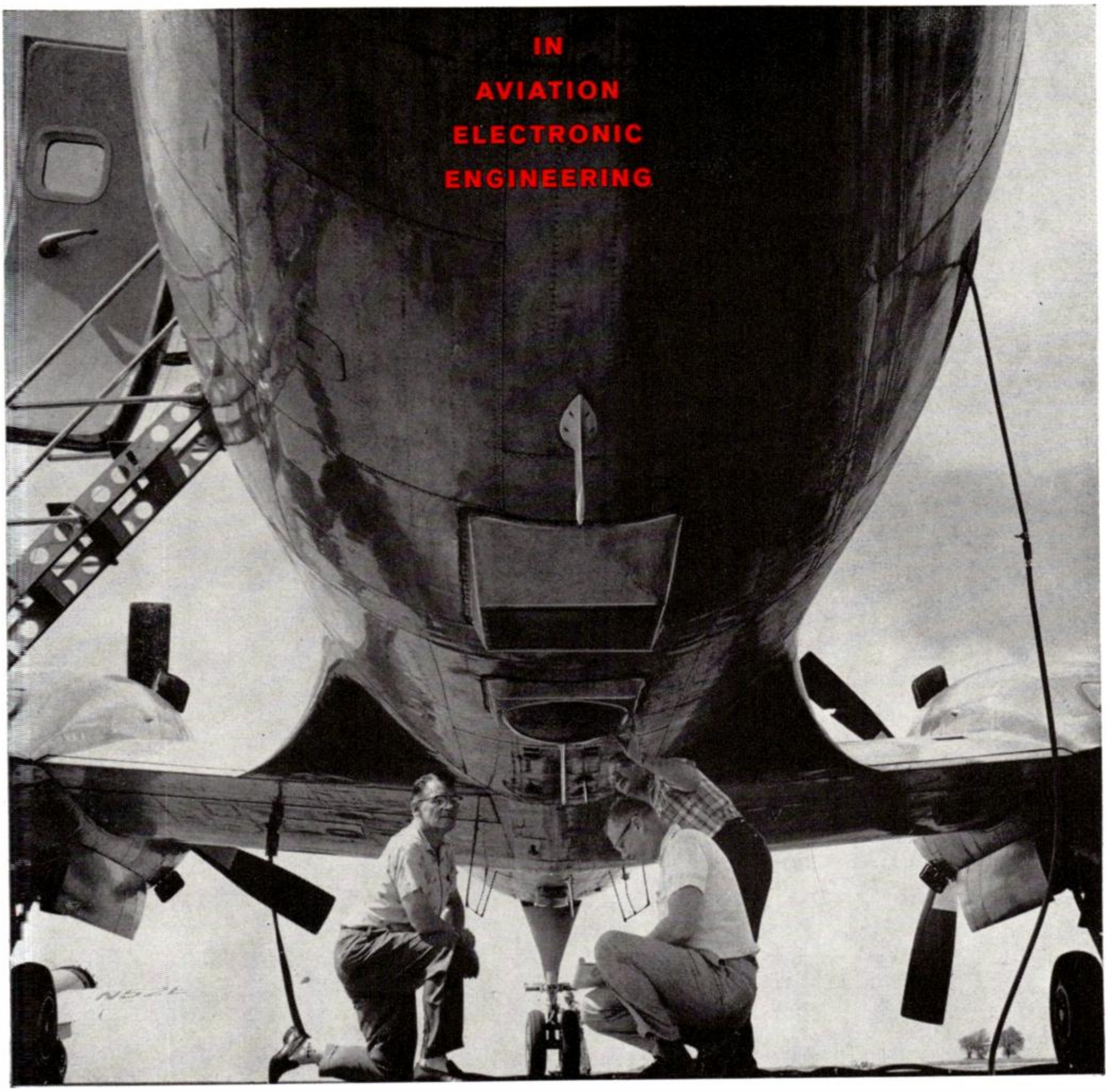
OVERSEAS

Robert College, in Istanbul, Turkey, presents a challenge in education where East meets West. An opportunity to contribute significantly to the development of a young republic is available to specialists in engineering, business administration and economics, the sciences, and the humanities. Graduate degrees required. Address inquiries to Dr. Howard P. Hall, Vice President, Robert College, Bebek Post Box 8, Istanbul, Turkey; with copy to the Near East College Association, 548 Fifth Avenue, New York 36, N.Y.

AUDIO ENGINEER

BSEE or BS Physics. 3 to 5 years experience in the design and development of moving coil microphones and/or miniature dynamic speakers. Manufacturing experience desirable. Progressive company manufacturing electroacoustic transducers in the Metropolitan area. Salary competitive. Send complete resume to Box 2062.

(Continued on page 96A)


IN
AVIATION
ELECTRONIC
ENGINEERING

Positions Open

(Continued from page 94A)

PROFESSORS

Ph.D.'s required in electrical engineering (both new and those established in their field) for a rapidly expanding and state supported urban University with a strong Graduate School. Outstanding opportunities in growth and salary—all located in an excellent total environment involving a research-based community in New York State. Box 2064.

PROFESSOR

Ph.D. wanted to head E. E. Department in metropolitan New York school, September 1962. Educational & industrial experience desired. Excellent opportunity to participate in teaching, research & growth of department. Write box 2065.

TEACHING-ELECTRICAL ENGINEERING

Electrical Engineering staff openings in an expanding undergraduate program. Candidates should possess graduate degrees (M.S. or Ph.D.). Encouragement is given to private consultation and research. Rank and salary commensurate with academic attainment and experience. Apply: Wm. Shewan, Chairman, Dept. of E. E., College of Engineering, Valparaiso University, Valparaiso, Indiana.

SCIENTISTS AND ENGINEERS

For JPL's Lunar and Planetary Programs with experience in any of the following areas: Celestial Mechanics, Orbit Determination, Space Navigation Theory, Trajectory Studies, and Systems Analysis. A new technology, based on celestial mechanics, mathematical statistics, and engineering systems analysis, is evolving in the space exploration program. Both theoretical and project positions are available for experienced people in this expanding field. Send resume to P. I. Locke, Jet Propulsion Laboratory, California Institute of Technology, 4800 Oak Grove Drive, Pasadena, California. An equal opportunity employer.

ENGINEERS

The Federal Aviation Agency is filling three engineer positions:

1. General Engineer (Aircraft Electrical/Electronic) GS-13, $10,635 per annum.
2. Aeronautical Design Evaluation Engineer GS-12, $8955 per annum.
3. Aeronautical Power Plant Design Evaluation Engineer (Engines) GS-12, $8955 per annum.

Incumbents will serve as technical specialists in the Engineering and Manufacturing Branch, Flight Standards Division. Some travel is required. Call Jamaica, N. Y. OL 9-7000 ext. 209 for further information.

ELECTRONICS INSTRUCTORS

Permanent positions as radio and electronics instructors for energetic young men up to 30 years old. Excellent and unlimited pay for persons with the right desire and qualifications. College definitely preferred. Excellent advancement opportunities. Send complete resume to Elkins Institute of Radio and Electronics, 2603 Inwood Road, Dallas 35, Texas.

PROFESSOR

Professor of Engineering to teach electronics and work in the development of undergraduate and graduate programs. Ph.D. and teaching experience in the field are required. Also, Assistant

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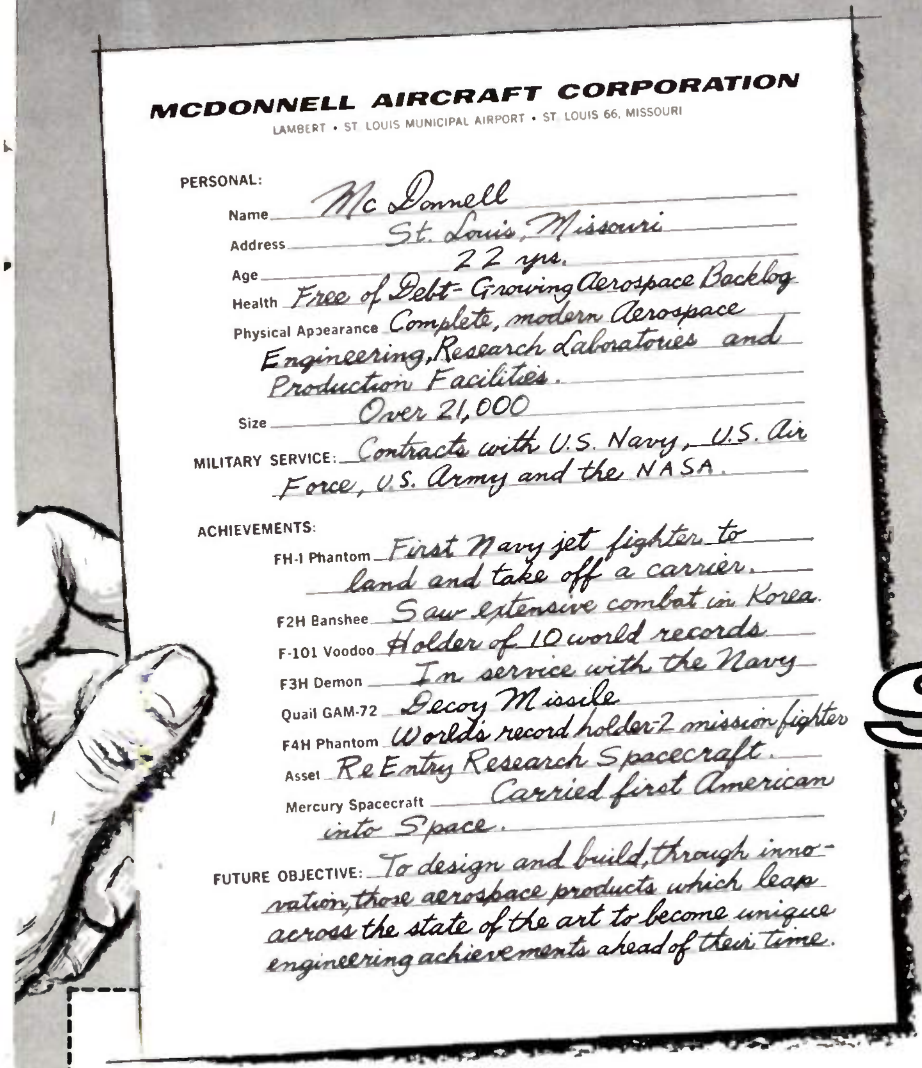

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FH-1 Phantom First Navy jet fighter to land and take off a carrier.

F2H Banshee Saw extensive combat in Korea.

F-101 Voodoo Holder of 10 world records.

F3H Demon In service with the Navy

Quail GAM-72 Decoy Missile

F4H Phantom World's record holder-2 mission fighter

Asset ReEntry Research Spacecraft.

Mercury Spacecraft Carried first American into Space.

FUTURE OBJECTIVE: To design and build, through innovation, those aerospace products which leap across the state of the art to become unique engineering achievements ahead of their time.

ENGINEERS SCIENTISTS PHYSICISTS

LET'S SWAP RESUMES

Please complete this form and forward to: Mr. D. F. Waters, Professional Placement, Dept. 62, McDonnell Aircraft, St. Louis 66, Missouri. *This is not an application for employment.* Your qualifications will be reviewed by our placement staff and you will be advised of positions at McDonnell for which you qualify. You may then make application if you wish.

Name__________ Address__________

City & State__________ Phone__________ Age_____

Primary Experience Area__________ Number of Years_____

Secondary Experience__________ Number of Years_____

Additional Comments__________

Education: AE____ ME____ Math____ Physics____ Chemistry____ EE____ Astronomy____ Other____

Degree: BS_____ Date MS_____ Date PHD_____ Date I would like to receive application form_____

I would like to receive literature about professional opportunities at McDonnell

My primary interest is:

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☐ Antenna & Microwave Techniques
☐ Automatic Test Equipment
☐ Computer Development
☐ Digital Technique & Logic Design
☐ Military Airborne & Solid State Electronics
☐ Research
☐ Thin Film Techniques
☐ Weapons Systems Analysis

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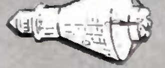

Phantom II Fighter and Attack Aircraft • Mercury, Asset and Aeroballistic Spacecraft • Talos and Typhon Missile Airframes and Engines • Quail Decoy Missiles • Rotorcraft • Electronic Systems • Automation

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Silver Springs, Maryland

Armour Research Foundation, a leading scientific and engineering organization, is now offering exceptional long term opportunities for electronic engineers to work in Silver Springs, Maryland.

The Foundation is currently forming a highly competent research team to make significant contributions toward the development and evaluation of an advanced radar system having search, track, and track-while-scan modes of operation. These openings afford a unique opportunity to work on an entire weapon system, along with opportunity for professional growth in the areas of development and analysis. To qualify you should have a B.S. degree in E.E., with an above average academic record, with a minimum of 2 years experience in circuit design, radar development, or system development and evaluation.

Staff members receive ATTRACTIVE SALARIES, up to 4 weeks vacation, generous insurance and retirement benefits, and tuition paid graduate study. *An equal opportunity employer.* Please reply in confidence to Mr. W. A. Massey at:

ARMOUR RESEARCH FOUNDATION

P. O. BOX 453
SILVER SPRINGS, MARYLAND

Positions Open

(*Continued from page 96A*)

Professor to teach electronics. M.S. required, Ph.D. desirable. Excellent opportunities in a rapidly growing college. Apply to the Head of the Engineering Department, Los Angeles State College, Los Angeles 32, California.

ASSISTANT or ASSOCIATE PROFESSOR

For January or September 1962, assistant or associate professor of Electrical Engineering required to teach undergraduate and graduate courses. Ph.D. required. Specialty networks, computors, or electromagnetic theory. Write to Dr. H. Mahrous, Electrical Engineering Dept., Pratt Institute, Brooklyn 5, New York.

ENGINEER-PHYSICIST FOR ELECTRO-THERMAL RESEARCH

Cornell Aeronautical Laboratory of Cornell University has several attractive openings for qualified Engineers or Scientists to participate in research in the thermodynamics of advanced electronic devices.

Projects include investigations of the thermal aspects of microminiature and new solid state electronic devices.

Appropriate experience in thermodynamics, cryogenics, solid state physics, and electronics required.

Contact E. P. Rentschler, Cornell Aeronautical Laboratory, Inc., 4455 Genesee Street, Buffalo 21, New York

CRITICAL VACANCIES AT GRIFFISS AIR FORCE BASE, ROME, N. Y.

Contract Assistant	GS-7	$5355 pa.
Technical Pub. Editor	GS-7	$5355 pa.
Management Analyst	GS-7	$5355 pa.
Research Psychologist (Physiological, Experimental & Engineering)	GS-7	$5355 pa.
Supvy. Engineering Data Specialist	GS-9	$6435 pa.
Librarian	GS-9	$6435 pa.
Industrial Engineer	GS-9	$6435 pa.
Electrical Engineer	GS-9	$6435 pa.
Research Psychologist (Physiological, Experimental & Engineering)	GS-9	$6435 pa.
Industrial Engineer	GS-11	$7560 pa.
Electrical Engineer	GS-11	$7560 pa.
Electronic Engineer	GS-12	$8955 pa.
General Engineer	GS-12	$8955 pa.
Mechanical Engineer	GS-12	$8955 pa.

COMPUTERS AND CONTROL SYSTEMS

- OPTICAL SPECIALISTS
- LOGIC DESIGNERS
- PROJECT ENGINEERS
- ADVANCED ELECTRONICS

Scientists and engineers needed with experience in all phases of digital computer and control system design.

Systems organization, logic design, magnetic core and drum memories, dynamic analysis, and electro-optical correlation devices. Also advanced areas such as high-speed tunnel-diode techniques, thin films, and hybrid analog-digital techniques. Applications include airborne digital equipment, numerical machine control, photogrammetric equipment, and special-purpose control computers. Both commercial and military programs, emphasizing advanced development and research. We think you will find this work unusually stimulating and satisfying. Comfortable and pleasant surroundings in suburban Detroit.

If interested, please write or wire A. Capsalis, Research Laboratories Division, The Bendix Corporation, Southfield, Michigan.

Research Laboratories Division

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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.

ACADEMIC POSITION

B.S. and M.S. in E.E.; present position Assistant Professor; 7 years teaching experience and 1 year training in Guided Missiles. Desires teaching position in Fall of 1962 with opportunity for graduate study. Box 3961 W.

ELECTRICAL ENGINEER

Professional E.E. degree (MIT). Four years experience in digital control system development, circuit design, logic design and programming. Speaks 4 foreign languges. U.S. Citizen. Seeks challenging position in U.S. or Europe. Box 3967 W.

HAWAII

Communications Systems Engineer. Data transmission, HF, VHF, UHF, microwave, common carrier; SSB, MUX, tropo and iono scatter, mobile, CCTV; Systems planning, design, procurement, installation, operation. Recent experience in high speed data transmission. Desires position in or leading to assignment in Hawaii. Available about first of 1962. Box 3968 W.

V.P. or GENERAL MANAGER

Age 42, BSEE. 24 years extremely diversified engineering, manufacturing and management experience with 5 major U.S. companies, in telecommunications, test ranges, aerial navigation and intelligence equipments and systems. Presently management consultant but will consider major responsibilities with smaller company or strictly autonomous division of larger one. East Coast (suburban area) preferred. Box 3970 W.

(*Continued on page 100A*)

Ph.D's, MS's, BSEE's

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opportunities
with

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Sylvania Mountain View, 40 minutes south of San Francisco, offers you challenging scientific work on defense systems, plus ideal living conditions.

The work involves systems studies, design and development in frequency ranges from DC to daylight; it offers growth into scientific or management positions of responsibility.

Immediate openings exist in the following fields:

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- Operations Research
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  - Antennas, Receivers, Transmitters, Transceivers, Servos
- Analog Computers
- Mechanical Design
- Field Engineering

Sophisticated San Francisco's theatres, restaurants and major league sports are minutes away. The Pacific Ocean is near; hunting, fishing and skiing 3 hours' drive.

Furthering your education? Sylvania encourages, sponsors graduate study at Stanford and other nearby institutions.

Send resumé to Wayne Pearson, Sylvania, Box 188, Mountain View, California

U.S. citizenship required.
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Government Systems Management
for ***GENERAL TELEPHONE & ELECTRONICS***

GENERAL SYSTEM

7468 C

RADIOTRACER

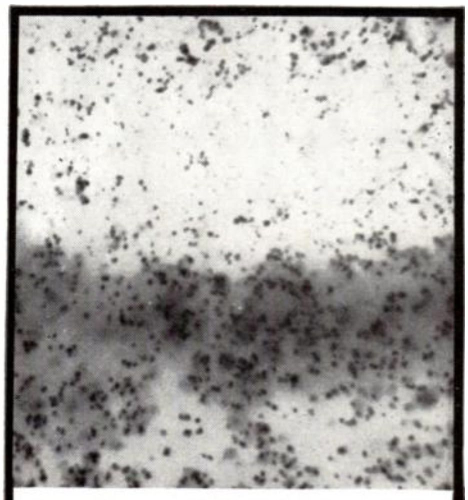
SURFACE STUDIES

Philco scientists have recently developed sensitive radiochemical methods which can detect about one-thousandth of a monolayer deposit on the hyperpure surface of a high frequency solid state device. Using radioactive tracers, it is possible to monitor the spatial distribution of materials deposited on the surface during synthesis of the device and thereby correlate measurements of device characteristics with observations of the physical properties of the surface under a variety of conditions. In this way, better insight is gained not only into the relation of surface phenomena to device performance but also into the mechanisms which influence the surface states.

Radiochemistry is one aspect of Philco's research program in solid state physics. Other projects include such studies as high vacuum mass spectrometry, structure of oxidic films, and nucleation of epitaxial films.

The Research Division offers attractive staff positions to scientists in a number of areas in addition to solid state research. Plasma physics, artificial intelligence, and communication theory are among the areas of interest. Letters of application or inquiry may be addressed to:

Mr. W. P. Diefenbach
Personnel Director
Philco Research Center
Blue Bell, Pennsylvania



Autoradiograph (1200 X) showing distribution of 0.06 monolayers of silver at edge of a metal-semiconductor contact.

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- Senior Design Specialists: Guidance & Control or Systems analysis & advanced design.

SOUTHEAST to $40,000

- Vice-Pres. for airborne instrumentation group. Must combine good technical background with ability to strengthen present contract acquisition capability.

SOUTHWEST to $17,000

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Personnel Search Specialists
225 E. Redwood St., Baltimore 2, Md.
LExington 9-6604

Positions Wanted

By Armed Forces Veterans

(Continued from page 98A)

LIBRARIAN

MS (Library) 9 years general, aeronautical, and electronic library experience. Familiar Government documents. Favor southern LI or midtown NY City. Box 3971 W.

ENGINEER

Military retired. Over 20 years training and experience in electronics maintenance and operations management for USAF. Some R&D associated with 8 years assignments on special projects. Interested in electronics or geophysical operations with possible assignments in Australia or New Zealand. Box 3978 W.

TEACHER

B.S. and M.S. in E.E.; present position Research and Development Engineer with six years experience. Desires teaching position in Fall of 1962. Box 3979 W.

SYSTEMS ENGINEERING

M.S.E.E., 5 years experience digital and analog systems. Interested in process analysis, simulation, feasibility study, and proposal preparation related to development and application of process control computers. Box 3980 W.

ELECTRONICS ENGINEER

Age 32, married. Desires challenging position offering security and advancement possibilities. Twelve years experience in ground radar and communications. Five years of experience has been in systems engineering, primarily microwave, training and supervisory experience; Commercial pilot. Box 3981 W.

Professional Group Meetings

(Continued from page 86A)

CIRCUIT THEORY
INFORMATION THEORY

Omaha-Lincoln—October 18

Symposium on Transmission Line Problems:

"600 Ohm Bridged H Pads for Telephone Line Attenuation," R. W. Hicks, AT&T, Omaha, Neb.

"300 Ohm H Pads for TV Broadcast Receivers," E. Olney, Columbia TV Service, Omaha, Neb.

"Theory of T & H Pad Attenuators," H. W. Becker, P. E., Omaha, Neb.

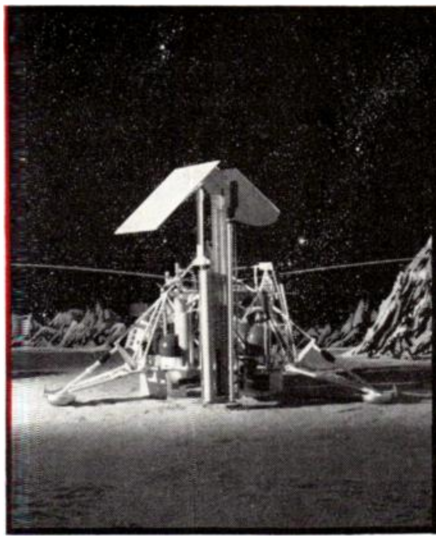
"Telephone Line & Repeater Problems," R. Barnett & O. Stauers, Western Electric, Omaha, Neb.

"Types of Faders in Broadcast Studios," S. Davidson, WOW-TV, Omaha, Neb.

(Continued on page 102A)

ELECTRONICS & SPACE
ENGINEERS

New long-range contracts at Ryan have created top-salaried opportunities in sunny San Diego



RYAN ELECTRONICS *was selected by Hughes Aircraft Company, Inc., to design, develop and fabricate the Radar Altimeter and Doppler Velocity Sensor equipments for NASA's Surveyor Lunar Soft Landing Spacecraft.*

SATURN LAUNCH VEHICLE *will employ radar altimeters now under development at Ryan Electronics. Engineers interested in a space age future are invited to get in touch with fast-moving Ryan Electronics right now.*

WORLD LEADER *in the design and production of Doppler Navigators for all types of aircraft now flying or projected, Ryan Electronics seeks engineers for challenging opportunities in sunny San Diego.*

Ryan is on the move! New contracts call for **years** of research, design and development work on such space age projects as radar altimeters for the Surveyor lunar soft landing vehicle and the Saturn launch vehicle. Ryan Electronics is also the world leader in Doppler navigation systems, as well as in the field of space guidance and control. If you are a career engineer interested in a top-salaried job with stable, diversified Ryan... if you would like to live in cool, sunny San Diego on the blue Pacific where the living is America's finest... **Send resume in complete confidence to: JAMES C. KERNS, Ryan Aeronautical Company, 5650 Kearny Mesa Rd., San Diego 12, California.**

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For project engineering assignments on automatic navigation, air data and flight control systems and devices. Also assignments in ground support equipment in the above areas, as well as in our new ordnance equipment group. Experience in electromechanical/optical systems desirable.

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Professional Group Meetings

(Continued from page 100A)

COMMUNICATIONS SYSTEMS

Rome-Utica—October 2-4

"More than 50 papers given at the Seventh National Communications Symposium at the Banquet," F. Virden, Rear Admiral US Navy.

Syracuse—October 24

"Survey of Satellite and Space Communications," R. P. Haviland, Gen'l Elec. Co., Philadelphia, Pa.

Toronto—October 23

"Engineering Considerations for the CNT Alaska Microwave System," C. J. Bridgland, Canadian National Telegraph, Toronto, Canada

Washington, D. C.—September 11

Panel Discussion on SPACE COMMUNICATIONS

Panel Moderator: Mr. E. W. Allen

Participants: Dr. H. Trotter, Jr., Pres., Gen'l Tel. & Elctronics Labs., Dr. E. A. Laport, RCA Labs.; Mr. R. P. Gifford, Gen'l Electric Co.; Dr. R. Kompfner, Bell Telephone Labs.

COMPONENT PARTS

Houston—October 10

"Characteristics and Trends of Film Capacitors," D. Harris, H & M Research Corp., Cleveland, Texas

A question and answer session was held at the conclusion of the paper.

ELECTRON DEVICES

Boston—October 5

"Microwave Power Varacters for Harmonic Generation," K. Gunn and R. Fekete, Sylvania Electric Products Inc., Woburn, Mass.

San Francisco—October 18

"Electron Beam—Plasma Interaction for Microwave Amplification," Dr. M. A. Allen, Stanford University, Stanford Calif.

Washington—September 25

"Thin Film Active Devices," J. P. Spratt, Philco Co., Blue Bell, Pa.

ELECTRONIC COMPUTERS

Baltimore—October 10

"Thin Films," Dr. M. Lauriente, Westinghouse Electric, P. O. Box 746, Baltimore 3, Md.

Boston—October 18

"The FX-1 Computer," J. Mitchell, K. Konkle and J. Raffel, Lincoln Lab., Lexington, Mass.

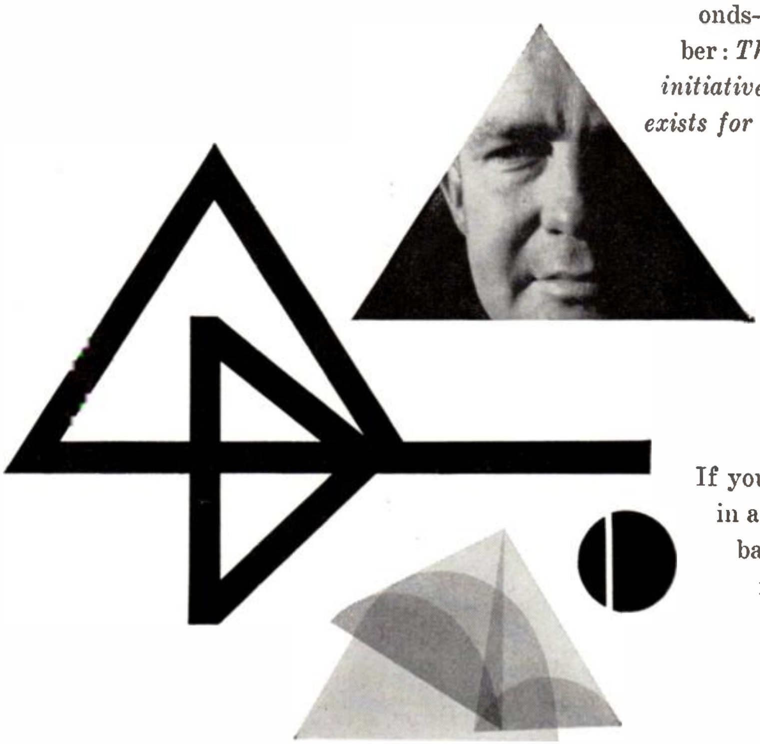
(Continued on page 104A)

ELECTRONICS

ACTIVITY REPORT

HOW TO MEASURE 2⅓ INCHES IN 186,000 MILES

▲ The task was a tough one for electronics engineer Larry Mills at Martin in Orlando, Florida: come up with a more accurate means of measuring time lapse between pulses in radar, missile, and lab devices. ▲ His solution: measure centers of energy from pulse to pulse rather than, conventionally, the leading edges. How? By specially processing the wave forms; then comparing against easily measurable calibrated standards. Result: accuracies to ± 0.2 millimicroseconds—using off-the-shelf components. ▲ Point to remember: *This outstanding development is a product of individual initiative at work in the creative technical environment that exists for scientists and engineers at Martin in Orlando, Florida.*



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If you have demonstrated similar initiative and ability in any of these fields,* write to us now. Outline your background and experience so that we can best match your interests to our own current or future work in advanced programs, in a wide variety of technical areas, or on one of our *six* major prime-contract missile or electronic systems. Tell us about yourself today. ▲ Send information to: C. H. Lang, Martin/Orlando, Orlando 63, Florida. ▼

**Immediate technical staff openings include:*

COMMUNICATIONS / INFORMATION THEORY / DIGITAL COMPUTER RESEARCH / OPERATIONS RESEARCH / WEAPONS SYSTEMS ANALYSIS / GUIDANCE & CONTROL SYSTEMS / HIGH RESOLUTION RADAR / FLUID DYNAMICS / AEROSPACE DYNAMICS / HUMAN FACTORS

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

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SYSTEMS
ENGINEERS

SALARY: TO $20,000

Several immediate openings exist at Hughes-Fullerton for Computer Systems Engineers qualified in the design and testing of large scale digital systems—and in the preparation of proposals for large scale digital computers.

These professional assignments involve such R & D areas as:

- **Solid state digital circuitry involving millimicrosecond logic**
- **Microwave carrier digital circuits**
- **Sub-microsecond core memory**
- **Thin film storage techniques**
- **Functional circuit concepts**
- **Micro-miniaturization concepts**
- **Tunnel diodes** ▪ **Microwave parametrons** ▪ **Circuit organization for maximal-speed computing.**

Located in Southern California's Orange County (the nation's fastest growing electronics center), Hughes-Fullerton offers you: a stimulating working environment; private or semi-private offices; long-term stability.

CALL COLLECT TODAY!
For complete information on these challenging assignments, call us collect **today!** Ask for:

Mr. B. P. RAMSTACK at:
TRojan 1-4080, ext. 3741.

Or, airmail resume to: HUGHES-FULLERTON R & D, P. O. Box 2097, Fullerton 1, California.

At Hughes, all qualified applicants will be considered for employment without regard to race, creed, color or national origin.

HUGHES

HUGHES AIRCRAFT COMPANY

Professional Group Meetings

(Continued from page 102A)

Long Island—October 24

"Delay Lines," J. Harwood, Ferranti Electric, Plainview, N. Y.

"Magnetic Drums," B. Slover, Bryant Computer Products, Walled Lake, Mich.

"Cores," G. Kruschke, Fabri-Tek, Amery, Wis.

Long Island—June 6

Hi Density Recording," A. Gabor, Potter Instruments.

Orange County—October 26

"High Density Magnetic Recording," Dr. A. S. Hoagland, IBM, San Jose.

Philadelphia—October 10

"A Hybrid Computer System," J. P. Heid, Mauchly Associates, Inc., Ambler, Pa.

Pittsburgh—September 27

"Basic Analog Computation (with film)," B. Johnson, Electronic Associates, Long Branch, N. J.

San Francisco—September 26

E. Koenigsberg, C-E-I-R., Inc., San Francisco, Calif.

Dr. Koenigsberg briefly reviewed the objectives of Operation Research and illustrated the utilization of Digital Computers by means of an example. Methods of machine solution utilizing linear programming techniques were briefly described.

Santa Ana—September 28

Panel Discussion: "Advanced High Speed Techniques," K. Uncapher, Rand Corp., Santa Monica, Calif.

Panel Members: M. Rosenberg, EMI; R. Brown, Burroughs; J. Rogers, STL; S. Nissim, TRW.

ENGINEERING MANAGEMENT

Rome-Utica—September 12

"Problems in System Engineering," R. Chipp, Rodd Chipp Associates, Bloomfield, N. J.

San Francisco—October 11

"Speculative Features of Electronic Stock," J. A. Chartz, Dalmo Victor Company, Belmont, Calif.

Syracuse—October 12

"Engineering: A Training Ground for Managers?" Dr. G. L. Haller, Gen'l Elec. Co., Syracuse, N. Y.

(Continued on page 106A)

PRATT & WHITNEY AIRCRAFT
DEPENDABLE ENGINES

instrumentation
and
controls
engineers
for
FLORIDA

The growing role of Pratt & Whitney Aircraft in the national aerospace program is apparent in the research and development projects and long-range product planning being conducted at their Florida Research and Development Center.

At present, openings exist at various experience levels in the following fields:

●

DATA PROCESSING ENGINEERS (BSEE, Engineering Physics):

Three to five years experience in set up, calibration, operation and data reduction techniques used in instrumentation operations with oscillograph, strip chart, and digital data systems.

●

CONTROL ENGINEERS (BSEE, BSME):

Three to five years experience in system design, trouble shooting, and calibration of electro-mechanical, electro-hydraulic, and electro-pneumatic servo control systems. Good mechanical background is necessary.

●

VIBRATION ENGINEERS (BSEE, BSME):

Three to five years experience in 1) operation of random control vibration equipment, 2) design of mounting fixtures for random vibration testing, and 3) working with accelerometer systems for measuring vibrations on rocket and jet engines.

Please submit resume, including minimum salary requirements, to:

MR. J. W. MORTON
OFFICE 57

PRATT & WHITNEY AIRCRAFT
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All replies will be handled promptly and in complete confidence.
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(An unusual bit of evidence to support our claim of "outstanding engineering opportunities")

As MANAGER OF ENGINEERING, and a key contributor to successful and far-reaching expansion plans, Dick Gifford feels the loss of one set of windows is a small personal price to pay for more than 22,000 square feet of new space for expanding Engineering operations.

But this is only about 20% of all the plant expansion now underway. A second wing will further increase total plant floorspace owned and leased from 400,000 to over 500,000 square feet by March, 1962.

In recent years, the Communication Products Department has built a healthy mix of industrial, commercial and military programs. Products represent a spectrum of communication systems—from tiny, personal two-way radios, to continent-spanning microwave radio relay networks. Now, by continuing to grow along virtually this entire front, the professional opportunities newly created carry a hard-to-find combination of two key career factors: rapid growth potential, *and near-optimum stability.*

High-Priority Programs for Defense; Highly Competitive Products for Industry—

In the military sphere, we recently added a new program of great importance to the nation's security—developing the intersite communications link for Titan II operational bases . . . a system which must be *survivable and reliable.*

In our commercial areas, a number of new opportunities have been created by expansion of such programs as: a 600 channel, fully transistorized and miniaturized multiplex telephone carrier system; a 600 channel microwave radio relay system (also fully transistorized) embodying a heterodyne repeater, parametric and traveling wave tube amplifiers; compact, transistorized two-way FM mobile radio; highly miniaturized personal communication equipment; tropospheric scatter equipment with modulation techniques, designed specifically for thin route applications.

Immediate Openings In These Areas:

Military Communication Systems Engineering
Military Communication Equipment Design
Telephone Carrier Circuit Design
RF and Video Circuit Design
Communication Equipment Advance Development
Solid State Receiver Design
2-Way Mobile Transmitter Design
2-Way Mobile Systems & Reliability Engineering
2-Way Radio Mechanical Design

■ Your letter or resume will receive careful attention and a prompt, personal reply. Write in confidence to Mr. W. J. Kelly, Div. 53—MA, Communication Products Department, General Electric, Mountain View Road, Lynchburg, Virginia.

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Advanced development of digital computer techniques and solid state devices on navigational, tactical and training projects for aircraft, missile and aerospace simulator systems.

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Installation, maintenance, operation and instruction on company and competitor electronic equipments, including communications, computers, radar and simulator systems. Assignments in this country and abroad.

Our modern facilities are located in residential suburbs of Washington, D. C. and New York City, affording you a choice of city, suburban or country living. Advanced study under tuition refund may be conducted at several nearby universities.

For details, contact: Mr. A. C. Sugalski, Manager, Professional Staffing, Dept. A-16 at our Riverdale facility.

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Professional Group Meetings

(*Continued from page 104A*)

ENGINEERING WRITING AND SPEECH

Philadelphia—October 11

"Learning Machines," J. S. Bryan, Philco Research Div., Blue Bell, Pa.

San Francisco—October 24

"Description of the Language-Translating Machine AN/GSQ-16," Dr. H. G. Kolsky, IBM Research Laboratory, San Jose, Calif.

HUMAN FACTORS IN ELECTRONICS

Los Angeles—October 10

"Logic and Language: Human Factors at Inception," T. Steel, Systems Development Corp., Santa Monica

"Circuits and Maintenance: Design for Future Trouble," W. King, Ramo-Wooldridge Corp.

"The Computer Center: Human Factors in User Organization," R. W. Rector, Space Technology Labs.

INDUSTRIAL ELECTRONICS

Omaha-Lincoln—October 25

"Standardization and Programming of Automatic Test Sets," P. P. Povilus, Western Elec. Co., Millard Plant, Neb.

Tour of Western Electric Test Set Room and Assembly Floor.

Schenectady—October 10

"Digital Applications of Thermoplastic Recording," R. G. Reeves, General Electric Co., Schenectady, N. Y.

INFORMATION THEORY

Boston—October 23

"Is an Analytical Auditory Theory Possible," Prof. W. Siebert, M.I.T.

San Francisco—October 18

"Networks of Noisy Channels," Prof. P. Elias, MIT, Cambridge, Mass.

MICROWAVE THEORY AND TECHNIQUES

Long Island—September 19

"Information Theory in Antenna Design," W. D. White, Airborne Instruments Lab., Deer Park, N. Y.

San Francisco—November 8

"Optical Masers," Dr. A. L. Schawlow, Stanford University, Stanford, Calif.

(*Continued on page 108A*)

ELECTRONIC ENGINEERS
Research and Development

Armour Research Foundation, one of the largest independent research organizations, offers exceptional opportunities in Chicago for qualified Electronics research engineers to work on a variety of industrial and military research projects in the fields of

RADAR
ANTENNAS
CIRCUIT DEVELOPMENT
MICROWAVES
GUIDANCE & CONTROL
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COMMUNICATION
SIGNAL PROCESSING
SEMICONDUCTOR DEVICES
MEASUREMENT TECHNIQUES
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RELIABILITY ANALYSIS
COMPONENT DEVELOPMENT

Minimum requirements include a B.S. degree in Electrical Engineering with above average academic grades. Research experience in one of the above fields is highly desirable, though not absolutely necessary.

Our environment combines the better features of academic and industrial research and offers you the opportunity to work on advanced research programs.

Staff members receive attractive salaries, up to 4 weeks vacation, generous insurance and retirement benefits, and tuition paid graduate study. Our Chicago location offers excellent cultural and recreational advantages. All qualified applicants will receive consideration for employment without regard to race, creed, color, or national origin. Please reply in confidence to Mr. Ron C. Seipp.

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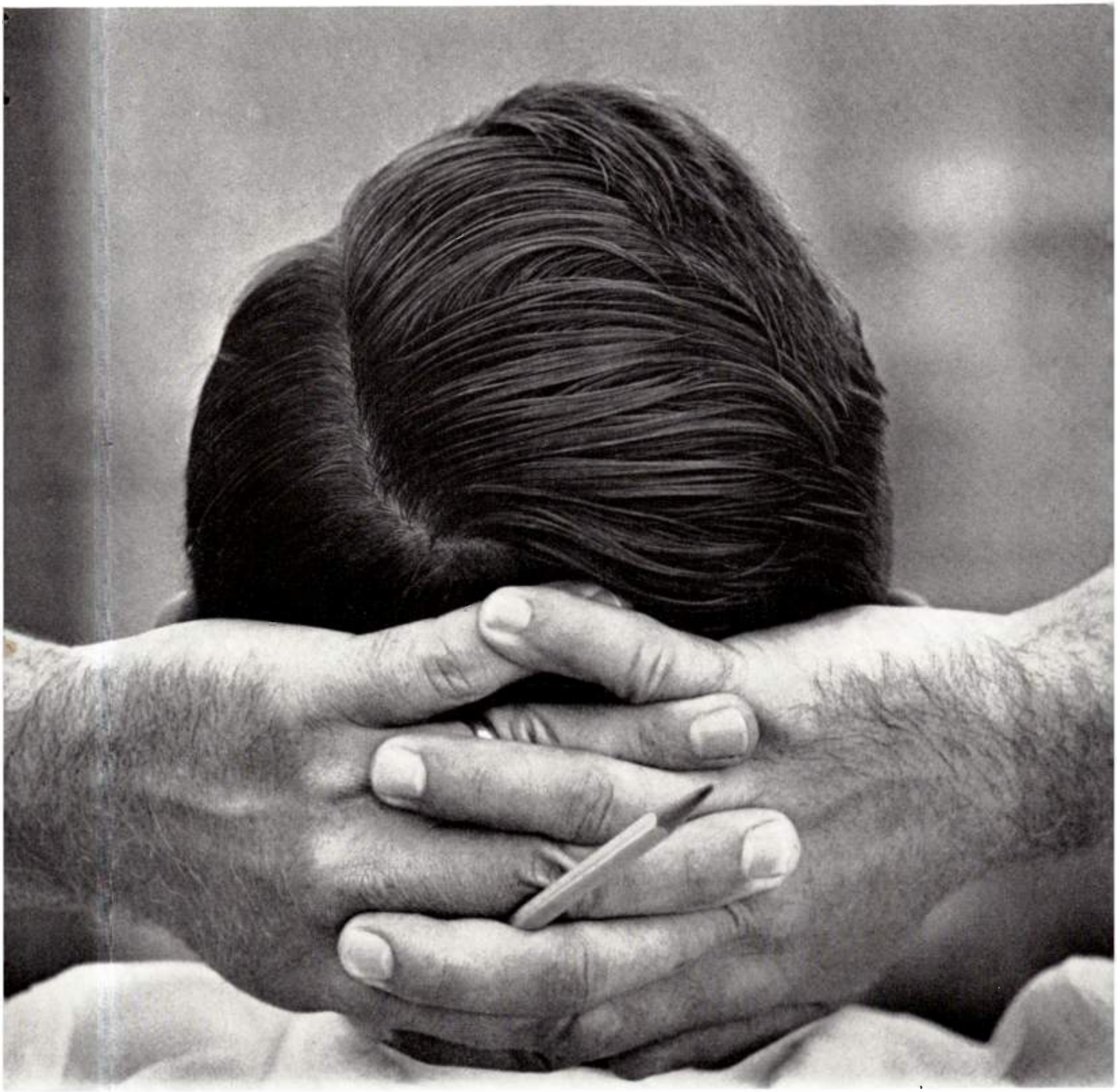
How many of his dreams do you share?



This man gets paid for dreaming. He seeks out new questions to ask, new goals to aim at. His insights shape the course of tomorrow's technology.

Do you, too, look beyond tomorrow? Then come to Northrop. Work in such uncluttered areas as space guidance and astro-inertial navigation systems, aerospace deceleration and landing systems, man-machine and life-support systems for space, laminar flow control techniques, automatic test equipment or world-wide communications systems. Or come and set your own goals where the challenge is greatest. You'll find more than 70 advanced projects already on the boards, and all the creative freedom you could ask for.

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- **MICROWAVE TUBES & DEVICES**
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- **CATHODE RAY TUBES**
  Design or manufacture of black and white PICTURE TUBES, and INDUSTRIAL and MILITARY TUBES.

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Westinghouse
ELECTRONIC TUBE DIVISION ELMIRA NEW YORK

Professional Group Meetings

(Continued from page 106A)

Washington—October 10

"The Relative Microwave Absorption Cross-Section of Mankind," O. M. Salati, Moore School of E.E.

Panel to stimulate discussion and bring out points of interest: Dr. D. E. Goldman, R. A. Sparka, W. Carley, C. Arsem.

MICROWAVE THEORY AND TECHNIQUES
ANTENNAS AND PROPAGATION

Baltimore—October 11

"Criteria for Conical Scan Tracking Antennas," T. Cheston, Applied Physics Lab., Johns Hopkins Univ., Silver Spring, Md.

MILITARY ELECTRONICS

Boston—October 11

"NIKE-AJAX and Missilemaster in Air Defense," Lt. M. L. Brady, NIKE-AJAX Site, Reading, Mass.

"Missilemaster in Air Defense," C. L. Weichert, Martin Co., Ft. Heath, Mass.

Fort Huachuca—October 12

"Engineering Aspects of the Electromagnetic Environmental Test Facility (EMETF) Interference Prediction Model" Dr. A. J. Hoehn, Bell Aerosystems Co., Tucson, Ariz.

Long Island—October 17

"Atlantic Missile Range Instrumentation," J. Halks, Pan American WA, Cape Canaveral, Fla.

Los Angeles—October 20

"DOD Plans to MEET the Scarcity of Future Engineering Talent," Honorable J. M. Bridges, DODRE Office of Secretary of Defense, Wash., D. C.

San Diego—October 11

"Atmospheric Probe for Planet Venus," R. W. Clapp, Hughes Aircraft Company, Los Angeles, Calif.

NUCLEAR SCIENCE

Chicago—October 13

"Neutron Activation Analysis," Dr. J. Haffner, Armour Res. Foundation, Chicago, Ill.

Los Angeles—October 17

"Radiological Defense Programs in California," J. W. Davis, State of Calif. Disaster Office, Pasadena

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

(Continued on page 110A)

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MICROWAVE COMMUNICATIONS ENGINEERS

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Work involves engineering, supervision of installation, operation and maintenance of microwave communications systems.

College Degree Mandatory

- - - - - - - -

MICROWAVE COMMUNICATIONS TECHNICIANS

Must be qualified by education and experience to inspect installation, operate and maintain microwave communications systems to insure quality and performance standards.

Technical or trade school training mandatory.

- - - - - - - -

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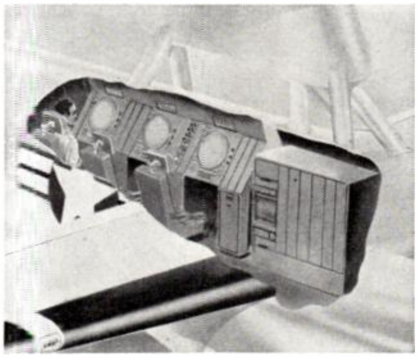
ADVANCED AIR DEFENSE SYSTEMS WITH FIRST-DAY CAPABILITY

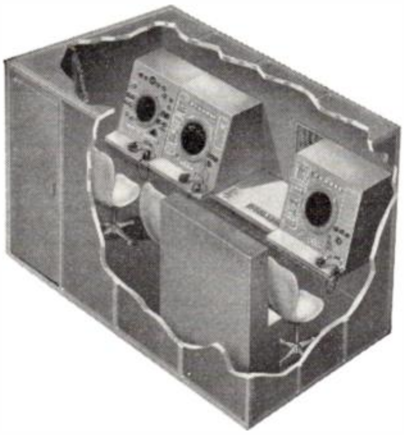
The needs of today's air defense systems pose a problem that would have seemed insoluble ten short years ago. The problem of furnishing mixed-weapons command and control, with first-day capability, in a system that is portable to any place in the world.

Here is how that problem has been solved through creative engineering utilizing a decade of industry progress in tactical data systems.

Systems already delivered by Litton to the military, or in the advanced state of development and production, include: Airborne Tactical Data Systems (AN/ASQ-54, AN/ASA-27) for the U.S. Navy, the Marine Corps Tactical Data System (AN/TYQ-1, AN/TYQ-2) for the U.S. Marine Corps, and the AN/FSG-1 Retrofit Improvement System (OA-3063/FSG-1 (V)) for the U.S. Army.



The first of these, the Airborne Tactical Data Systems, provides a capability for the mission of Airborne Early Warning and Control (AEW & C) in defense of large land masses, attack carrier task groups and other naval units. Both the AN/ASQ-54, installed in a land-based AEW & C aircraft, and the AN/ASA-27, installed in a carrier-based AEW & C aircraft, furnish early warning data on enemy raids to surface elements of an air defense network and provide airborne control of interceptors.



The second of these systems, the Marine Corps Tactical Data System (MTDS), features capabilities for continuous and effective control of Combat Air Operations during an amphibious assault. Facilities are available for control of aircraft on missions such as close air support, reconnaissance, and interdiction and for air defense with mixed weapons, both ship-based and shore-based surface-to-air missiles and interceptors. An integral air traffic control system assists in initial and continuous identification of friendly aircraft.

The third, the AN/FSG-1 Retrofit Improvement System, significantly increases the counter-countermeasures capability of the AN/FSG-1 Missile Master System deployed within the Continental United States to furnish surface-to-air missile battery coordination in the defense of large cities and industrial areas.

Through the successful design, development and manufacture of systems for air defense missions, Litton has demonstrated its capability to proceed with even further advanced data systems. Such systems are now under conception and development at Litton.

Air defense systems that not only fulfill today's defense requirements but also defy obsolescence for years to come require engineering that is versatile, inventive, aggressive, and adaptable. This is the kind of engineering Litton expects from its people. If you are qualified to perform engineering at this level, you are invited to write to H. Laur, Litton Systems, Inc., Data Systems Division, 6700 Eton Avenue, Canoga Park, California; or telephone DIamond 6-4040.

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ELECTRONICS ENGINEERS

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PHYSICISTS

Theoretical/ Experimental Research on new

GYROSCOPE PROGRAM

An opportunity to join the Scientific Research Staff of Republic's new Paul Moore R & D Center, performing basic and applied research in the sciences related to aerospace flight. Initial assignment will be theoretical and experimental research on new gyroscope program employing principles of paramagnetic resonance.

Applicants should have, or be working towards, PhD degree in Physics or Electrical Engineering; 7-14 years research/development experience (or equivalent graduate study); and broad circuit analysis and synthesis capability. Experience in magnetics, gyroscopics, or magnetic resonance techniques desirable.

Write in confidence to
Mr. George R. Hickman
Technical Employment
Manager
Dept. 14A

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Professional Group Meetings

(Continued from page 108A)

PRODUCT ENGINEERING AND PRODUCTION

Philadelphia—October 11

"Development, Design and Application of Structural Welded Wire in Product Engineering," T. W. Whipple, E. H. Titchener & Co., Binghamton, N. Y.

RADIO FREQUENCY INTERFERENCE

Philadelphia—October 17

"The International Special Committee for Radio Interference—Its Organization and Program," R. M. Showers, University of Pennsylvania, Phila., Pa.

San Francisco—September 19

The Chairman reported on the National IRE Professional Group meeting held August 23, 1961.

Members discussed the forthcoming National PGRFI Symposium (1962). There was no speaker.

RELIABILITY AND QUALITY CONTROL

Columbus—October 12

"Modern Quality Control Techniques," R. Corwin, Kimball Glass Co., Columbus, Ohio

Los Angeles—October 16

"Technical Justification," J. Orme, Aerojet-General, Azusa.

"Company Standards Lab Requirements," E. A. Anderson, Navy-BuWeps, Pomona.

"Standards Problems of Customers," W. Hogan, Leeds & Northrop, Monterey Park.

Philadelphia—October 17

"User Approaches to Higher Reliability Specifications for Component Parts," Panel Discussion.

Panel Moderator: W. T. Sumerlin, Philco Corp., Phila., Pa.

Panel Speakers: M. Chamow, G. E., Phila., Pa.; W. Garner, Westinghouse, Baltimore, Md.; A. Schwartz, RCA, Camden, N. J.; P. Cook, Corning Glass, Bradford, Pa.; L. Jacobson, IRC, Phila., Pa.; Dr. D. Van Winkle, Continental Devices Corp., Hawthorne, Calif.

SPACE ELECTRONICS AND TELEMETRY

Los Angeles—September 19

"Physiological Telemeter Systems," A. D. Bredon, Spacelabs, Inc., Van Nuys, Calif.

RELIABILITY ENGINEERS and QUALITY ASSURANCE ENGINEERS

Symposium Interviews
Washington, D.C.
January 9-11, 1962

STL's expanding Reliability and Quality Assurance Program has created positions with responsibility for the developing, implementing and auditing of these programs for advanced weapons systems and space programs. Responsibility extends through all phases, including:

- Proposal Preparation
- Program Planning
- Hardware Development
- Procurement
- Manufacturing & Test
- Field Support

Requirements:

Reliability Engineers—BS in Engineering or Physical Sciences with 5 or more years experience in aerospace or related industries in reliability programs, preferably including hardware design experience.

Quality Assurance Engineers—Positions range from BS in Engineering or Physical Sciences with 2 or more years experience in aerospace or related industries in quality assurance programs—to BS in Engineering with 10 or more years experience, including 5 years in quality assurance planning, and 3 years design engineering experience.

Symposium interviews arranged at headquarters hotel by calling Mr. H. R. Stevens, EXecutive 3-5030, Tues., Jan. 9 thru Thurs. Jan. 11, or send resumes to Mr. H. R. Stevens at STL, an equal opportunity employer.

SPACE TECHNOLOGY LABORATORIES, INC.
P.O. Box 95005I
Los Angeles 45, California
a subsidiary of Thompson Ramo Wooldridge Inc.

"Signal Attenuation During Missile Staging," E. L. Tarca, Seigler Corporation, Inglewood, Calif.

San Francisco—October 17

"Recent Developments in Optical Pumping," Dr. A. Bloom, Physics Corp., Palo Alto, Calif.

Washington, D. C.—October 24

"Techniques of Bandwidth Compression for Telemetry," Dr. H. Blasbalg, Signal Analysis Group, IBM, Rockville, Md.

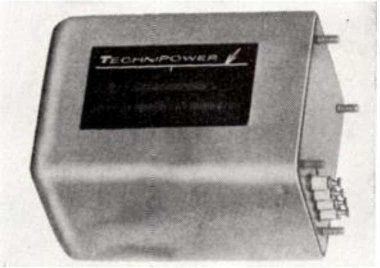
NEWS New Products

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

(Continued from page 45A)

AC-DC Power Supplies

Based on the acceptance of their existing line of DC Power Modules, **Technipower, Inc.**, 18 Marshall St., So. Norwalk, Conn., has now introduced a wide line of high temperature dc power modules.



Using all silicon semiconductors and tantalum capacitors, these modules have a temperature rating of −40 to +100°C. Over 90 models are available ranging from 2.8 to 52 volts at powers of 1.0 to 20 watts. Other specifications are as follows: input voltage range −105 to 125 volts, 50–400 CPS.

Regulation, both 0.05% and 0.5% models available.

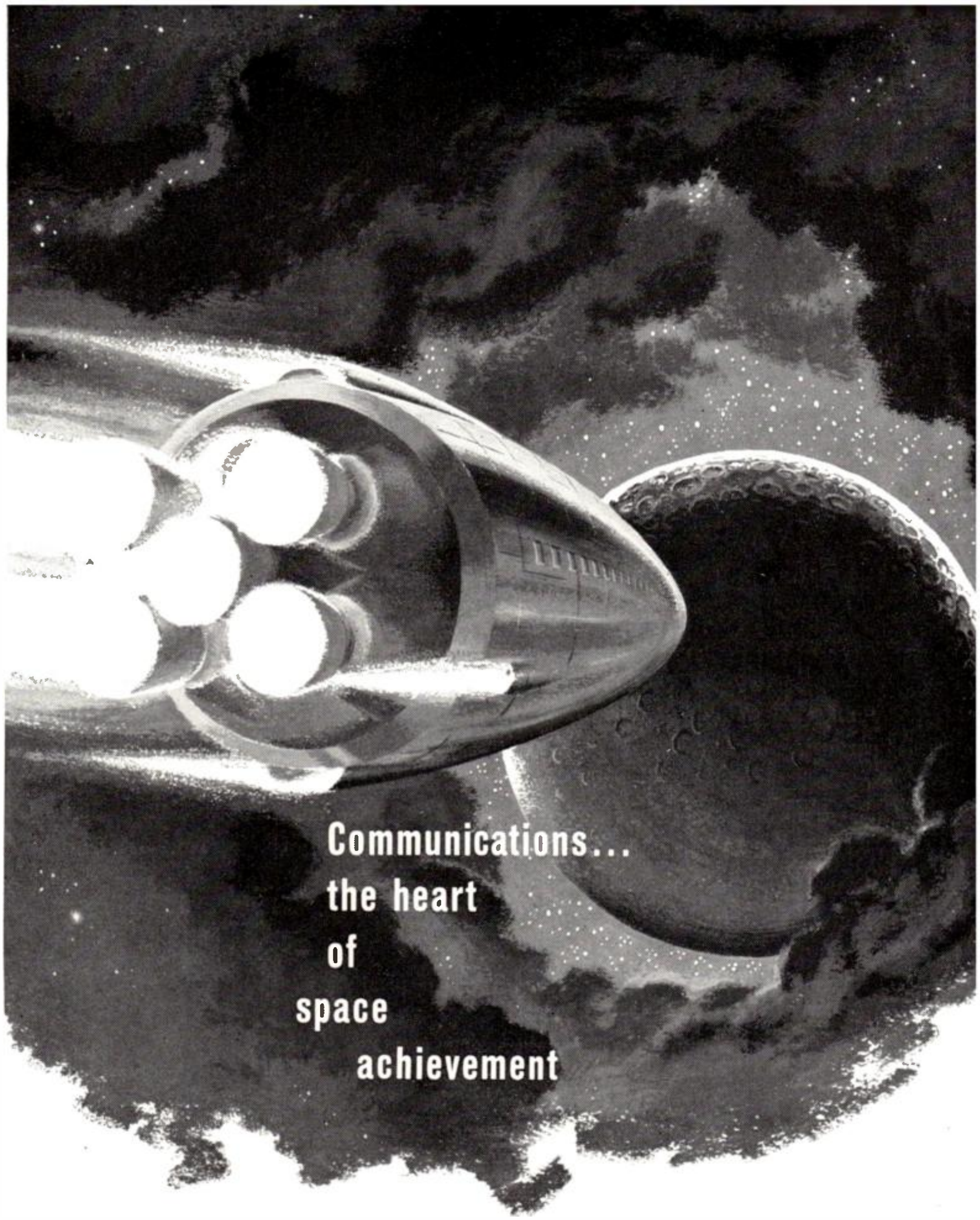
Ripple is 1mv RMS for 0.05% regulation and 5mv RMS for 0.5% regulation.

An 8-page brochure is available with complete data.

Metal Film Resistors

The Vamistor® line of precision metal film resistors by **Weston Instruments Div., Daystrom, Inc.**, 614 Frelinghuysen Ave., Newark 14, N. J., has been expanded to include new commercial Weston Class C units with a temperature coefficient of 0±25 ppm/°C over a range of +165°C to +25°C and 0±35 ppm/°C over a range of +25°C to −55°C. This class of unit is suited to precision test equipment applications at room temperature and above, but where low temperatures are seldom encountered.

(Continued on page 112A)



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

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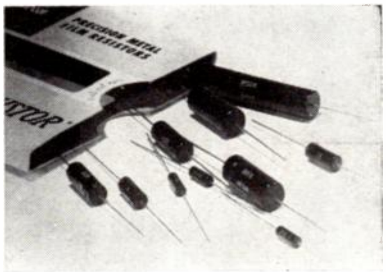


The range presently available includes a military ⅛ watt model and both military and missile (gas-filled) models in 125°C ratings of 0, ½, 1, and 2 watts. The missile ½ watt type is the size of a ¼ watt unit and the 1 watt unit the same size as a ½ watt unit. Both these units are available in the "F" designation under MIL-R-10509D.

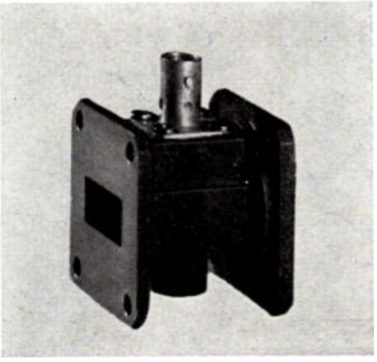
The component is virtually impervious to weather, humidity, and salt spray conditions and combines good high frequency performance, long range stability, good shelf life, and freedom from noise.

Construction features include use of a low temperature coefficient alloy internally deposited on a steatite tube and spiralled to resistance value. Conducting bonds are fire-bonded silver, and encapsulation is of epoxy resin.

For more information on the line and a complete list of new prices, write to the firm.

SPST Diode Switch

Consolidated Microwave Corp., 850 Shepherd Ave., Brooklyn 8, N. Y., announces the addition of CMC Model X110 to its new line of solid state microwave switching devices. The unit contains a semiconductor diode which allows high speed switching with low control power requirements. Compact, lightweight, and ruggedized construction lends to its reliability for military applications.



CMC Model X110 operates in the frequency range of 8.2 to 12.4 Gc with typical isolation of 20 db at 9 Gc and insertion loss of 1.3 db. Its switching speed is less than 10 nanoseconds. RF power handling capacity is 4 watts CW and 150 watts

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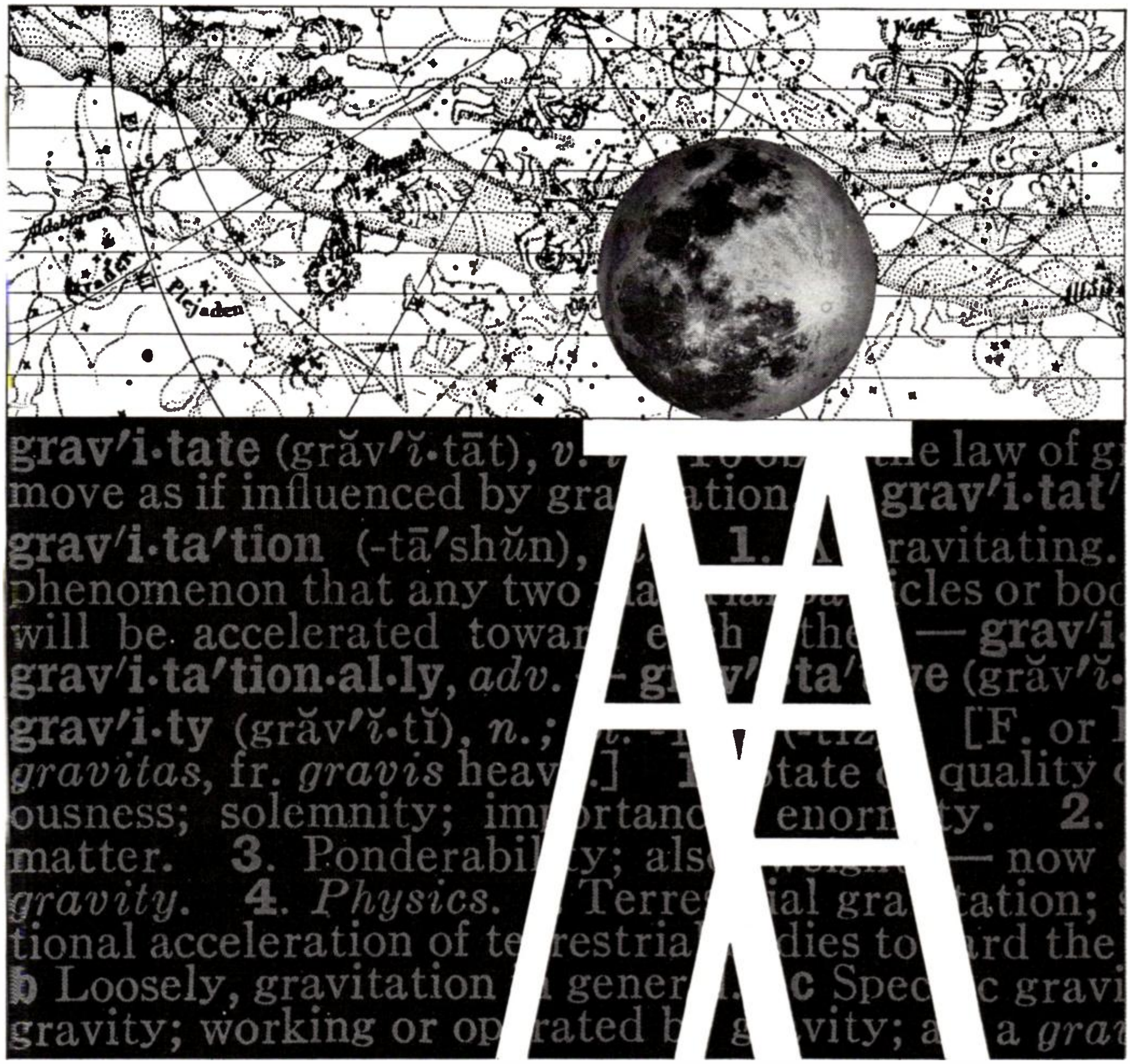


purposeful imagination....in space

The men of Aerospace probe all disciplines of the space sciences. Chartered exclusively to serve the United States Government, Aerospace Corporation applies the full resources of modern science and technology to advanced space and ballistic missile systems. ☐ As part of the Air Force-science-industry team, the men of Aerospace evaluate and stimulate the free flow of information that results in the imaginative concepts required for national leadership in space. From concept to countdown and beyond, Aerospace provides advanced systems analysis and planning; theoretical and experimental research; general systems engineering and corresponding technical direction of programs. ☐ Aerospace Corporation, an equal opportunity employer, now needs more men to meet these responsibilities. Highly skilled engineers and scientists with advanced degrees, knowledgeable in interdisciplinary problem solving, are urged to contact Mr. George Herndon, Room 109, Aerospace Corporation, P. O. Box 95081, Los Angeles 45, California. ☐ Organized in the public interest and dedicated to providing objective leadership in the advancement and application of science and technology for the United States Government.

AEROSPACE
CORPORATION
2400 EAST EL SEGUNDO BLVD., EL SEGUNDO, CALIFORNIA

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

(Continued from page 112A)

peak. The temperature range is −50 to +80°C. An insertion length of only 1½ inches makes this unit extremely compact. Among many other applications, this device can also be used as an RF modulator and a voltage controlled attenuator.

Flutter Meter

This sensitive measuring instrument was engineered by **Amplifier Corp. of America,** 398 Broadway, New York 19, N. Y., to comply with the standards set by the IRE for flutter and wow. It is designed to give a visual indication of wow and flutter content of all types of tape recorders and playback equipment including 33⅓, 45, and 78 rpm discs, and 16 and 35 mm sound film mechanisms. A built-in preamplifier and high impedance (1 megohm) input attenuator will accept voltages ranging from 1 millivolt to 300 volts. Connection may be made directly across magnetic tape playback heads, or across high level circuits delivering up to 300 volts.

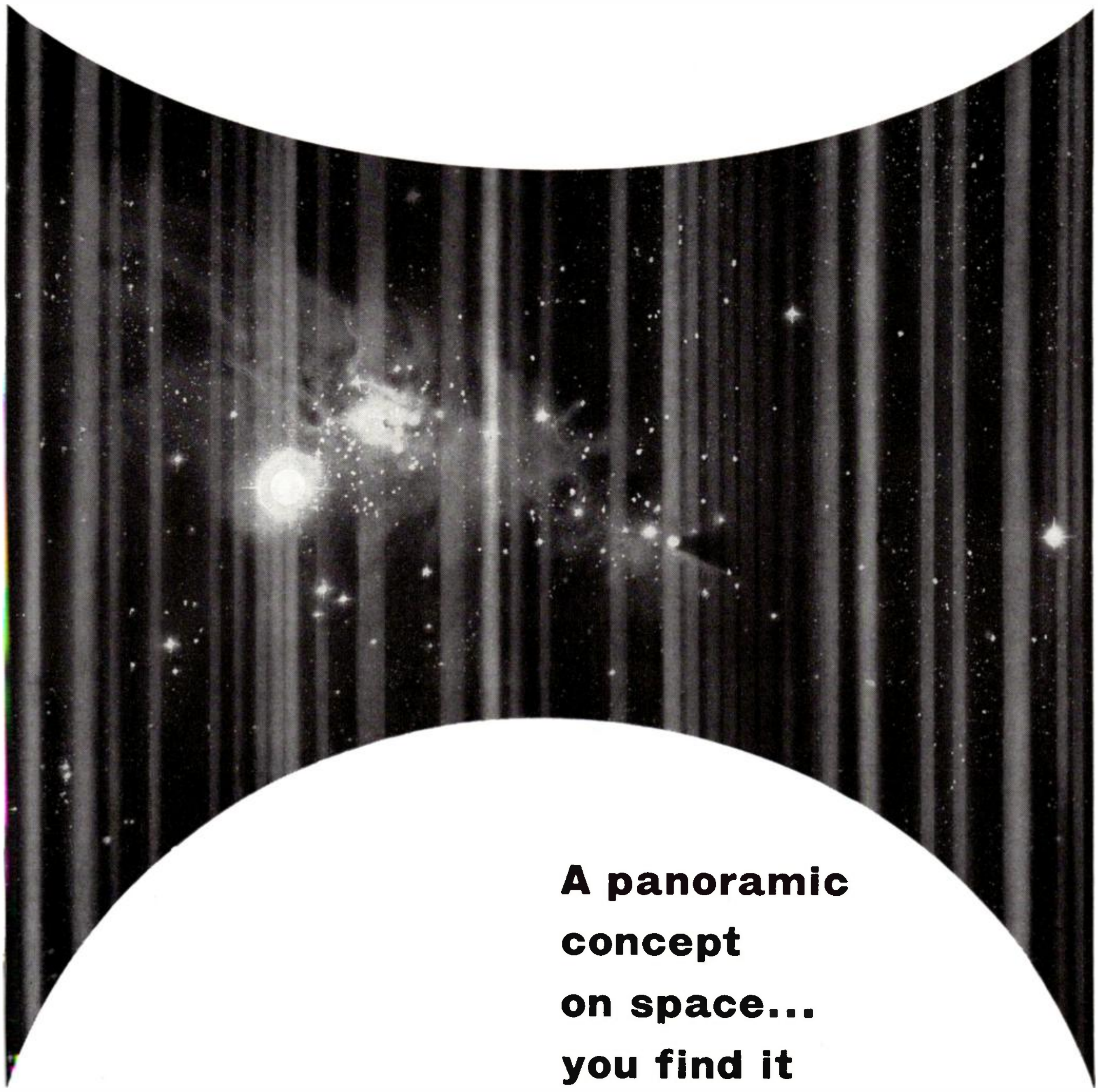
Flutter (the high frequency components) is measured through a high-pass filter while wow (or low frequency components) is measured with a low-pass filter in the circuit.

A built-in 3,000 cps oscillator is incorporated for recording purposes. This permits the Flutter Meter to be used as a complete instrument in itself and eliminates the necessity of calibrating and using external oscillators. A three-range filter is included to study and isolate flutter and wow components. These ranges are 0.5 to 6 cps for wow, 5 to 250 cps for flutter, and the combined flutter and wow of 0.5 to 250 cps.

A frequency discriminator is used to demodulate the flutter signals which are then read on a calibrated meter as RMS value of sine-wave flutter components. The built-in limiting preamplifier prevents erroneous readings from amplitude modulation components.

Hum, noise, switching surges, and other extraneous transients have no effect upon the reading or stability of the instrument. Three scales: 0.3%; 1.0%; and 3.0% are calibrated for flutter and wow. Significant readings of flutter or wow can be made down to 0.01% with reliability. Employs a 7-inch meter with two 5½-inch scales for three ranges.

(Continued on page 116A)



A panoramic concept on space... you find it at Lockheed

Long experience and proven capability give Lockheed a great advantage: perspective on space. This results in programs that cover the broad spectrum of aerospace dynamics; that present the most far-reaching technological problems; that offer endless creative challenge to Scientists and Engineers; that give them new concepts to explore—new goals to reach.

This far-ahead program in the concepts and vehicles of space creates a climate most favorable to advancement in the state of the art as well as in professional status.

Scientists and Engineers with an eye to a secure and rewarding future will find these openings at Lockheed worthy of serious consideration: Aerodynamics; thermodynamics; dynamics; electronic research; servosystems; electronic systems; physics (theoretical, infrared, plasma, high energy, solid state, optics); electrical and electronic design; structural design (wing, empennage, fuselage); human engineering; reliability; wind tunnel design. Write today to Mr. E. W. Des Lauriers, Manager Professional Placement Staff, Dept. 1801, 2402 N. Hollywood Way, Burbank, California. An equal opportunity employer.

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A DIVISION OF LOCKHEED CORPORATION

Maintaining the Lead in Infrared Reconnaissance

Leadership in the field of research requires a constant supply of new talent and fresh ideas. Engineers and physicists have maintained HRB's lead in airborne infrared reconnaissance by developing outstanding new IR systems and making improvements and modifications in current systems.

Evidence of HRB's superiority in the IR field is the famous Manhattan strip photo. Taken with HRB's RECONOFAX®* camera under conditions of complete darkness, the photo was accurate enough to reveal undeveloped areas in Central Park and residential, commercial and industrial areas throughout New York City.

Extensive facilities provide top scientific talent with the best conditions possible for advanced research and development work in infrared, but new talent is constantly required to maintain and expand this leadership role.

In addition to the many challenging technical programs, HRB offers unusual employee benefits. These include company-paid graduate study at The Pennsylvania State University, generous hospitalization, life insurance and vacation programs, as well as a company-paid retirement program.

If you are interested in learning more about career opportunities at HRB-Singer, write George H. Rimbach, Supervisor of Personnel, Dept. R-1.

**for technical data write Dept. I.*

HRB

HRB-SINGER, INC.

A SUBSIDIARY OF THE SINGER MANUFACTURING COMPANY
P.O. Box 60, Science Park, State College, Pa.

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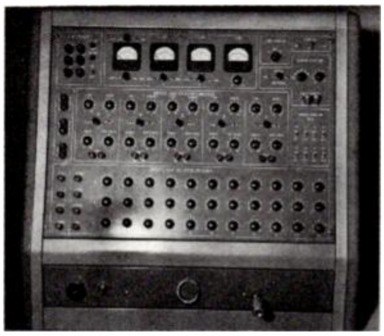
(Continued from page 114A)

In the "calibrate" position, the meter will also measure variations in amplitude of any rate from 0 to 40 cps. Controls on the panel consist of an input voltage selector, a vernier calibration adjustment, a capacity balancing control to compensate for differences in linear speeds of tape mechanisms, turntables and film mechanisms, a filter selector and a scale selector switch.

Assembled on a standard rack panel, dimensions are 8¾ inches high, 19 inches wide, 8 inches deep. Net weight is 18 lbs. Complete technical specifications and prices may be obtained by writing to the firm.

Printed Circuit Card Tester

This Printed Circuit Card Tester has been developed by **Ess Gee, Inc.**, 15 Havens St., Elmsford, N. Y., to simulate environments found under normal or extremely adverse operating conditions while testing or trouble shooting printed circuit cards.



The electrical outputs of any card may be analyzed using simulated inputs while the card is subjected to temperatures ranging from −50 to +185°F, and acceleration loads of 2g and 5g. Minor modification will allow this unit to simulate the pressure and humidity environments as required by card application. Input and output voltages may be monitored and variable input-output loading conditions may be accomplished at the control panel.

The Printed Circuit Card Tester is designed for desk, or bench-mounted operations. It is small enough and light enough to be readily movable.

Temperature control is accomplished through the means of a vernier thermostat which is continuously variable. A vibration switch selects the acceleration modes. Marginal voltage test switch is incorporated to vary the operating voltages ±10%. Input and output loading is accomplished through switch banks ranging from 0–100,000 ohms in ten-ohm steps. The environmental chamber has a re-

(Continued on page 118A)

The Calibre of An Aerospace Organization is Evidenced by the Calibre of the Problems Put To It by its Customers — and by the Company to its Engineers

Engineers, Scientists

what's your opinion:

1

ON HOW TO CREATE A REALISTIC SPACE SIMULATOR?

This is a question put to Bell Aerosystems last year by the Air Force. If you had been a member of Bell's technical staff then, you might have been one of those who developed the space environment simulation concept now being engineered into actuality here. It is based upon a high-resolution, closed-loop television system superimposing images generated by 3 studios. Many stimulating problems, however, remain to be solved in: development of an electronic planetarium, real time digital flight dynamics, television image insetting techniques, TV noise alleviation techniques.

2

ON HOW TO PROVIDE OPTIMUM VEHICLE TOUCHDOWN DISPERSIONS WITH A FULLY AUTOMATIC, ALL-WEATHER LANDING SYSTEM?

Your informed opinions on any aspect regarding landing of boosters or space vehicles will receive thorough consideration. Bell engineers have already created a system in landing conventional aircraft providing a 30% improvement in this area, as demonstrated by 4000 landing tests made by U.S. Navy and Air Force planes. Vehicle landing systems and specialized equipments are advanced concepts now under study.

3

ON HOW TO PROVIDE AN ASTRONAUT WITH POSITIVE, FLEXIBLE STEERING CONTROL IN SPACE?

Your answer will certainly contain the statement that much depends upon the particular space mission but that reaction controls will be the foundation of the system.

This revolutionary concept was originated by Bell engineers for Bell's own X-1B high altitude research plane. More refined Bell Systems were developed for the X-15, Mercury capsule, Agena booster/satellite and Centaur program. Bell's Reaction Controls are applicable to missiles, satellites, re-entry vehicles — indeed, any object requiring a continuous/intermittent or variable thrust vector. One unique recent application is the Bell rocket belt for the infantry man.

The three problems high-lighted above are representative of the wide range of unusual challenges offered engineers at Bell. Others concern such diverse projects as advanced Agena* rocket engines for NASA's Ranger and Marine vehicles; Hydroskimmer (air cushion) amphibians for the U.S. Navy; battlefield surveillance systems; automatic check-out equipment; missile & drone recovery systems. Another important field at Bell is Inertial Guidance and Navigation Systems, many of which employ an appropriate member of Bell's family of Hipernas** inertial platforms. *If you would like to pit your talents against tough-calibre problems, look into current openings with Bell.*

*Bell's Agena — most reliable rocket engine yet designed. Already used in Discoverer program, Midas, other satellites.

**Hipernas (HIgh PERformance NAvigational System) — the most accurate, pure-inertial, self-compensating navigation and guidance system known.

INERTIAL GUIDANCE

Various degree levels with 2 to 8 years specific experience in one of the following areas: logic design and magnetic memory development; low frequency design and development involving servos; systems and component testing, resolver chains, gyros, accelerometer integrators; inertial instrument development.

RADAR

MS with 5 to 10 years research and development experience on radar equipment, pulse circuitry, feed-back circuits, CW covering full frequency. Openings also in Tucson, Arizona.

ELECTRONIC SYSTEMS

MS preferred, with minimum 5 years experience for synthesis and analysis of advanced systems such as air traffic control, target locators, visual simulators and feed-back controls. Openings also in Tucson, Arizona.

SYSTEMS ENGINEERING

BS in ME, Applied Math, AE or Physics. Experienced in the dynamics of closed-loop systems, guidance, mechanics of vehicle flight, space mechanics, analog and digital computing systems.

ba

Please address resumes to Mr. George Klock

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(Continued from page 116A)

movable cover to permit easy card installation, provide maximum protection to the operator, permit visual inspection of the card while under test, and stabilize the temperatures in the immediate card environment. Outputs are analyzed by means of a voltmeter or oscilloscope.

Fuzed Quartz

The General Electric Co., Nela Park, Cleveland 12, Ohio, has developed a new "ultra-pure" fused quartz, the first to be made in the United States from domestic raw materials, it was announced by Ernest A. Howard, general manager of the company's Lamp Glass Department. Fabrication of the new quartz into tubing and rod is now underway by the department.

The advantages of a domestic source of supply and of a high degree of purity will be of major significance to the defense, space, electronics, lamp and other industries which rely on fused quartz for critical applications, the firm stated.

Flame from torch causes tube of imported materials to sag sharply. Domestic tube is affected considerably less.

Eight years of research and development were applied to perfecting the new process of purifying the raw material, silicon dioxide, which is available domestically. Previously all silicon dioxide used for fused quartz production had to be imported, the only known sources of reliable quality being in Brazil and Madagascar.

The quartz will be particularly beneficial to the nation's research and development laboratories where, in many cases, even the smallest trace of impurities presents a problem. The semiconductor industry will also benefit from a much lower content of alumina and boron, previously a major problem in the manufacture of transistors.

Because of its high purity, the new quartz has higher thermal resistance, en-

(Continued on page 120A)

HIGH STANDARDS

The creative work of research personnel is the basic product of the Electro-Physics Laboratories. Therefore, men willing to devote their full talents and energy to achieve rapid and economical scientific progress are essential. **THE PROBLEM** The useful applications of long range propagation of electromagnetic radiation in the VLF to VHF portion of the spectrum. These applications are concerned with communications and the measurements of natural and man-made disturbances of the geophysical environment. **THE MAN** He is imaginative in the application of classical theory to today's problems. He accepts challenging technical problems with the characteristically positive approach of the successful investigator. He is enthusiastic about the pursuit of useful knowledge along uncharted pathways. If he detests mediocrity and gadgeteering . . . if he seeks the best possible solution to the fundamental problems regardless of the approach . . . then this is our man. Staff appointments are now available for personnel having, or eligible for, security clearances. For additional details, please write to:

This 100-foot high, HF helical antenna, designed and built by the Electro-Physics Laboratories, is one of the many unusual antennas used to increase our capabilities in experimental investigation of ionospheric phenomena.

Mr. William T. Whelan, Director of Research and Development

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ACF ELECTRONICS
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NEWS New Products

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(Continued from page 118A)

abling it to withstand high temperatures for longer periods of time. Tests show, for example, that the new material will deform at less than one-half the rate of the former quartz when subjected to temperatures of 1200°C and higher. In addition, tubing and rod produced from the domestic source will devitrify much more slowly, and remain more transparent, at these high temperatures.

Although changeover of all production processes will not be completed for several months, the new ultra-pure fused quartz is now available for immediate shipment in all sizes. Special shapes can be manufactured to customer specifications. Price of the improved new material, designated Type 204-A, is the same as that of the previous standard line.

DC to DC Airborne Power Supplies

Solid state DC to DC airborne power supplies have been developed by **Computer Engineering Associates,** an affiliate of **Susquehanna Sciences, Inc.,** 350 N. Halstead, Pasadena, Calif., to power telemetering transmitters or in any application requiring reliable and efficient conversion of low-voltage DC to high-voltage DC. Power supplies are completely encapsulated in a semi-flexible epoxy to provide protection against shock, vibration and environmental conditions and to meet the following tests.

High and low temperature operation from +185 to −40°F. Ten second sustained acceleration at 60 g, one shock at 50 g average acceleration, 6 shocks at 30 g average acceleration. Six minute vibration of 10.0 g (peak) from 10 to 1,000 cps, 6 hour vibration of 3.75 g (peak) at 25 cps. Temperature cycling from −65 to −160°F for 72 hours, with a sweep frequency vibration at 3.0 g (peak) while at temperature extremes. Twelve minute exposure to white noise at 150 db above 2×10^{-4} dynes per square cm. High altitude operation. 100-hour life test at +150°F with unit cycled

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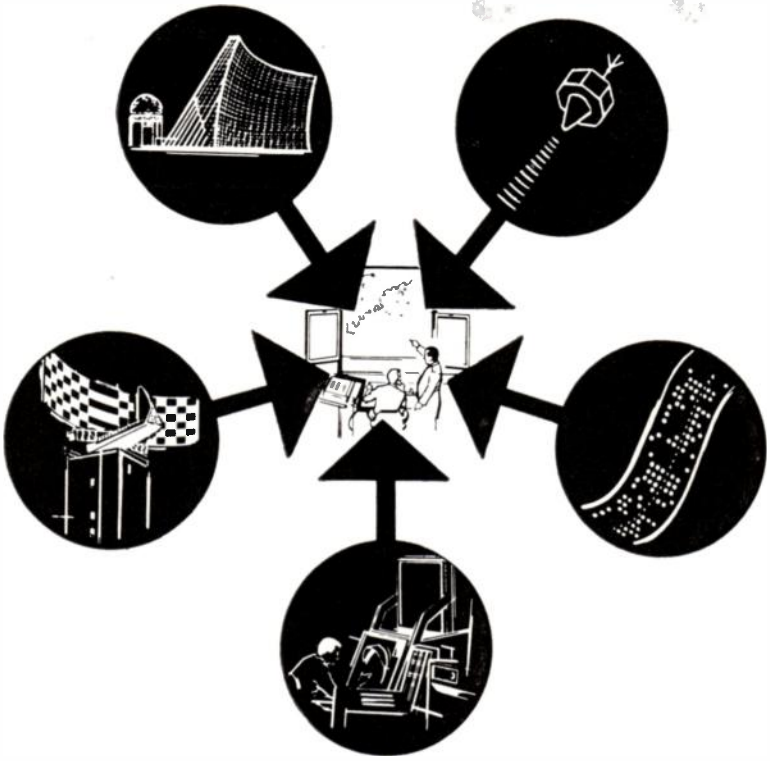

VITAL: command and control systems which collect, transmit, process and display information to military commanders responsible for compressed-time decisions.

THE
MITRE
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is now engaged in designing and developing computer-based command and control systems which collect, sort, transmit, and display information vital to those who make command decisions. SAC's Command and Control System, SAGE, BMEWS, NORAD, MIDAS and other urgent military programs are all within the scope of MITRE's system integration work.

Formed in 1958, MITRE is an independent nonprofit corporation with a growing technical staff engaged in systems planning, engineering and management for the Air Force Electronic Systems Division.

You are invited to join this scientific community and participate in important work which is performed in an atmosphere of objectivity and free inquiry.

Engineers and scientists interested in the vital field of command and control technology may inquire about openings in:

- **Operations Research**
- **Communications**
- **Human Factors**
- **System Cost Analysis**
- **Econometrics**
- **Radar Systems and Techniques**
- **System Analysis**
- **Advanced System Design**
- **Computer Technology**
- **Mathematics**
- **Air Traffic Control System Development**
- **Antenna Design Microwave Components**

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- Systems Analysis • Bionics
- Space Communications Systems
- Information Theory • Acoustics
- Thermodynamics • Aeronautics
- Horizontal Path Atmospheric Scintillation • Antenna Systems
- Advanced Detection Principles and Systems • Infrared Homing
- Multilayer Interference Filters
- Differential Gas Absorption Spectroscopy • Radar Signal Processing • Transistor Circuitry
- Electro Optical Information Collection • Microelectronics

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FOR FURTHER DETAILS WRITE TO
PROFESSIONAL EMPLOYMENT
MANAGER
JOHN HAVERFIELD
OR CALL
JEFFERSON 4-6000

MELPAR INC
A SUBSIDIARY OF WESTINGHOUSE AIR BRAKE COMPANY
3000 Arlington Blvd.
Falls Church, Va.

An equal opportunity employer.

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 120A)

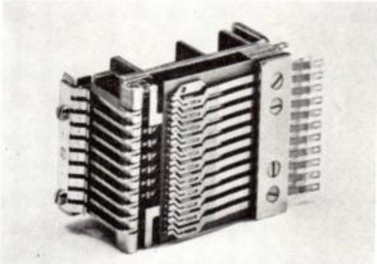
on-off. Units can also be designed operable to temperatures up to +260°F to various other required specifications. Delivery: 8 weeks. For further information, write to Max Polling at the firm.

Polonium Alpha Sources

Hazleton Nuclear Science Corp., 4062 Fabian Way, Palo Alto, Calif., has available a catalog sheet which describes the polonium alpha sources which they have recently placed on the market. Uses are: effective ionization, dissipation of electrostatic changes, and instrument calibration. The mica window permits 70% of alpha transmission. Range is 1 microcurie to 50 millicuries.

Electro-Magnetic Pulse-Counter

The Semiconductor Dept., International Telphone and Telegraph Corp., Components Div., has announced the availability of the type ZM-53 electromagnetic pulse counter. This relay-like device has 10 armatures operating sequentially on application of a pulse series. Precious-metal contacts are closed by the armatures.



The nominal minimum pulse duration for operation of the device is 20 milliseconds and the minimum interval between pulses is also nominally 20 milliseconds. The actuating winding of the counter is available in both 24- and 60-volt designs with corresponding current ratings of 650 and 324 milliamperes. At the conclusion of each counting sequence, the device is returned to zero by application of a 40-millisecond pulse to a release winding.

(Continued on page 124A)

GOODYEAR
AIRCRAFT
CORPORATION
Arizona Division

is expanding its Electronic facilities and has long range positions that will challenge your talent in the technical areas listed below:

SR. DEV. ENG'R.
Develop wideband i-f amplifiers, i-f oscillators, i-f isolation amplifiers, & related circuits.

SR. DEV. ENG'R.
Develop X-band radar circuits & equipment including wave guide assemblies, & related circuits.

SR. DEV. ENG'R.
Servomechanisms, design close loop control system.

SR. DEV. ENG'R.
Develop radar transmitters & modulators including pulse forming networks, transformers, protection circuits, X-band amplifiers & related circuits.

SR. PACKAGING ENG'RS.
Layout packaging & detailing layout of electronic sub-assemblies including i-f and r-f units.

All qualified applicants will receive consideration for employment without regard to race, creed, color, or national origin.

Request Application or Send Resume To:

B. A. Watts, Engineering Personnel
Goodyear Aircraft Corporation
Litchfield Park, Arizona

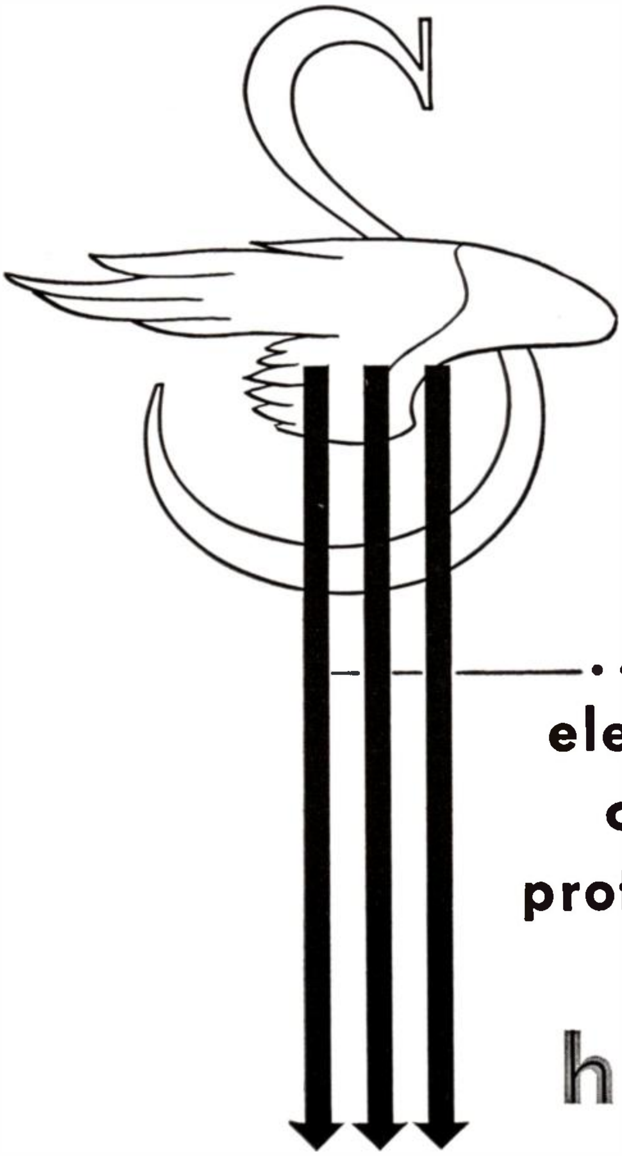
Similar Positions at Goodyear Aircraft Corporation, Akron, Ohio.

ENGINEERING MANAGER

Continued expansion by the Automatic Equipment Division has created an outstanding opening for a graduate electrical or mechanical engineer preferably with a master degree to be responsible for all engineering activity including development, design and application engineering for a newly organized product department; automated electro-mechanical industrial machinery. Must have at least 10 years management experience with overall project level responsibility for the design and development of highly engineered systems. Should have a strong background in several of the following areas: logic circuitry, servo networks, hydraulic servo controls, tape recorders. Must also be experienced in the design of practical hardware for such systems, and should have demonstrated leadership and managerial capabilities. *This division is located in a Chicago suburb and is a major new industrial machinery division. A.M.F., a leader in the manufacture of automated machinery, offers an exceptional opportunity for personal and professional growth.* These positions offer many excellent employee benefits and a salary commensurate with experience. Please reply in confidence to:

MR. R. B. MARTIN
Personnel Manager
AMERICAN MACHINE
& FOUNDRY COMPANY
Automatic Equipment Division
2100 East Lunt Avenue
Elk Grove Village, Illinois

pioneer and leading manufacturer
of rotary wing aircraft



...at SIKORSKY electronic engineers can broaden their professional horizons in the field of helitronics

Today, as Sikorsky Aircraft's activities expand in significance and scope, particular emphasis is being directed to a sphere of activity which we term "*helitronics*". This area embraces a blending of two major technologies: *heli*copters and elec*tronics*. Specifically, helitronics means the integration of guidance and navigation systems, specialized electronic search and detection equipment to enhance the mission capability of the helicopter; specialized sensors and automatic controls to increase its versatility as an optimum military weapon system and a commercial carrier.

The need for more sophisticated electronic systems offers exceptional opportunities to competent electronic engineers with particular skills in: design • instrumentation • test • development • air-borne systems • production and service support equipment • trainers and simulators.

Unusually interesting openings as Avionics Instructors also exist for men with aircraft electronics experience and a desire to teach.

If you are interested in these career opportunities, please submit your resume, including minimum salary requirements, to L. J. Shalvoy, Personnel Department.

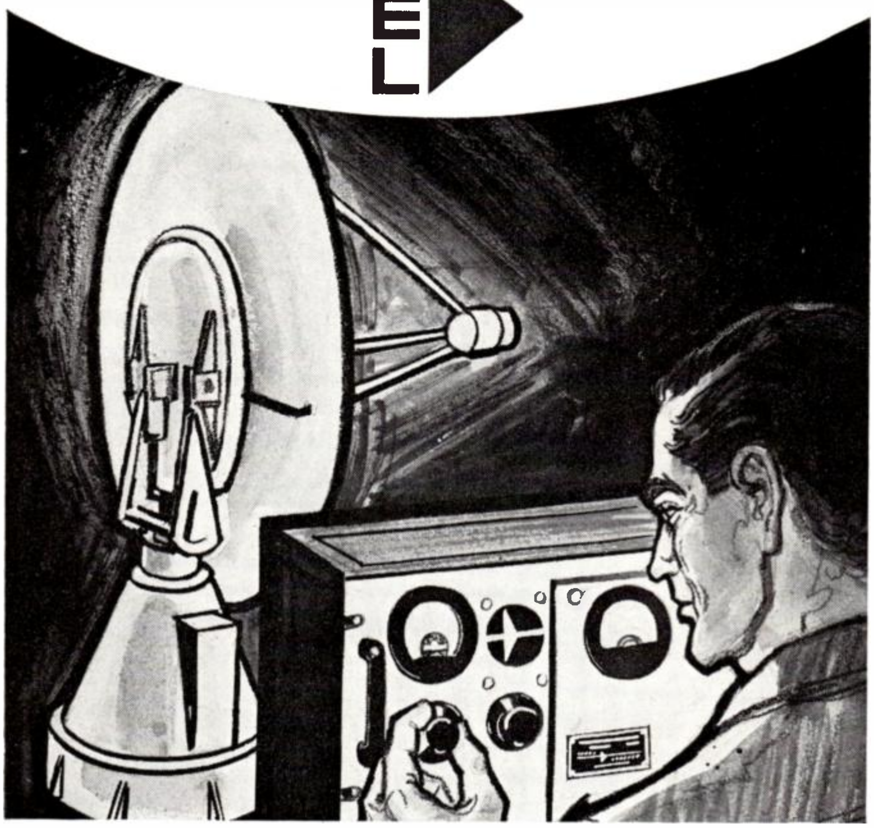
SIKORSKY AIRCRAFT **STRATFORD, CONNECTICUT**

DIVISION OF UNITED AIRCRAFT CORPORATION

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

GEL Advances the Art of Microwave Telemetry Systems

GEL



Sophisticated research and development programs in Microwave Telemetry in the 1500 to 2200 mc region are now underway at General Electronics. Now, increasing workloads demand additional highly qualified, experienced engineers for junior and senior positions in this, and other fields. This is an opportunity to advance the art, and yourself, in a company guided and managed by engineers. Openings are in suburban city locations at GEL facilities in Cambridge, Mass., and Silver Spring, Md. Many excellent benefits include profit sharing, insurance, tuition refund programs, and the opportunity to join a company with an average 50% annual growth rate over the past six years.

MICROWAVE SYSTEMS AND ANTENNA ENGINEERS— Experienced in design and development of microwave antennas, receivers and components such as multipliers, mixers, filters and cavities.

TRANSMITTER DESIGN ENGINEERS— Experienced in design and development of high-powered HF and VHF Transmitters with some experience in FM Transmitters desirable.

RECEIVER ENGINEERS— Experienced in design and development of VHF and/or UHF Receivers. Solid-state experience using devices such as varactors, transistors and diodes desirable.

AUDIO ENGINEERS— Experienced in design of audio and tape transport systems. Solid-state and printed circuit experience helpful.

PARAMETRIC MULTIPLIER DESIGN ENGINEERS— Experienced in development of solid-state devices associated with varactor frequency multipliers.

Write or call: Mr. B. T. Newman, UNiversity 4-8500

GENERAL ELECTRONIC LABORATORIES, INC.

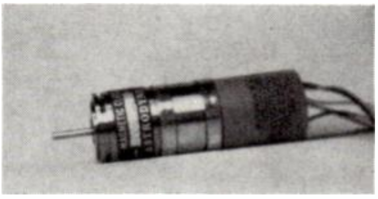
195 Massachusetts Avenue, Cambridge 39, Mass.

An Equal Opportunity Employer.

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

(Continued from page 122A)

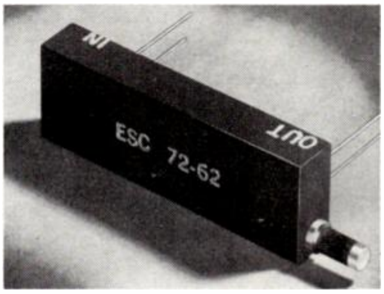
Clutched Synchro



Integral clutch, spring return mechanism and synchro has been developed by **Astrodyne Inc.**, 121 Clinton Rd., Caldwell, N. J. The size 10 package contains an electromagnetic clutch, spring return mechanisim and synchro for computer applications. The return mechanisim is set to ±1 minute of electrical zero of the synchro and repeats synchro position to within ±1 minute. The synchro can be rotated through 360°. Units are also available in size 8 or 11 or with potentiometer and commutator switches.

Miniature Trimmer Delay Line

ESC Electronics Corp., 534 Bergen Blvd., Palisades Park, N. J., has announced a new trimmer delay line with wide applications where miniaturization is a factor.



With a delay time of 0.05 μsec minimum at maximum position, the unit is 2.25" in length, 0.35" wide, and 0.75" deep (not including leads). It has a terminating resistance of 1000+10%. Its output rise time is 0.05 μsec maximum with input rise time a step function. The dielectric strength is 500 VDC.

For further information write to the firm.

Miniaturized Power Packs

Electronic Research Associates, Inc., Cedar Grove, N. J., announces the availability of a new two-page bulletin which contains detailed mechanical data covering their complete line of Transpac miniaturized power packs.

Included are case size data, tolerances,

(Continued on page 126A)

ELECTRICAL ENGINEERS

B.S. DEGREE

Electrical engineer with experience in electronic test equipment, DC amplifiers, FM equipment, magnetic tape recorders, pulse circuitry, and electrical control system.

Work would include part or all of the following jobs:

1. Maintaining and trouble shooting a data acquisition system used in rocket testing.
2. Modifying and improving existing instrumentation, which includes analyzing individual problems, recommending possible solutions, obtaining equipment for prototype fabrication, checking and modifying and turning completed project over to operations.
3. Setting up and operating special instrumentation tests such as hydraulic testing, dynamic balancing, vibration equipment, and frequency analysis.

HERCULES has immediate openings at the nation's foremost rocket research and development facility at Rocket Center, West Virginia— Allegany Ballistics Laboratory.

All inquiries will be acknowledged promptly. Write to
DR. W. R. LOWSTUTER, Technical Personnel Dept. H.

HERCULES POWDER COMPANY
INCORPORATED
Cumberland, Maryland

All qualified applicants will receive consideration for employment without regard to race, creed, or national origin.

U. S. CITIZENSHIP REQUIRED.

Increased technical responsibilities in the field of range measurements have required the creation of new positions at the Lincoln Laboratory. We invite inquiries from senior members of the scientific community interested in participating with us in solving problems of the greatest urgency in the defense of the nation.

RADIO PHYSICS and ASTRONOMY

RE-ENTRY PHYSICS

PENETRATION AIDS DEVELOPMENT

TARGET IDENTIFICATION RESEARCH

SYSTEMS:
- Space Surveillance
- Strategic Communications
- Integrated Data Networks

NEW RADAR TECHNIQUES

SYSTEM ANALYSIS

COMMUNICATIONS:
- Techniques
- Psychology
- Theory

INFORMATION PROCESSING

SOLID STATE
- Physics, Chemistry, and Metallurgy

• *A more complete description of the Laboratory's work will be sent to you upon request.*

All qualified applicants will receive consideration for employment without regard to race, creed, color or national origin.

Research and Development
LINCOLN LABORATORY
Massachusetts Institute of Technology
BOX 16
LEXINGTON 73, MASSACHUSETTS

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 124A)

terminal connections, mounting information, and related data. Drawings covering case sizes and terminal orientations are also included.

The catalog sheet has been given the designation of Bulletin #102—Mechanical Data sheet.

Size 11 Synchros

Kearfott Div., General Precision, Inc., 1150 McBride Ave., Little Falls, N. Y., has announced the production of a line of Size 11 Bu/Ord synchro specification MIL-S-20708A. These transmitters, control transformers, control differential transmitters, and torque receivers are employed in systems used for fire control, radar, navigation, missile functions, and other applications requiring a high level of precision, endurance, and reliability. Each of these units operates in an ambient temperature range of −55°C to +80°C and each weighs 4.7 ounces. Other mechanical data includes: rotor moment of inertia of 2 gm cm^2, friction at 25°C of 0.05 in./oz. (maximum), friction at −55°C of 0.07 in./oz. (maximum). Electrical characteristics differ with specific units, and synchros having accuracies of five or seven minutes are available.

Measurements of VSWR In Coaxial Systems

PRD Electronics, Inc., subsidiary of Harris-Intertype Corp., 202 Tillary St., Brooklyn, N. Y., manufacturer of microwave and electronic test instruments, has released a new 8 page PRD Report entitled "Measurement of VSWR in Coaxial Systems" by Leonard Sweet and R. A. Lebowitz.

The paper describes in detail the current methods used in obtaining standing wave measurements of microwave components, with emphasis on coaxial systems. By informative diagrams, graphs, and photographs of coaxial test equipment, this report describes the common types of coaxial lines used in systems, such as rigid air lines and solid dielectric cables. Numerous coaxial connectors, including Types N, C, BNC, HN, LC, LT, UHF, miniature and others, are also described. In various sections of this publication, the

(Continued on page 128A)

ENGINEERING PHYSICIST

Challenging position offered by dynamic young organization, located in Central New Jersey.

Successful applicant would participate in amply funded operation of exceptionally well equipped R & D Laboratory, concerned with mechanical movement or displacement produced by the interaction of electrical, magnetic or electrostatic fields in microminature sizes. Experience or knowledge of fundamental design aspects with similar problems (for example miniature relays, choppers, switches, etc.) an asset.

For interview please send brief resume to Personnel Department, P.O. Box 483, Princeton, New Jersey or call collect WAlnut 4-5990.

ENGINEERS for MINUTEMAN MISSILE SITE in Montana

Space Technology Laboratories, Inc., responsible to the Air Force for the systems engineering and technical direction of ICBM weapon systems programs, has immediate opportunities for qualified engineers to serve as systems test advisors in the system test and integration of the Minuteman missile system at Malmstrom AFB, Great Falls, Montana.

Requirements include BSEE or BS in Physics, or equivalent, with 5 or more years experience in the design, development and installation of communication, control, computer, or data processing systems.

To investigate these vital, challenging opportunities, please write Dr. R. C. Potter, Manager of Professional Placement and Development. STL is an equal opportunity employer.

SPACE TECHNOLOGY LABORATORIES, INC.
P.O. Box 95005I
Los Angeles 45, California
a subsidiary of
Thompson Ramo Wooldridge Inc.

ANTENNA & PROPAGATION ENGINEERS: To carry on theoretical & experimental studies of advanced nature for all radio frequency bands. EE's. (Advanced degree preferred.) Minimum of 6 years experience. Ⓒ

Senior Staff Consultant

for all System Level Experimentation in Electronics for all MSVD programs

System problems include: stabilization and attitude control of satellites and re-entry vehicles; space/ground communications; orbit navigation and control; mid-course and terminal spacecraft navigation and guidance.

An advanced degree is essential for this outstanding opportunity for high level, original work; also a substantial background in systems design. Experience in radar & communications helpful. Ⓔ

MANAGER

ADVANCED MISSILE & SATELLITE ELECTRONICS

A Significant Opportunity for an Engineer with a Record of Achievement. An advanced degree – MS or PhD and a minimum of 10 years applicable experience is mandatory for the highly competent and creative man required for this position. He will guide an organization responsible for the development of very advanced components for missile & satellite applications. These include incoders, command circuitry, transmitters, digital electronics, etc.

Ⓠ

Ground Support Electronic Systems Engineer

Design and development of ground electronic systems through the RF spectrum for application in the support systems of re-entry or space vehicles. A technical specialist position requiring a BSEE, 6 years experience and a background in such fields as communications, telemetry or radar.

Ⓨ

ENGINEERS & SCIENTISTS

IMMEDIATE OPENINGS AT THE MISSILE & SPACE VEHICLE DEPARTMENT

PHILADELPHIA, PENNSYLVANIA

Your inquiry is invited regarding any of the positions — encompassing many areas of space science and engineering — listed on this page. Please use the INQUIRY CARD provided for your convenience. A prompt answer is guaranteed; your confidence will be respected.

Analog Computer Specialists

To plan sequence of computer operations and determine circuitry for engineering problems connected with research, development and design of advanced satellite, spacecraft and missile systems. Set up and operate computer programs in a growing analog facility equipped with Electronic Associates and Reeves Analog Equipment, a combined Analog-Digital facility, and passive element analog computer.

Ⓓ

Diversified Assignments for Design Engineers

Logic Circuit Design

Provide high level technical evaluation of digital techniques as applied to automatic programming systems for check-out equipment and airborne instrumentation. Requires EE with broad digital background from systems philosophy to final design. Ⓢ

Digital Components & Circuits

Design and evaluate A/D and D/A converters, multiplexers, parity check generators, digital storage devices, binary counters, etc., for both airborne and ground digital equipment. EE with 2-5 years experience. Ⓣ

Diversified Assignments in Systems Engineering

SYSTEMS DESIGN
Navigation & Control

Challenging opportunity for engineer capable of assuming responsibility for integration, coordination and project engineering of special navigation and control systems. These include satellite attitude control, orbit control, space power. Work requires a broad background in controls; 5 or more years engineering experience; EE or ME. Ⓕ

SYSTEMS ANALYSIS
Varied Space Systems

Conduct preliminary studies to optimize space systems through sub-system trade-offs, in order to provide a balanced systems synthesis effort. Define performance requirements and specific interface relationships between sub-systems. Degree plus 4 or more years experience in systems analysis required.

Ⓖ

RADAR SYSTEMS
Re-Entry Guidance & Space Navigation

Apply radar systems principles to problems of re-entry-guidance and space navigation; define parameters of the technical concept and maintain project control over full engineering effort. Requires strong background in radar systems with EE.

Ⓗ

GENERAL ELECTRIC

All qualified applicants will continue to receive consideration for employment without regard to race, creed, color, or national origin.

PROFESSIONAL EXPERIENCE (Two most recent or most applicable jobs)

Company ____________________ Position ____________________

Years of experience (from) __________ (to) __________ Salary __________ Assigned Duties __________

____

____

Company ____________________ Position ____________________

Years of experience (from) __________ (to) __________ Salary __________ Assigned Duties __________

____

PERSONAL DETAILS

Name ____________________

Home Address ____________________ City ____________________

State ____________________ Telephone ____________________ U.S. Citizen Yes ☐ No ☐

Education Undergraduate ____________________
College Degree Year of Graduation

Graduate ____________________
College Degree Year of Graduation

Position Objective (primary job preference, level of responsibility, salary expectations) ____________________

____

____

I am interested in the following position(s)

Navigation & Control	E, F, H
Instrumentation & Communication	C, E, Q, S, T
Advanced Systems	F, G, H
Ground & Space Support	Y
Analog & Digital Techniques	Q, S, T
Applied Mathematics	D
Systems Analysis	G

▶ Simply circle the appropriate letters of those positions that meet your professional interests and qualifications, fill in the questions to the left — attach a resume if you have one handy — and airmail to:

Mr. Frank Wendt
Missile & Space Vehicle Department
General Electric Company
3198 Chestnut Street
Philadelphia 4, Pennsylvania

GENERAL ELECTRIC

464A

Scientists • Engineers

BASIC STUDIES IN
COMMUNICATIONS
TECHNIQUES
...at Sylvania's
APPLIED RESEARCH LABORATORY

At ARL, basic research in applications of statistical communication theory offers liberal opportunities for original work. As the central research facility for Sylvania Electronic Systems, ARL's multi-disciplined staff covers an unusual range of short and long range studies in commercial and military areas.

Pictured above is the new laboratory facility, located in Waltham, Massachusetts, convenient to the academic environment of the Boston area, and the suburban residential areas.

Current *active* research projects in Communications Techniques include:

- **Secure Communications Techniques**
- **Controlled Data Rate, Communications Feedback Systems**
- **Adaptable Communications Techniques**
- **Modulation Techniques for Satellite Relay Systems**
- **Communication Reconnaissance Systems**

☐ Laboratory research programs are carried out under both Government contracts and, to a very substantial extent, under internally-directed research funds.

☐ The Laboratory's staffing reflects a high degree of experience and/or academic training. Of the 80 scientific personnel presently employed in the Laboratory, over one-third have PhD degrees.

☐ If you hold a graduate degree or have equivalent experience in an applicable technical field, the advantages of a position with the Applied Research Laboratory merit your serious consideration...

• Problems of Major Magnitude in Advanced Areas • Informal Professional Atmosphere and Extensive Opportunity for Creative Research • Ample Opportunity and Encouragement to Publish • Liberal Opportunities for Advanced Education.

☐ Communication scientists and engineers at ARL enjoy important inter-disciplinary relationships with other research groups which include Mathematics, Radio Physics, Information Processing, and Engineering Research.

Please direct your inquiry to Dr. Seymour Stein, Manager, Communications Research Department

APPLIED RESEARCH LABORATORY
SYLVANIA ELECTRONIC SYSTEMS
Government Systems Management
for GENERAL TELEPHONE & ELECTRONICS

An Equal Opportunity Employer

40 Sylvan Road – Waltham 54, Massachusetts

ENGINEERS
for
CAPITRON DIVISION
Elizabethtown, Penna.

Send complete resume to T. J. Sullivan

AMP INCORPORATED
Harrisburg, Pennsylvania

An Equal Opportunity Employer

ELECTRICAL ENGINEER
(Power Supplies)

Will develop and design Solid State Hi Voltage power supplies to 30 KV. B.S. in E.E. or equivalent. Experience with power supplies, magnetic amplifiers and transistor circuitry helpful.

DEVELOPMENT ENGINEER

B.S. in E.E. or Physics, with experience in field of design and development of new products. Capable of doing creative work in the electronics field.

CAPACITOR DESIGN ENG'R

B.S. in E.E. or Physics preferred, for design in product engineering of rolled paper, Milar and Teflon capacitors, Isamica and Bentonite clay film capacitors.

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

(Continued from page 126A)

philosophy, definition and measurement of VSWR are compared and discussed. A section on applications covers the fundamental uses of equipment commonly employed in standing wave measurements.

The report also covers the latest progress made in VSWR measurement. It defines the terms used and discusses new techniques for making standing wave measurements. The errors inherent in the use of a slotted section, such as residual VSWR, slope, and irregularities, are also described. A bibliography is included.

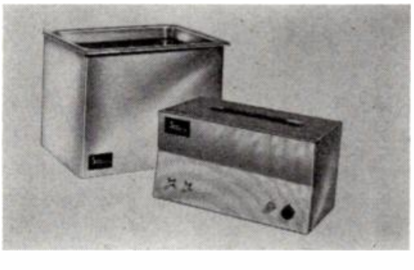
Copies of this latest issue of "PRD Reports," Vol. 7, No. 3, entitled "Measurement of VSWR in Coaxial Systems" is available free from the firm.

Kerr Cell Applications

Electro-Optical Instruments, Inc., 2612 E. Foothill Blvd., Pasadena, Calif., has available several data sheets on their standard Kerr cell instruments. The firm has information available about the new uses of Kerr cell instrumentation in such experimental research as: Electrical discharges in gases; Electrically exploded wire phenomena; High-speed ballistic range photography; Hypervelocity particle photograph; Light sources for optical doppler systems; Lightning stroke studies; Optical modulators for Lasers; semiconductor carrier lifetime test sets; Single crystal explosive detonation phenomena; Thermonuclear plasma studies; Ultra-high-speed shock tube photography; and Visual physiological testing and research.

Self-Tuned Ultrasonic Cleaner

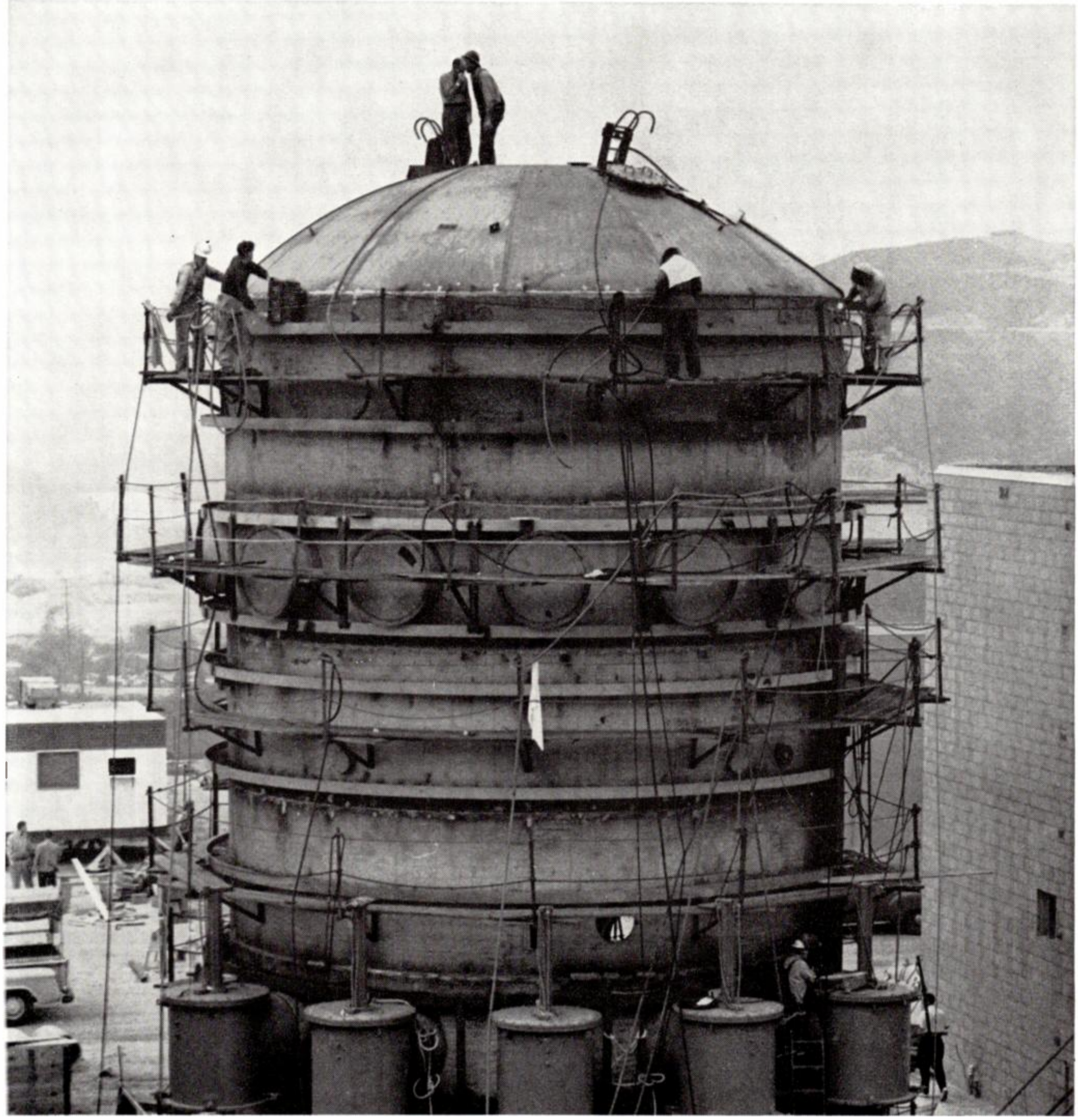
The 3-gallon multisonic Model MSS 1500 series produced by **Sonic Systems, Inc.,** 1250 Shames Drive, Westbury, N. Y., combines self tuning, automatic operation, high reliability and ruggedness at the cost of some one gallon systems.



The generator has 150W average power and 600W peak. High efficiency, half wave self rectifying circuit scans transducer producing even cavitation throughout complete cleaning tank. Generator dimensions are 13″×10″×8½″.

Transducer tank MST 1501 with ID of 19½″×12¼″×6″ deep can operate at temperatures of 230° F permitting use of boiling aqueous solutions. The price of the system is $449.

(Continued on page 130A)



Putting outer space in Pasadena

How will sub-zero cold and intense heat and light from he Sun effect planet-bound spacecraft moving through the void of outer space? The scientists and engineers at Cal Tech's Jet Propulsion Laboratory think they know.

But they're building this giant space simulator at JPL's Pasadena facility just to make sure.

The 80-foot high simulator will soon begin environmental tests on the Mariner—first spacecraft scheduled to fly-by Venus. In a 25-foot-in-diameter chamber, the Mariner may be exposed to a vacuum of 5×10^{-6} millimeters of mercury, a wall temperature of −320°F, and a radiant flux of sun light and heat found as near to the Sun as Venus and as far away as Mars.

Next-to-the-real-thing testing is part of any R & D work. But in JPL's job—exploration of our Moon and planets—the stakes were never higher.

The odds for JPL's (and the nation's) success in space are only as good as our scientists and engineers are good. JPL is looking for good people. The best people. Why not write to us and find out if you can improve the odds.

JET PROPULSION LABORATORY
4804 OAK GROVE DRIVE, PASADENA, CALIFORNIA
Operated by California Institute of Technology for the National Aeronautics & Space Administration

All qualified applicants will receive consideration for employment without regard to race, creed or national origin / U. S. citizenship or current security clearance required.

NEWS New Products

(Continued from page 128A)

Function Generator

The newly developed Type 250 function generator introduced recently by **Exact Electronics, Inc.**, P. O. Box 234, Hillsboro, Oregon, makes available five different waveforms–square, triangular, sine, cosine, and ramp–with independent or simultaneous output of all five over a frequency range of from 0.001 to 10,000 cps.

The unit features external triggering in addition to manual triggering and continuous operation; delivers 30 v at 25 ma at the main output, or 30 v at 5 ma at auxiliary outputs. Reference level is variable for above and below-ground waveforms. Amplitude variation over the instrument's 10-million-to-one timing range is said to be under 5%.

Flexibility of the Exact Type 250 suits it for use as a general research laboratory tool as well as in such fields as servo work, analog computer function generation, medical applications, and mechanics.

The Type 250, complete with operating accessories and instruction manual, is priced at $600.

Detailed specifications can be obtained from the manufacturer.

Program Analyzer

Applied Data Research, Inc., 759 State Rd., Princeton, N. J., has developed an IBM 709–7090 program "analyzer." The analyzer examines all instructions in an IBM 709–7090 program and produces a list, cross referencing each operand address with the instruction and location affecting it.

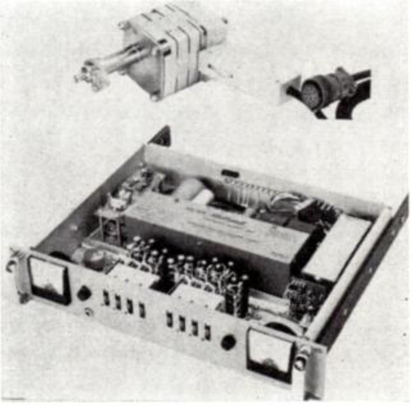
A program analysis is useful in locating errors during program checkout and in making changes to instructions and constants during and after debugging.

Virtually an entire 32,000 word memory can be analyzed at one time at output speeds of 640 instructions per minute. The program can produce a listing for successive memory locations or for only those locations being referenced.

A binary program deck, operating instructions and sample listing can be obtained without charge, from the firm.

Signal-Conditioning Equipment

Astra Technical Instrument Corp., 9905 W. Jefferson Blvd., Culver City, Calif., has announced a new line of accessories specifically for use with signal-conditioning circuitry: the Calibration Monitor Panel Model CMP-1 and Servo Gun Model SG-1. These provide semi-automatic balancing of strain gages; semi-automatic biasing of thermocouples; selectable, six-range null voltages; and automatic decimal point indication. The CMP-1 provides a convenient monitor for high- and low-level signals in the same instrument at the same time. The easily read and used unit has extreme reliability and very great accuracy through digital readout switching. The hand-held servo gun consists of a



(Continued on page 132A)

Philco WDL has immediate and challenging engineering positions open in Discoverer, Midas and Advent programs in these categories:

- Data systems engineering
- Technical Staff — communications and data systems
- Systems test planning
- Systems reliability analysis
- Design Engineering — surface and vehicle electronics
- Human factors and operations analysis
- Field tracking station activation
- Tracking and control systems design
- Logistics and station support
- And others

U. S. citizenship or current transferable Department of Defense clearance required.

if your future is for growing...

People at Philco's Western Development Laboratories are proud of their achievements, secure in the steady growth of their company, enthusiastic about their future.

In just four years, Philco WDL has expanded from a staff of 18 to more than 2,000. At the end of this year WDL will open a new 250,000-square-foot facility which will accommodate a continually expanding complement of engineers, scientists and supporting staff members.

Philco WDL, the space organization which designed and built the Courier satellite, conducts an ever-growing development program in tracking, satellite instrumentation, communications, data processing and command—moving hand in hand with the federal government in space exploration and space age defense.

Your growth with a growing company, ideal living on the Northern California Peninsula, professional and monetary advancement commensurate with your own ability—these are some of the advantages and satisfactions of working at Philco Western Development Laboratories.

If yours is one of the fields listed at the left, write today to Mr. W. E. Daly, in confidence, of course, Department R-1.

This new 250,000-square-foot addition to Philco Western Development Labs is further indicative of your future there.

EQUAL EMPLOYMENT OPPORTUNITY
FOR ALL QUALIFIED APPLICANTS

PHILCO
Famous for Quality the World Over

WESTERN DEVELOPMENT LABORATORIES
3875 Fabian Way, Palo Alto, California

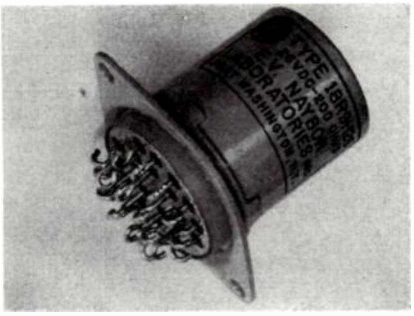
NEWS New Products

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

(Continued from page 130A)

housing and balancing servo motor with a no-load speed of 144 rpm and stall torque of 23 ounce-inches. The SG-1 is always used with the CMP-1 and any signal-conditioning equipment with a driven control of the proper mechanical configuration. Insertion of the gun automatically couples the output of the signal-conditioning equipment to the CMP-1. The servo motor will not operate until the pisto-grip trigger is pressed. Request further information from R. W. Smith, at the firm.

General Purpose Relay



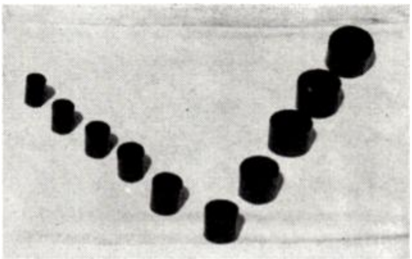
E. V. Naybor Laboratories, Inc., Port Washington, N. Y., has developed the 'R' series relays, general purpose subminiature types designed and manufactured in accordance with military specification MIL-R-5757D.

A new type of rotary motor, for which patent application has been made, actuates the multiple contact arrangements. The high torque developed by this relay results in higher contact pressures, thus providing greater reliability.

Specifications

Contact Arrangement—Two-four-and six-pole, double throw at two amperes resistive, or one ampere inductive, or eight amperes overload. These ratings are for voltages of 26.5 vdc or 115 vac; 60 to 400 cps. Coil—standard coil resistance for 6 PDT relay is 200 ohms for 26.5 volt operation. Other resistance values are available from 6 to 14,000 ohms. Temperature Range—From −65°C to +125°C. Vibration—15 G's to 2,000 cycles. Shock—50 G for 11 milliseconds. Altitude—To 70,000 feet. Dielectric—Test at 1,000 Volts RMS. Weight—2.6 ounces.

Epoxy Case Shells



Cycle Products Co., Inc., 123 Central Ave., Newark 2, N. J., is manufacturing a standard line of epoxy shells for electronic component packaging. Material is epoxy mineral filled *non-burn* is standard and available from stock, other epoxy materials and other plastics such as silicone, diallyl phthalate and phenolic are available on special order.

Colors are black which is the standard stocked color, red, green, blue, tan orange are available on special order.

Shapes and sizes. Other shapes, square, rectangular, and so forth, and sizes can be moulded on special order.

Standard OD	Stocked ID	Sizes Height
.562	.492	.750
.625	.547	.750
.688	.610	.750
.750	.672	.750
.813	.753	.750
.875	.797	.750
1.062	.984	.750
1.188	1.126	.750
1.250	1.172	.750
1.312	1.235	.750

Silicon Diode Catalog

A new catalog describing a line of silicon double diffused alloy diodes for use in core driving, clamping, gating and similar circuits in high speed digital computers, has just been announced by **National Transistor Mfg., Inc.**, 500 Broadway, Lawrence, Mass.

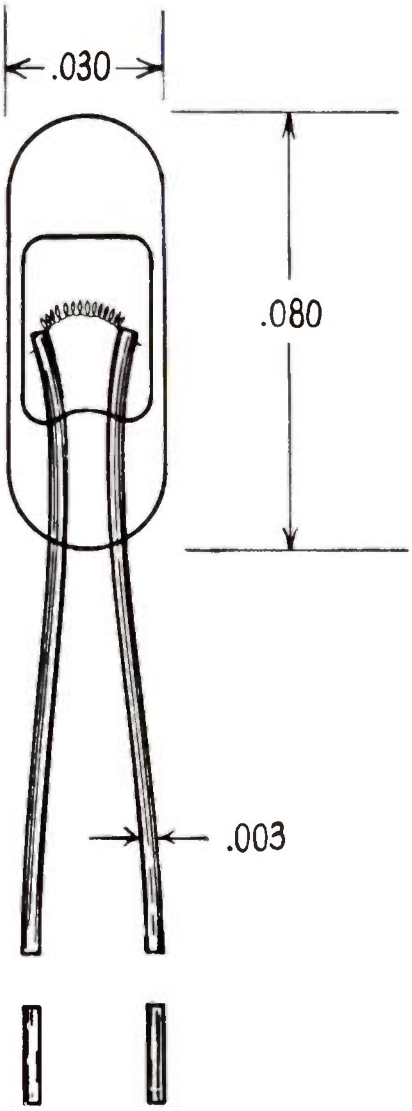
Conductances of the order of 500 ma at 1.0 volt, and switching speeds of 500 ma at 15 millimicroseconds are typical of the diodes listed.

In addition, these diodes will withstand 60,000 g's centrifuge for 3 minutes acceleration and 150 pounds per square inch for 30 minutes in an aqueous solution. They are encased in a one piece subminiature glass envelope.

The new catalog B-101 can be obtained by writing to the firm.

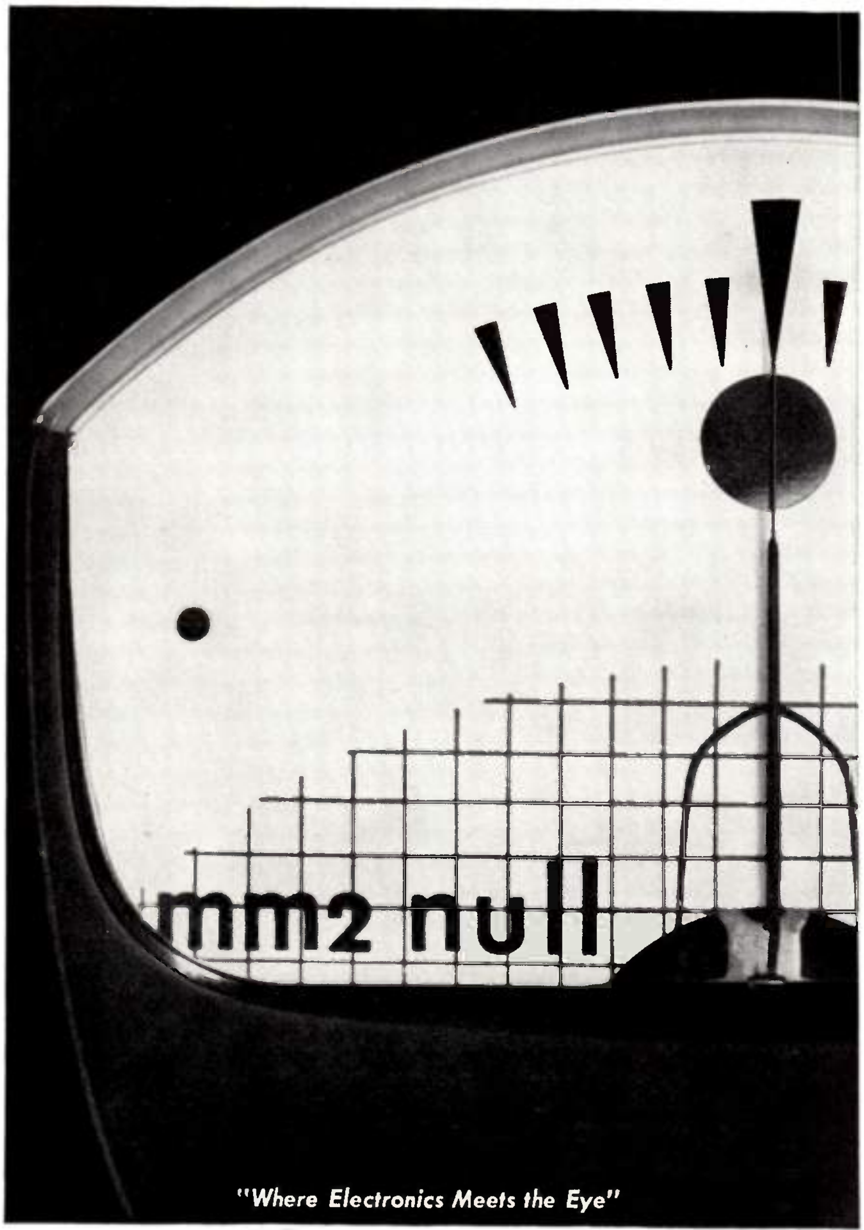
Incandescent Spotlight

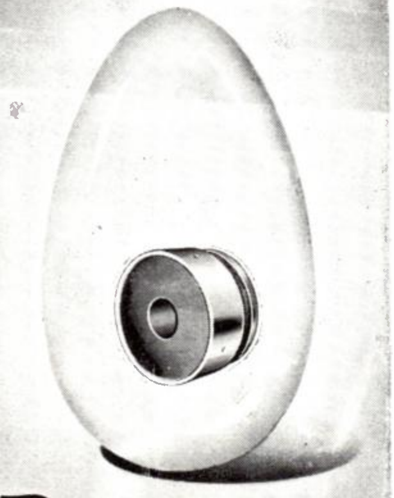
The Pinlite, "Model Lens," believed to be the world's smallest incandescent spotlight, has been announced by **Kay Electric Co.**, Dept. PI, Old Bloomfield Ave., Pine Brook, N. J., manufacturers of microminiature lamps and precision electronic test instruments.



Measuring 0.080" long by .030" in diameter, the new lamp utilizes new manufacturing techniques that permit a built into lens effect. The directional characteristics of the light source intensify the projected light to provide a microminiature

(Continued on page 134A)





NEWS New Products

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(Continued from page 133A)

spot. Up to ten times as much light is concentrated in the direction of the beam.

Total light output is 60 millilumens.

Aside from its general applications as an optical point source, other potential areas are in readout devices, medical and industrial light probes, transistor indicators, and pointer and meter scale illuminators. In addition, low thermal time-constant provides a fast light modulating capability and thus application to audio frequency sound recording.

Specifications: operating voltage (volts) —1.2; operating current (milliamperes)—12; diameter (inches)—0.030; length (inches) 0.080; and a life expectancy (hours)—1000.

Literature and prices may be obtained by writing the manufacturer.

FM Antenna Catalog

Andrew Corp., P.O. Box 807, Chicago, Ill., manufacturers of antenna systems, has published a new catalog on their "Multi-V" FM antenna system. The Multi-V antenna was introduced by Andrew 10 years ago and is currently being used by many FM stations. When used in conjunction with HELIAX, flexible coaxial cable, the Multi-V provides a matched system with mimimum VSWR across the entire channel. The system offers complete flexibility for multiplexing.

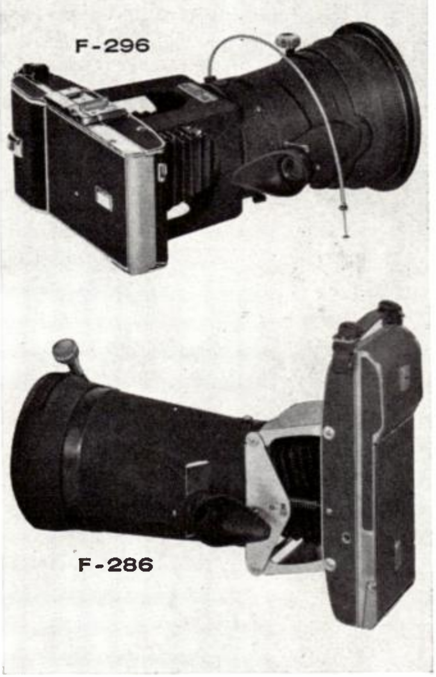
All electrical specifications shown in the Andrew Catalog B are based on measured data. The catalog is available by writing to the firm.

Quast Appointed by E D P Corp.

The appointment of Harold R. Quast as manager of Systems Engineering for the **E D P Corporation**, 3501 S. Orange Blossom Trail, Orlando, Fla., specialists in electronics, instrumentation and systems research, was announced recently by William Vassar, E D P president.

A graduate of Tri-State College where he received a B.S. degree in radio engineering, Quast has many years of design experience with missile instrumentation, tracking and data handling systems.

Quast's appointment marks E P D Corporation's expansion into the field of large electronic instrumentation and data handling systems. He will be responsible for developing the "Systems Approach" to engineering problems, in which facets of a task, from initial engineering concept



through final equipment fabrication and installation, are continuously coordinated on a management level.

Spectrofluorometer Bulletin Offered

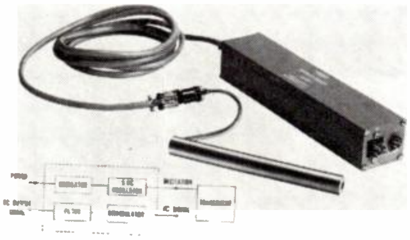
Farrand Optical Co., Inc., 238th St. and Bronx Boulevard, New York 70, N. Y., has prepared a new brochure covering the features and operation of their improved Spectrofluorometer.

Among the improvements is a new attachment which now permits the measurement of transmission or absorption of the sample.

For a copy of the brochure contact the firm.

Transducer Converters

A new series of transducer converters permits the use of differential transformer-type transducers with 28 vdc and 115 vac power sources. Developed and manufactured by the new **Transducer Div., Sanborn Co.,** 175 Wyman St., Waltham 54, Mass., these converters produce DC signals proportional to the input, ready for monitoring or recording with a general-purpose DC amplifier. No carrier amplification is needed.



The two models (592-100 and 592-200) that accept 28 vdc contain a 5 kc oscillator, demodulator and filter. The first produces a DC signal that varies linearly with both displacement and input voltage. The second (592-200), with a voltage regulator, produces an output signal that is independent of input voltage variations. A third model, 592-300, accepts 115 V ±10%, 60-cps ac and produces an output signal that is independent of input voltage variations. All converter circuitry is contained in a 9¾″×1$\frac{9}{16}$″×2″ aluminum housing which has input, output and transducer connectors.

Complete information on these transducer converters and transducers for linear motion, pressure, etc., is available from the firm.

Micro-Miniature Ceramic Capacitors

Centralab, The Electronics Div., Globe-Union Inc., 900 E. Keefe Ave., Milwaukee 1, Wis., announces the availability, in production quantities, of a new series of microminiature, 3-v Ultra-Kap® ceramic capacitors. Described as being half the size of a matchhead, these units measure less than $\frac{3}{32}$″ in diameter. They

(Continued on page 136A)

SENSITIVE RESEARCH

POCKET POTENTIOMETER

New! .05% accurate Miniature DC Potentiometer!

Model PC *

PocketPot

Precision DC measuring potentiometer with self-contained galvanometer and battery operated standardization circuit. Direct "in line" readout. Additional voltage and current ranges can be obtained by using it in combination with the Model PC-S, a switch controlled, .05% accurate "plug in" unit of the same size.

Additional Specifications

ACCURACY: ± .05% of reading or ± .5 mv., whichever is greater.
RANGES: 0–5.100 v; when used with PC-S, 0–500 v. and 0–1 amp.
RESOLUTION: Continuous. 1 mv. divisions on slide wire.
SENSITIVITY: Infinite resistance at null. When used with PC-S, 2,000 Ω/v.
SIZE: 9″ x 4¼″ x 1¾″.
WEIGHT: 3 lbs.
PRICE: Model PC, $450.00; Model PC-S, $175.00.

*Copyright U.S.A.
Patent Applied for

SENSITIVE RESEARCH INSTRUMENT CORPORATION
NEW ROCHELLE, N.Y

NEWS New Products

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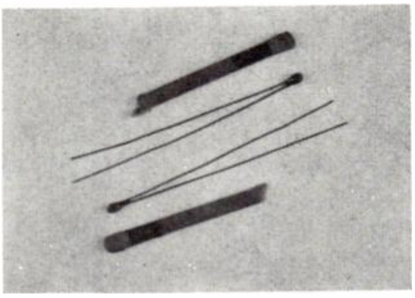
(Continued from page 135A)

are available in 0.005, 0.01 and 0.02 μf capacities. A 0.1 μf unit, approximately $\frac{7}{32}$″ in diameter is also available.



Power factor is a maximum of 10% at 1 kc, measured at 3 v rms maximum. Leakage resistance is in excess of 30K ohms, 1.4 v dc.

Specifically designed for low voltage semiconductor circuits, these high capacity units have excellent temperature stability from −55°C to +85°C, according to the manufacturer.

Cost ranges from $175.00 to $250.00 per thousand depending on rating and quantity. Delivery time is approximately four to six weeks. For more detailed information on these units write for bulletin EP 1350, available without charge from the firm.

Rectifier

A new compactron device for high-voltage rectifier service in television receivers has been developed by **Generla Electric Co.**, Receiving Tube Dept., Owensboro, Ky.

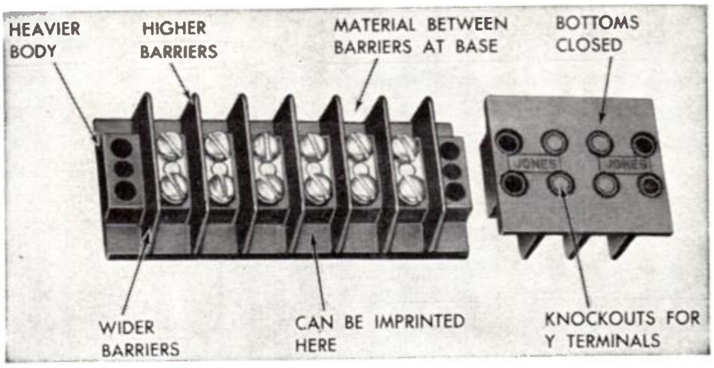
Conventional 1G3 Type 2AH2

Designated type 2AH2, the new 12-pin compactron rectifier is designed to supply power to the anode of the picture tube. It is immediately available in production quantities at a suggested user price of 71 cents.

Lower seated height and a 2.5 volt

(Continued on page 138A)

AT LAST—The IDEAL BARRIER TERMINAL STRIP



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Wider and higher barriers for increased creepage distances. Closed bottoms for complete insulation. Material between barriers at the base adds to the strength and maintains the same creepage distance between contact to contact and contact to ground. Can be imprinted here. No insulating or marker strip required. Three series—540, 541 and 542 having the same terminal spacing as our 140, 141 and 142 series.

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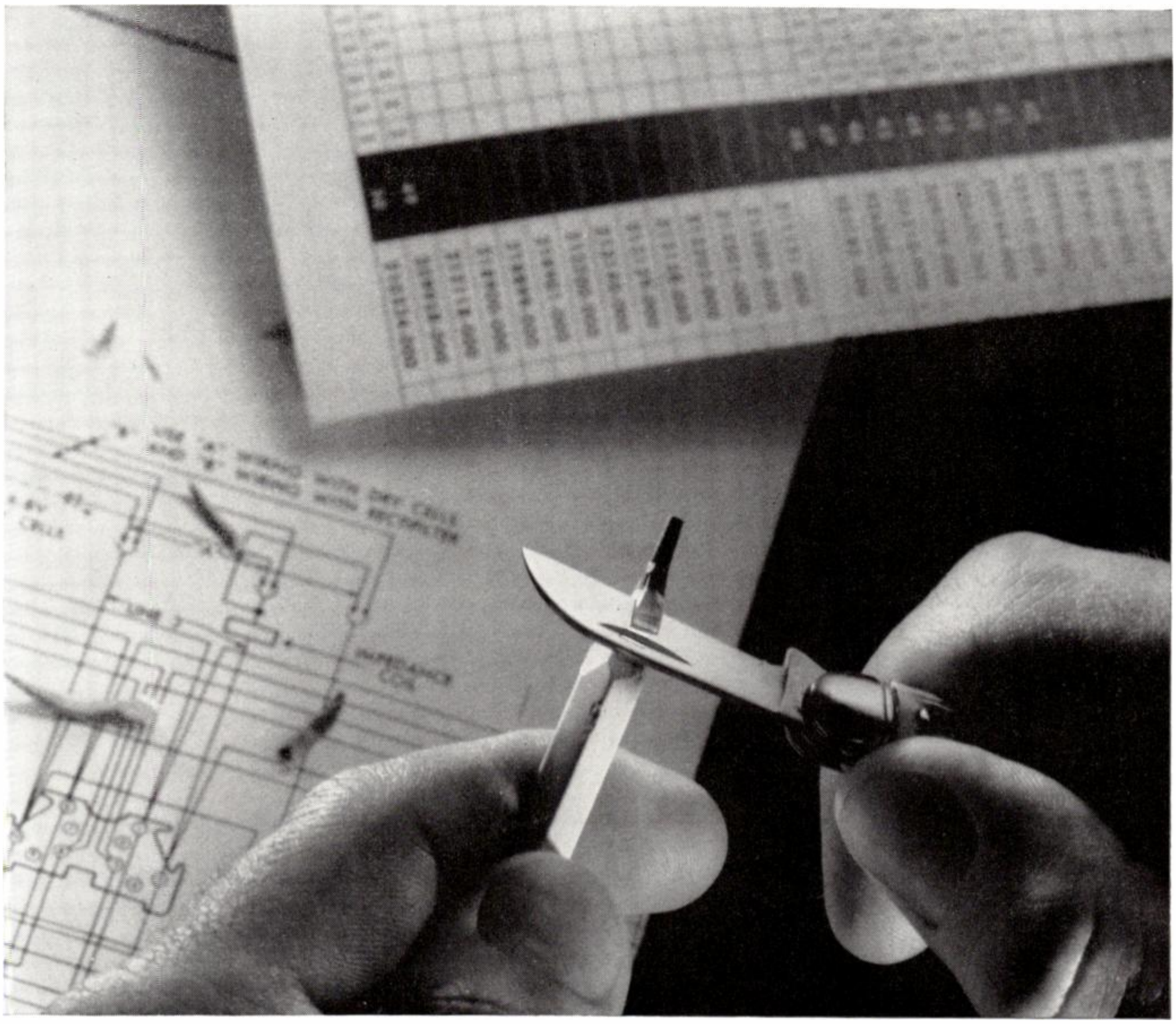
CINCH MANUFACTURING COMPANY
CHICAGO 24, ILLINOIS
DIVISION OF UNITED-CARR FASTENER CORP.



Sorry,
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Now Corning supplies total ΔR less than 3% with no derating

Start a design with known limits of resistance deviation as tight as that and you can specify other components with more certainty and more freedom.

Drop an amplifier stage? Use broader tolerance, cheaper tubes or transistors?

Model		Resistance (ohms)	Corning Design Tolerance
NF (Meets Mil-R-10509D)	60 65	100 to 100K 100 to 348K	3%
N (Meets Mil-R-10509D)	60 65 70	10 to 133K 10 to 499K 10 to 1 meg.	3%
C (Meets Mil-R-22684)	20 32 42S	51 to 150K 51 to 470K 10 to 1.3 meg.	5% (plus purchase tolerance of either 2% or 5%)

That's what our new *Corning Design Tolerances* are all about. They give you a percent deviation from nominal that includes the purchase tolerance, maximum ΔR due to TC, and maximum load-life drift. They're based on extended performance at full power and 70°C. ambient for over 30,000 hours.

We've assigned them to resistors that cover the 10 ohm—1 meg. range: the fusion-sealed NF, the precision N, and the low-cost, high-performance general purpose C.

A new folder, "Design Tolerances for Tin Oxide Resistors," gives you full information. Write for a copy to Corning Glass Works, 542 High St., Bradford, Pa. . . . and sharpen your pencil.

CORNING
Electronic Components

NEWS New Products

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(Continued from page 136A)

cathode-type heater are unusual features of the new compactron diode as compared with conventional-tube type counterparts, such as the 3A3, 1B3, 1K3 and 1G3.

The cathode-type heater used in the 2AH2 enables it to handle substantially higher voltages than other rectifiers with filament-type heaters. Maximum peak inverse plate voltage rating of the 2AH2 is 30,000 volts for total DC and peak; 24,000 volts in the DC component. Maximum steady-state peak plate current is 80 ma and maximum DC output current is 1.5 ma.

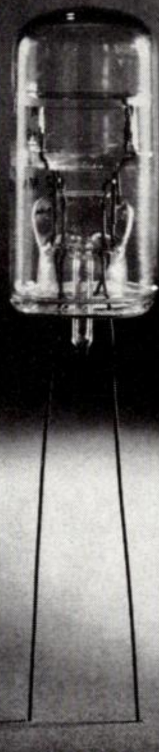
Current Indicator and Integrator

A combined measuring and integrating instrument that is widely applicable to the fields of chemistry, nuclear physics, and radiography has been announced by its manufacturer, **Elcor, Inc.**, 1225 W. Broad St., Falls Church, Va.

The Model A309B can be used in the nuclear physics field for measuring and integrating beam current in electron or positive ion accelerators. Chemical applications encompass a wide variety of important uses. For coulombmetric measurements, it will continuously indicate the progress and extent to which a reaction has progressed by providing a digital reading of the accumulated charge. With a calibrated ionization chamber the A309B will measure X-ray and gamma radiation intensity and measure the accumulated exposure or dossage over a wide range and for extended periods of time.

The A309B can also be used as an automatic control for processes, experiments, and monitoring. When a given amount of charge (easily and accurately predetermined by front panel controls) has been collected, the instrument will stop or continue integrating as determined by user, give a signal and/or sound an alarm, and shut off associated equipment being controlled by the A309B.

Full-scale input current sensitivity can be varied from 1 milliampere to 3 millimicrocoulombs per second. The integration period can extend from a fraction of a second to 280 hours, and can be terminated either upon reaching a preset charge or at a preset time by using an external



clock. Current of either polarity can be measured and integrated.

New ISOPLY circuitry is employed to obtain a drift of less than 0.01% per hour. A mercury-cell current source is provided internally for calibration purposes and for convenience in checking performance. Accuracy exceeds 1% of full-scale. Other features include the ability to operate with relatively low signal voltages or signal source impedances and provision for driving either a 1 ma or a 10 mv recorder.

Additional information and complete specifications are given in Elcor Bulletin No. 36-1261. The price of the Model A309B Current Indicator and Integrator is $1274.00 with delivery three (3) weeks after receipt of order.

IBM Promotes Marcy

H. Tyler Marcy has been promoted to vice president of the **General Products Div., International Business Machines Corp.,** 112 East Post Rd., White Plains, N. Y. The announcement was made by John W. Haanstra who was recently promoted from the post to president of the division.

In his new position, Mr. Marcy will be responsible for product development for the division——one of two divisions in the company handling product development and manufacturing of commercial data processing equipment. Development laboratories are located in Endicott, N. Y., San Jose, Calif., and Rochester, Minn. Marcy will have his office in White Plains.

Marcy was formerly manager of the IBM development laboratory in Poughkeepsie, a part of the Data Systems division. He joined IBM in February, 1951 as a technical specialist at the Airborne Computer Laboratory the company then operated in Vestal, N. Y.

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

Delay Lines

PCA Electronics, 16799 Schoenborn St., Sepulveda, Calif., introduces a series of PCDL lumped constant delay lines packaged in hermetically sealed cases and specifically designed for application on printed circuit cards.

Developed to meet specific height and volume requirements, the case height of new delay line is 0.400 inch maximum. Width: 1.400 inches. Length may be varied as required. Time delay to rise time ratios of 5:1 require a delay line length of 1.6 inch, for example. Ratios of 25:1 may be obtained with a length of 5.6 inch.

Up to 8 terminals are available on a standard case, but more may be added as

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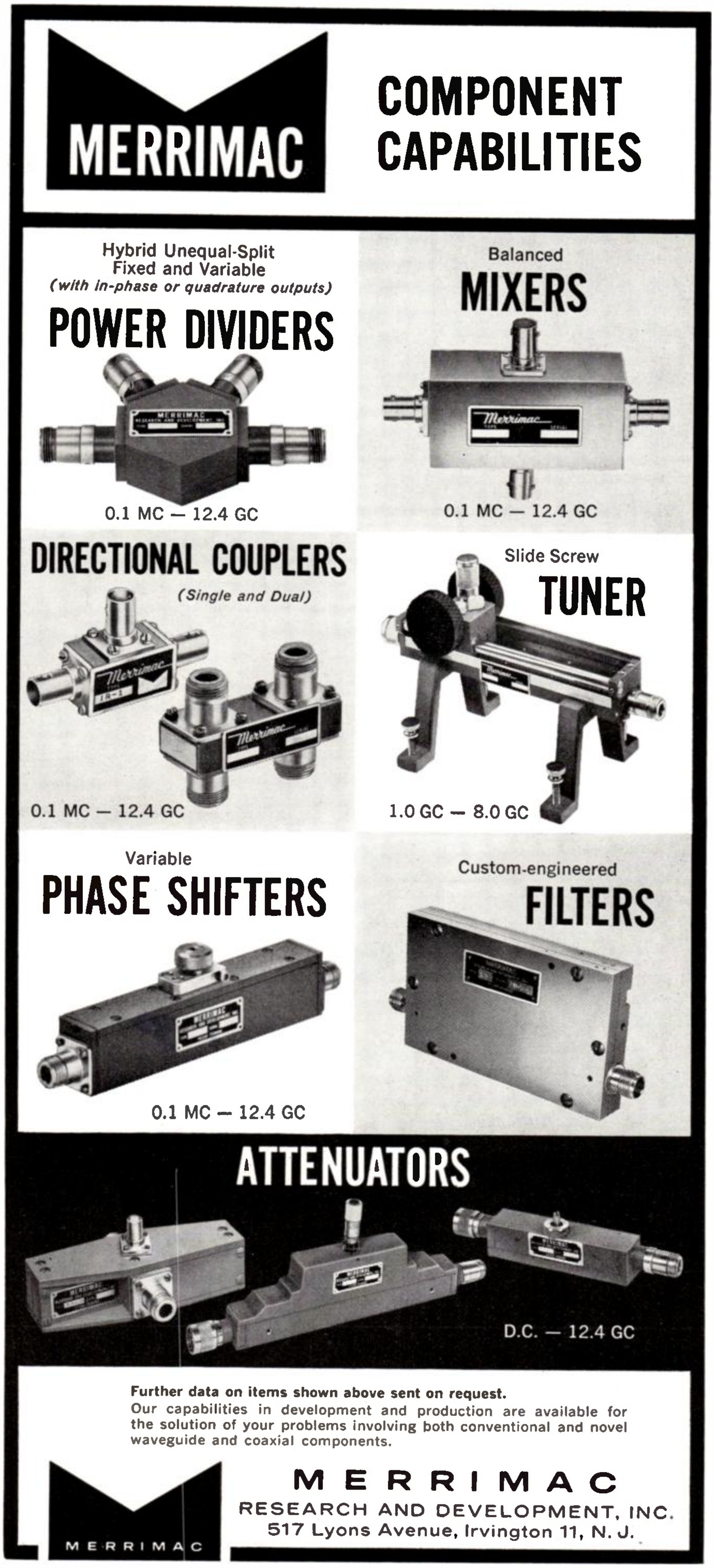

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(Continued from page 139A)

required. Terminals are 0.030 inch diameter pins on 0.200 inch, centers. A 6-32 insert on each end of line secures it to the board.

PCDL delay lines have a 500 VDC rating, low temperature coefficient, a maximum attenuation of 0.7 db/μsec for rise times less than 0.15 μsec., and 0.35 db/μsec for rise times greater than 0.15 μsec. Tolerances can be maintained through an environmental temperature range from −55° C to 125° C. Delivery date for PCDL delay lines is approximately 30 days. Price: from $20.00 to $150.00 per unit, depending on performance requirements.

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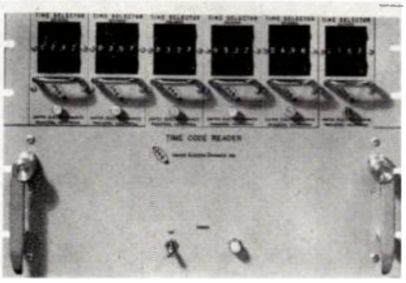
Antenna Laboratory: Smithtown, N. Y.

Time Code Reader and Selectors

The Time Code Reader developed by **United ElectroDynamics, Inc.,** 200 Allendale Rd., Pasadena, Calif., decodes serial IRIG time code for operation of the Time Selectors to provide precise automatic programming of instrumentation and control systems. A typical on-line application would be automatic control of a test sequence as follows:



1. Start tape recorders, cameras, oscillographs, etc.
2. Activate system calibration.
3. Initiate test function (engine run, shock tube firing, etc.).
4. Terminate test function (shut down engine, etc.).
5. Recalibrate.
6. Stop recorders, cameras, etc.

When used with played-back tape signals, the unit may be used for automatic editing and programming of data-processing sequences. Time Selectors may be pre-set to activate any given function with an accuracy of better than one millisecond. Resolution and range of times selected can be provided to meet specific requirements.

General Communications Appoints Whittier

Carl H. Whittier has been named to the new position of manager of Technical Development for **General Communication Co.,** 677 Beacon St., Boston, Mass., microwave engineering firm.

Mr. Whittier has been associated with the company since 1942, in advanced engineering and development work. In his new position, he will be responsible for the origination or initial investigation of technical developments applicable to the firm's expanded areas of activity, in microwave measurements, instrumentation and application.

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Full azimuth and elevation sweeps 360 degrees in azimuth. 210 degrees in elevation. Accurate to 1 mil. or better over system. Complete for full tracking response. Angle acceleration rate: AZ, 9 degrees per second squared EL, 4 degrees per second squared. Angle slewing rate: AZ 20 degrees per sec. EL. 10 degrees per sec. Can mount up to a 20 ft. dish. Angle tracking rate: 10 degrees per sec. Includes pedestal drives, selsyns, potentiometers, drive motors, control ampildynes. Excellent condition. Quantity in stock for immediate shipment. Ideal for missile & satellite tracking, antenna pattern ranges, radar system, radio astronomy, any project requiring accurate response in elevation and azimuth.
Complete description in McGraw-Hill Radiation Laboratory Series, Volume 1, page 284 and page 209, and Volume 26, page 233.

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Type CM 706A Freq. 3000 to 4000 mcs. CW. Output 200 Watts minimum. New, with full guarantee.

BROAD BAND BAL MIXER using short slot-hybrid. Pound type broad band dual balanced crystal holder. 1x5 wg size. $25. New.

WESTERN ELECTRIC TD-2 MICROWAVE LINK 4 KMC CRYSTAL OSC.

Basic crystal oscillator multiplying thru 2C37 cavities and 416B cavities with crystal stability output 3700-4200 MCS. New.

2C40 LIGHTHOUSE CAVITY

AN/APW-11A transmitter cavity for 2C40. Complete S band coverage at max. power. Temperature compensated. New $77.50 ea.

AN/AAS-6

Infrared Detection System, complete. New $5500.

60 MC I.F. STRIP

—95 DBM sensitivity. 130 DB gain. 1 MC & 5 Mc band widths. Mfg. Texas Instruments. $125 each w/tubes.

L BAND RF PKG.

20 KW peak 990 to 1040 MC. Pulse width .7 to 1.2 micro sec. Rep rate 180 to 420 pps. Input 115 vac. Incl. Receiver $1200.

MIT MODEL 9 PULSER 1 MEGAWATT-HARD TUBE

Output pulse power 25 KV at 40 amp. Max. duty ratio: 002. Uses 6C21 pulse tube. Pulse duration .25 to 2 microsec. Input 115 volts 60 cycle AC. Includes power supply in separate cabinet and driver. Fully guaranteed as new condition. Full Desc. MIT. Rad. Lab. Series "Pulse Generators."

2 MEGAWATT PULSER

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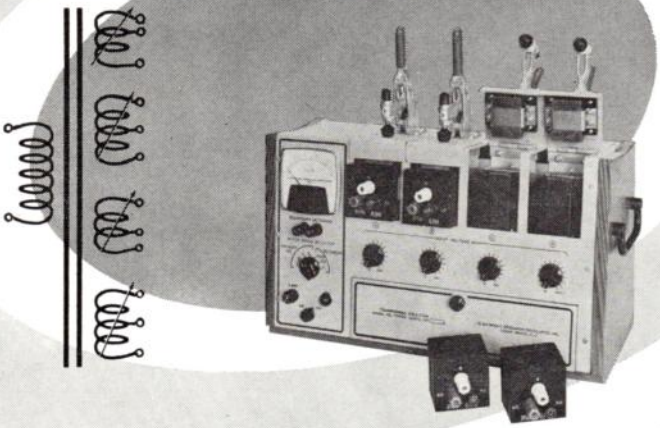

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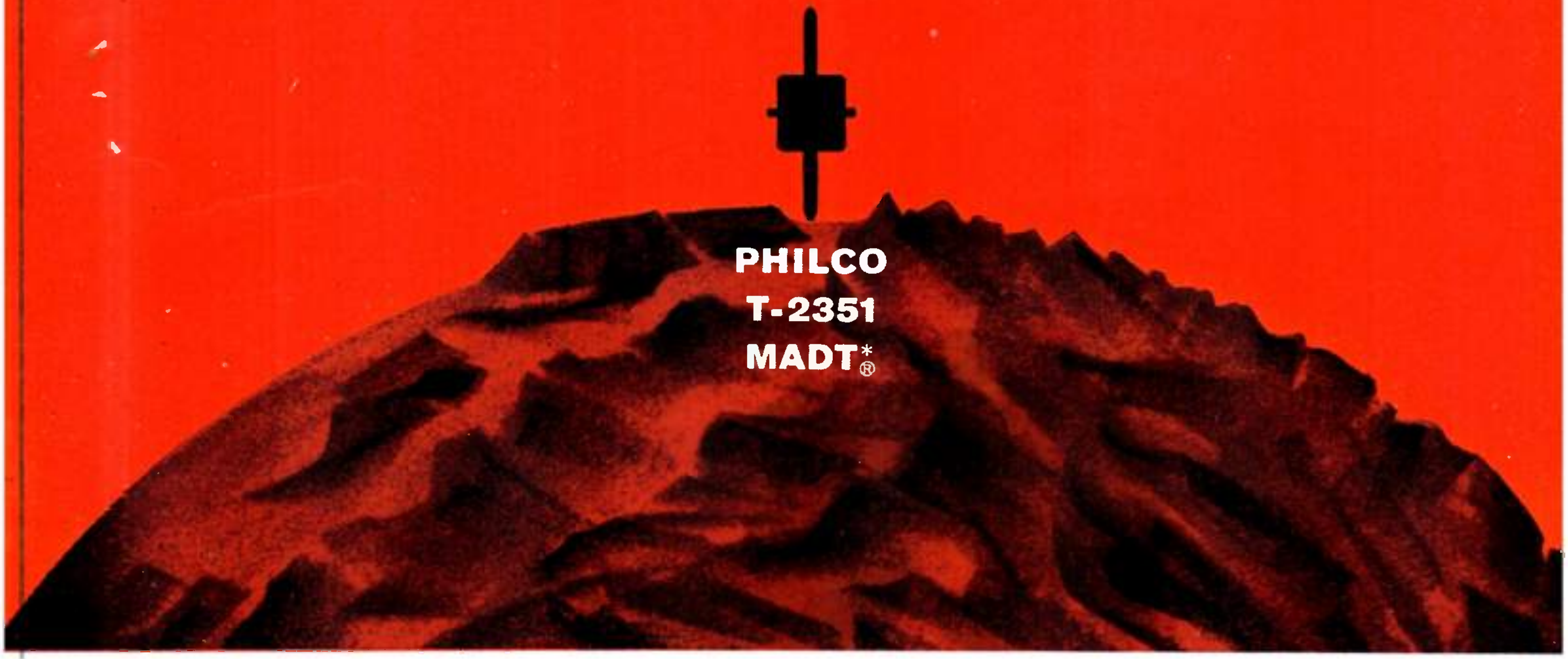

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