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Proceedings of the IRE

in this issue

10th

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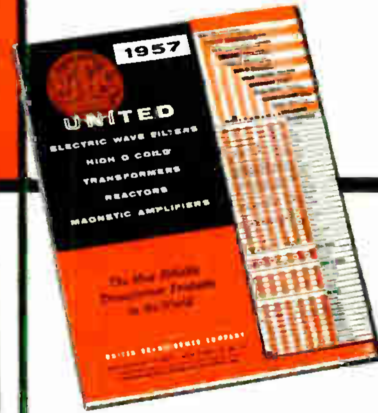

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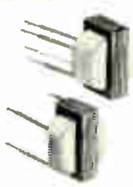

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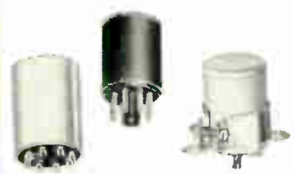

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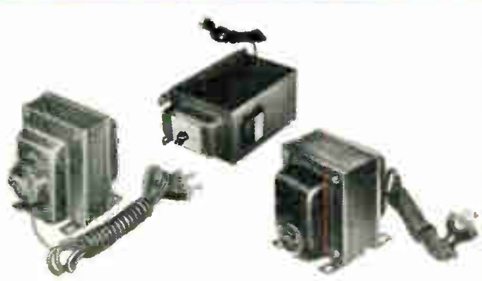

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VOLUME 46, NUMBER 5, PART I

May, 1958

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

contents

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

continued

COVER

The valuable service that the Joint Technical Advisory Committee, formed ten years ago this month by the IRE and the EIA, has rendered the FCC regarding utilization of radio frequencies is recounted in the article on page 823.

PROCEEDINGS OF THE IRE, published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N. Y. Manuscripts should be submitted in triplicate to the Editorial Department. Responsibility for contents of papers published rests upon the authors, and not the IRE or its members. All republication rights, including translations, are reserved by the IRE and granted only on request. Abstracting is permitted with mention of source.

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

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In the January 1958 issue of the Proceedings, Jack Greene, Section Head in our Department of Applied Electronics, discussed "Noise Factor and Noise Temperature." In this issue, he covers some related factors which result from living in a universe that is generating radio signals at power levels that are beyond our comprehension.

Antenna Noise Temperature

At low radio frequencies, it has long been recognized that the noise power received by an antenna from its surroundings is in general higher than that contributed by the receiver connected to it. As a result, the effective input noise level of an antenna-receiver system at low frequencies is primarily determined by antenna noise. At higher radio frequencies the antenna noise level decreases and consequently receiver noise becomes a dominant factor in determining system sensitivity. However, with the advent of extremely low-noise receivers useable at high frequencies (such as the maser and reactance amplifier), an exact knowledge of the antenna noise contribution becomes vital since it may determine overall system sensitivity.

The noise power received by an antenna from its surroundings is defined as the integral of the product of antenna gain and the noise power radiated by the surroundings. For a receiver of bandwidth, B, it is convenient to convert antenna noise power, P_A, to an antenna noise temperature, T_A, by the relation $T_A = P_A/KB$, where K is Boltzman's constant. One can also define an antenna noise temperature as the integral of the product of antenna gain and the noise temperature distributed in the antenna surroundings. The noise temperature distribution is in general composed of the following components: (a) noise due to galactic and extragalactic sources, (b) noise due to discrete "radio stars", (c) noise associated with atmospheric absorption, and (d) noise associated with terrestrial absorption. (Man-made interference is not included.) The noise temperatures due to sources (a) and (b) have been fairly well tabulated by radio astronomers, and in general decrease with frequency. The noise temperature due to (c) decreases with elevation angle above the horizon, is a function of weather conditions, and except for a few "windows" increases with frequency to an asymptotic value. The noise temperature due to (d) is a function of frequency and the condition of physical surroundings (such as the type of soil, calm or choppy seas, etc.) and in general increases with frequency to an asymptotic value equal to the physical temperature of the surroundings.

Other incidental contributions to antenna noise result from resistive losses in the antenna, feed, and transmission line coupling the feed to the receiver. These losses are especially serious in low noise systems since they simultaneously attenuate input signals and generate noise; thus they can degrade system sensitivity by more than the attenuation associated with the losses.

Computed antenna noise temperature for a typical 10 foot diameter parabolic reflector operating at 1000 Mc is shown as a function of antenna elevation angle in Figure 1. The data in this figure were computed assuming a) vertical polarization, b) the antenna is located on a seacoast and looks over the sea, c) the antenna is always pointed at the galactic center,* d) no intense "radio stars" are in the antenna pattern, e) resistive losses in the reflector, feed, and transmission line absorb 2% of the incident power, and f) that the feed produces a parabolic illumination taper. One further assumption requires some explanation. For a given illumination a theoretical antenna pattern can be predicted. However, in practical antennas, side lobes not predicted by this theory result from spillover, diffraction at the reflector edge and supporting structures, etc. Because these incidental minor lobe patterns vary so widely from antenna to antenna depending on the particular design, a constant gain of 0.2 is assumed for the whole of the incidental pattern. This gain figure is representative of typical parabolic antennas where no special effort has been made to minimize these effects.

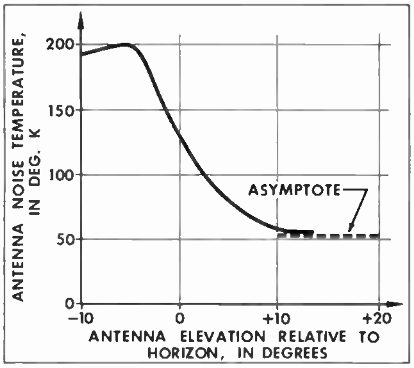

Figure 1 shows T_A is about 105 deg. K when the 3 db point of the main lobe is on the horizon, and decreases to an asymptotic value of about 55 deg. K as the elevation angle increases. At present, masers with an effective noise temperature of about 25 deg. K are available, and hence for the antenna described, the antenna noise temperature would be substantially higher than the maser noise temperature for all elevation angles. Because of this, efforts are now being made to minimize the incidental minor lobe pattern in designing antennas for low noise systems, since this not only minimizes the antenna noise temperature, but simultaneously increases antenna main lobe gain.

From the above discussion it is clear that when low noise systems are considered, one cannot simply assume that the receiver noise temperature determines system sensitivity, but must carefully evaluate the contributions of the antenna and its surroundings as well.

* This assumption is pessimistic since the galactic center has a higher noise temperature than the galactic pole, and an antenna does not generally always point toward the galactic center.

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

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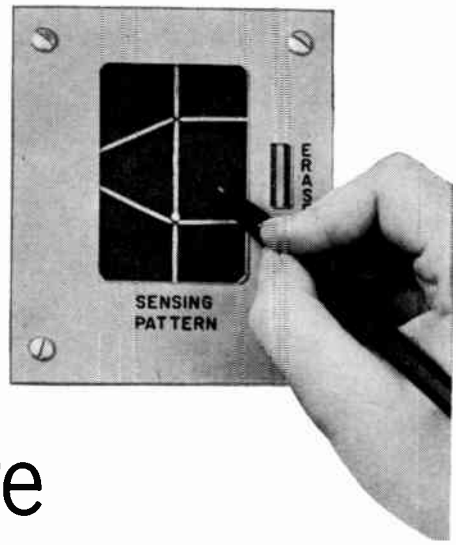

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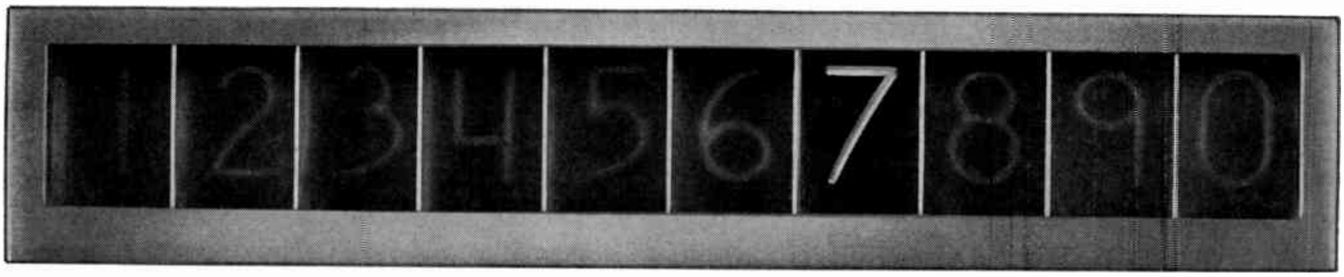
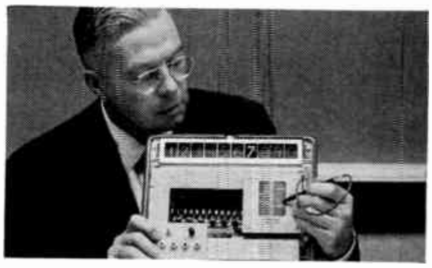


on new Bell Labs machine

A new device invented at Bell Laboratories "reads" a numeral while it is being written and instantly converts it into distinctive electric signals. The signals may be employed to make a numeral light up in a display panel, as above, or they may be sent to a computer or to a magnetic "memory" for storage.

The writing is done with a metal stylus on a specially prepared surface. Two dots, one above the other, are used as reference points. Seven sensitized lines extend radially from the dots. Transistorized logic circuits recognize numerals according to which lines are crossed.

The concept of a number-reader has interesting possibilities as a new means of communication from humans to machines. For example, in an adjunct to a telephone, it might provide inexpensive means of converting handwritten data into signals which machines can read. The signals could be transmitted through the regular telephone network to a teletypewriter or computer at a distant point. In this way, a salesman might quickly and easily furnish sales data to headquarters, or a merchant might order goods from a warehouse.



Tom Dimond, a B.S. in E.E. from the University of Iowa, demonstrates an experimental model of his number-reading invention. A similar device can also be made to read alphabetical characters. Small size and low power requirements result from transistor circuitry.

Modern communication involves many more fields of inquiry than the transmission and reception of sound. The experimental number-reader is but one example of Bell Telephone Laboratories work to improve communications service.

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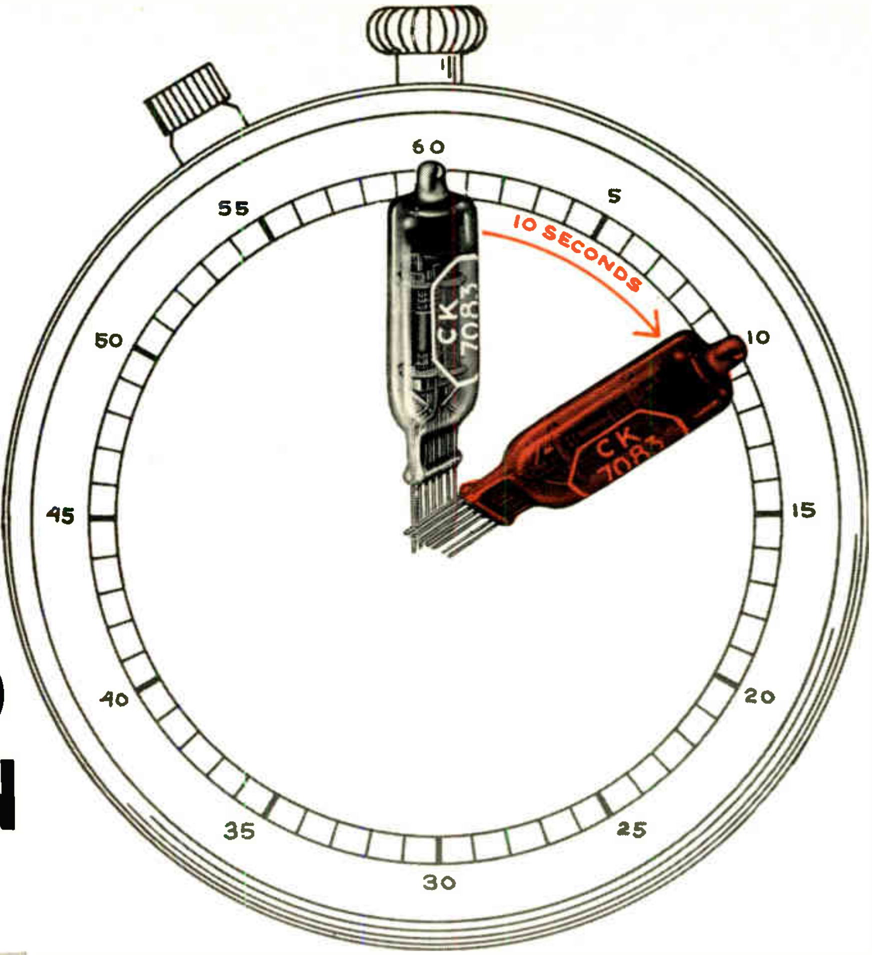

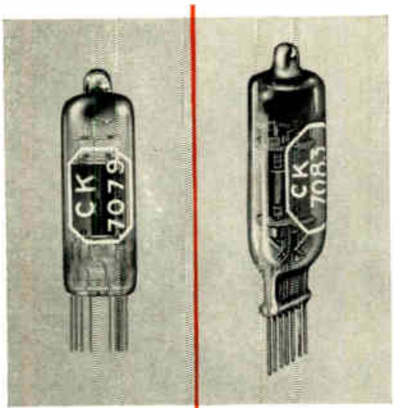

Actual Size

	CK7079 TWIN TRIODE	CK7083 AMPLIFIER PENTODE
Gm	5000*	5000
μ	20*	
Heater volts	6.3	6.3
Heater mA	300	200
Plate volts	100	120
Plate mA	8.5*	7.5
	Approximate prototype CK6111	Approximate prototype CK5702WA

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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

May 6-8, 1958

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

Exhibits: Mr. David F. Weinberg, Ramo-Wooldridge Corp., P. O. Box 45067, Airport Station, Los Angeles, Calif.

May 12-14, 1958

National Aeronautical & Navigational Electronics Conference, Dayton Biltmore Hotel, Dayton, Ohio.

Exhibits: Mr. John Kinnally, 203 Talbott Bldg., Dayton 2, Ohio.

June 4-6, 1958

Armed Forces Communications & Electronics Association Convention & Exhibit, Sheraton-Park Hotel, Washington, D.C.

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

June 5-6, 1958

Second National Symposium on Production Techniques, Hotel New Yorker, New York, N.Y.

Exhibits: Mr. Robert W. Swiggett, Photocircuits, Inc., Glen Cove, L.I., N.Y.

June 16-18, 1958

Second National Convention on Military Electronics, Sheraton-Park Hotel, Washington, DC.

Exhibits: Mr. L. David Whitelock, BuShips, Electronics Div., Dept. of Navy, Washington, D.C.

August 19-22, 1958

Western Electronic Show and Convention, Ambassador Hotel and Pan Pacific Auditorium, Los Angeles, Calif.

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

September 22-24, 1958

National Symposium on Telemetering, American Hotel, Miami Beach, Fla.

Exhibits: Mr. L. P. Clark, Tele-Dynamics, Inc., 5000 Parkside Ave., Philadelphia 31, Pa.

October 8-10, 1958

IRE Canadian Convention, Exhibition Park, Toronto, Canada.

Exhibits: Mr. Grant Smedmor, IRE Canadian Convention, 1819 Yonge St., Toronto, Canada

October 13-15, 1958

National Electronics Conference, Hotel Sherman, Chicago, Ill.

Exhibits: Mr. J. S. Powers, National Electronics Conference, 84 East Randolph St., Chicago 1, Ill.

October 20-22, 1958

Fourth Annual Symposium on Aeronautical Communications, Hotel Utica, N.Y.

Exhibits: Mr. Robert E. Gaffney, 50 Cambridge Rd., Whitesboro, N.Y.

(Continued on page 10A)

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Instrument & Cat. No.	Frequency Range (mc)	Noise Figure Range (db)	Output Impedance (ohms)	Price*
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Mega-Node 175-A	50-500	0-19	unbalanced—50	$325.00
Mega-Node 403-A	3-500	0-19	unbalanced—50	$790.00
Mega-Node-Sr. 250-B	10-3000	0-20	unbalanced as specified	$1395.00
Rada-Node 600-A	5-400	0-23.8 depending on impedance	unbal. nom. 50	$1865.00
	10-3000	0-20	waveguide	†
	1120-26,500	15.28 or 15.8		
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RG-69/U	UG-417/U	1200-1400	311-A	310-A	$395.00
RG-104/U	UG-435/U	1700-2600	**	870-A	$495.00
RG-112/U	UG-553/U	2200-3300	**	880-A	$495.00
RG-48/U	UG-214/U	2600-3900	261-A	260-A	$175.00††
RG-49/U	UG-149/U	3900-5850	271-A	270-A	$175.00††
RG-50/U	UG-344/U	5850-8200	281-A	280-A	$175.00††
RG-51/U	UG-51/U	7050-10,000	291-A	290-A	$175.00††
RG-52/U	UG-39/U	8200-12,400	301-A	300-A	$175.00††
RG-91/U	UG-419/U	12,400-18,000	521-A	**	$250.00
RG-53/U	UG-425/U	18,000-26,500	531-A	**	$250.00

†† Any three plus power supply: $595.00. Any in excess of three: $167.00 ea.
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Meetings with Exhibits

(Continued from page 8A)

October 27-29, 1958

East Coast Aeronautical and Navigational Electronics Conference, Lord Baltimore Hotel & 7th Regiment Armory, Baltimore, Md.

Exhibits: Mr. R. L. Pigeon, Westinghouse Electric Corp., Air Arm Div., P.O. Box 746, Baltimore, Md.

November 19-20, 1958

Northeast Electronics, Research and Engineering Meeting (NEREM), Mechanics Building, Boston, Mass.

Exhibits: Mr. Howard H. Dawes, General Radio Co., 275 Massachusetts Ave., Cambridge 38, Mass.

December 3-5, 1958

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

Exhibits: Mr. L. D. Whitelock, BuShips, Room 1025, Main Navy Building, Washington 25, D.C.

December 3-5, 1958

Second National Symposium on Global Communications, St. Petersburg, Fla.

Exhibits: Mr. Robert L. Lazarchik, Sperry Rand Corp., P. O. Box 1828, Clearwater, Fla.

December 4-5, 1958

PGVC Annual Meeting, Hotel Sherman, Chicago, Ill.

Exhibits: Mr. Frederick L. Hilton, 4501 Augusta Blvd., Chicago, Ill.

December 9-11, 1958

Mid-America Electronics Convention, Municipal Auditorium, Kansas City, Mo.

Exhibits: Mr. Leo Schlesselman, Bendix Aviation Corp., Box 1159, Kansas City 41, Mo.

March 23-26, 1959

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

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

April 16-18, 1959

South Western IRE Regional Conference & Electronics Show, Dallas, Tex.

Exhibits: Mr. Durward Tucker, WRR, State Fair Grounds, Dallas, Tex.

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

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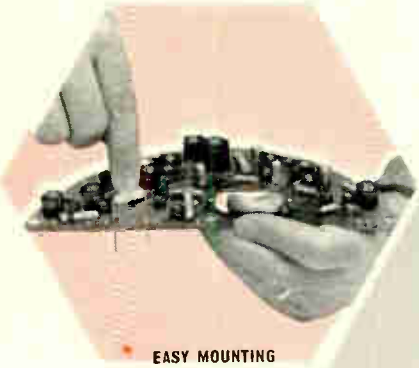

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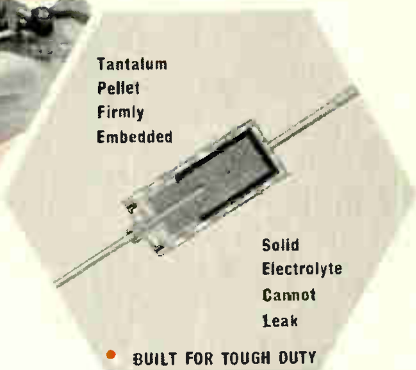

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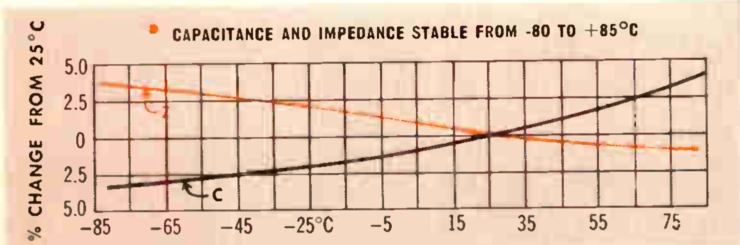

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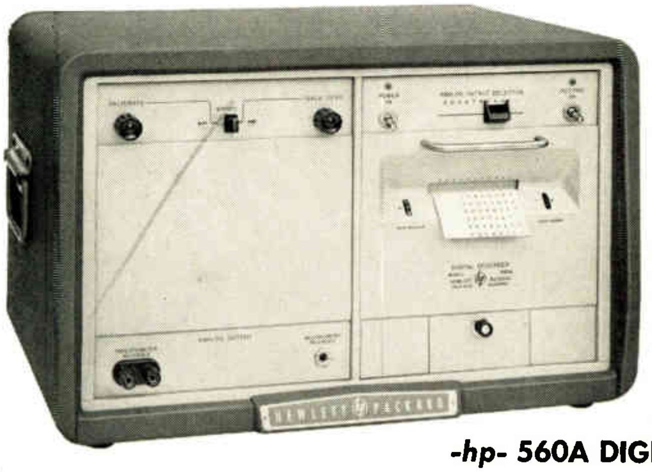

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-hp- 560A DIGITAL RECORDER

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-hp- 560A/AR Digital Recorder

Accuracy:	Identical to counter used.	
Printing Rate:	5 lines/sec maximum.	
Digit Capacity:	11 per line.	
Driving Source:	Parallel entry staircase voltages. Descending from 135 to 55 v, 0 to 9.	
Analog Output:	Proportional to any 3 digits selected. Maximum amplitude 1 ma or 100 mv.	
Print Command Signal:	1 μsec minimum, pos. or neg. 15 v per pulse.	
Paper:	3" roll or folded.	
Line Spacing:	Single or double, adjustable.	
Price: -hp- 560A/AR:	(11 digit, cabinet model)	$1,390.00
	(11 digit, rack mount)	1,375.00
	(6 digit, cabinet model)	1,265.00
	(6 digit, rack mount)	1,250.00

Data subject to change without notice. Prices f.o.b. factory.

464B

New *-hp-* 560A Digital Recorder works direct with all *-hp-* counters and most other precision electronic counters; no intermediate equipment is needed. It provides a complete record of all types of test data, plus, through an analog output, a convenient graphic record of very small data variations.

The analog output for driving a strip chart recorder is a voltage or current proportional to the number represented by any three consecutive digits of recorded data. The 560A permits expanded scale strip chart recording and the strip chart can never be driven off scale since range variation for the 3-digit scale is 0 to 999. Wider variation merely causes a repetition of the 0 to 999 sequence.

Model 560A is a complete, self-contained electronic instrument normally controlled by staircase voltages and a print command pulse from an electronic counter. It may, however, be controlled by other electronic or electro-mechanical devices. Printing speed is five, 11-digit lines per second; secondary or coding data may be entered simultaneously with primary data.

Maximum print capacity of the recorder is five, 11-digit lines per second but instruments can be supplied with any lesser number of digits desired.

hp offers the most complete line

lines per second!

Record Counters

-hp- Electronic Counters

-hp- 524B Electronic Counter. Measures frequency 10 cps to 220 MC, time interval 1 μsec to 100 days, period 0 cps to 10 KC. Basic -*hp*- 524B Counter, 10 cps to 10 MC, $2,150.00. Plug-in Frequency Converters, Video Amplifiers, Time Interval Unit and Period Multiplier, $150.00 to $250.00.

-hp- 521A/C Industrial Counter. Accurate, low cost industrial instrument. Measures frequency, speed, random events, rpm, etc. Direct readings 1 cps to 120 KC. Time of count 0.1 to 1.0 seconds, variable display time. -*hp*- 521A has 4 place registration; -*hp*- 521C has 5 place registration, crystal controlled time base. -*hp*- 521C, $650.00. -*hp*- 521A, $475.00.

-hp- 522B Electronic Counter. Compact, versatile instrument for frequency, period *or* time measurements. Measures frequency 10 cps to 120 KC, time interval 10 μsec to 10^5 sec. Reads direct in cps, KC, seconds, milliseconds. $915.00.

-hp- 523B Electronic Counter. All-purpose counter measures frequency 10 cps to 1.1 MC, time interval 3 μsec to 27.8 hours, period 0.00001 cps to 10 KC. Stability 2/1,000,000 per week. Reads direct in sec, msec, μsec or KC; automatic decimal. Display variable 0.1 sec to 5 sec or indefinite. $1,215.00.

See your -hp- representative or write direct for details

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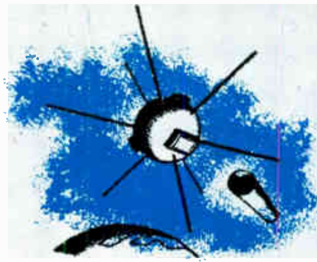


Satellite performance...

NEW -hp- RECORDER PROVIDES VITAL DOPPLER HEIGHT, RANGE DATA!

Two -*hp*- 560A Digital Recorders and two -*hp*- 523B Frequency Counters, connected to radio receivers, provided important orbital data on Soviet satellites. The installation in Figure 1 (photo courtesy Stanford Research Institute) shows the equipment arrangement which produced the tape showing frequency shift in cps (Figure 2) and strip chart recordings (Figure 3). Calculation based on Doppler shift is an efficient method of determining satellite range, height and other orbital information. Simultaneous Doppler records from differing frequencies provide propagation data. (See -*hp*- Journal Vol. 9, No. 3-4, for more details.)

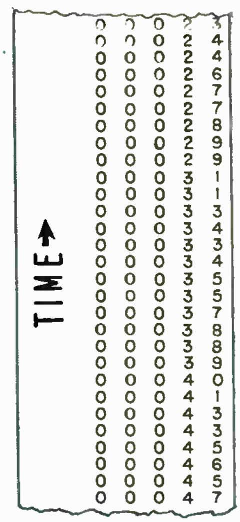


Figure 1. -*hp*- digital recorders, counters and dual trace oscilloscope assembled for satellite tracking.

Figure 2. Printed tape from -*hp*- 560A similar to that used during satellite measurements. Note Doppler frequency shift in cps.

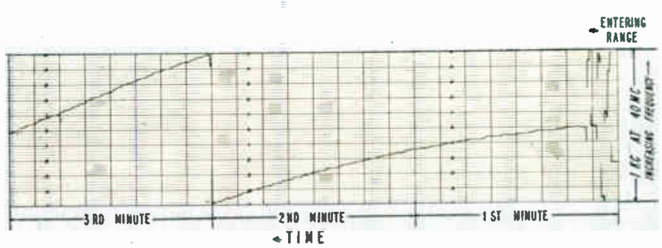


Figure 3. Strip-chart recording of 40 MC transmission from Soviet satellite. Note Doppler shift during 3-minute period when satellite transmitter was continuously keyed.

# of precision electronic counters

Recipients of the 1958 IRE awards are (*above left to right*): E. L. Ginzton, winner of the Liebmann Memorial Prize; R. L. Kyhl and H. F. Webster, co-winners of the Baker Award; C. P. Ginsburg, winner of the Zworykin Television Prize Award; and E. W. Allen, Jr., winner of the Diamond Memorial Award. *Below*—The 1958 IRE President, D. G. Fink (*left*), congratulates A. W. Hall, upon his receipt of the IRE's Medal of Honor for his inventions in the field of electron tubes.

1958 IRE NATIONAL CONVENTION

Record 54,500

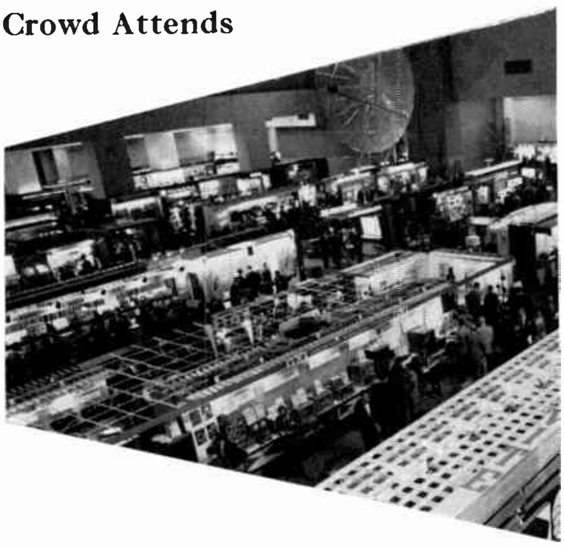
Above, left—Retiring IRE president J. T. Henderson (*right*) turns over the gavel to incoming IRE president D. G. Fink at the annual meeting. Above, right—An overflow crowd listens to the Tuesday evening session, a highlight of the convention. The session featured a panel on "Electronics in Space" which consisted of the following (*back row, left to right*): F. L. Whipple, Smithsonian Astrophysical Observatory; Maj. D. G. Simons, Holloman Air Force Base; Seville Chapman, Cornell Aeronautical Lab.; L. V. Berkner, Associated Universities, Inc.; C. S. Draper, M.I.T.; (*front row, left to right*), H. Preston-Thomas Canadian National Research Council; D. G. Fink, 1958 IRE President; Lee DuBridge, President, Calif. Inst. of Technology; and J. B. Weisner, M.I.T. Below, left—G. W. Bailey (*left*), convention chairman and IRE Executive Secretary, chats with C. E. Granqvist of Sweden, 1958 IRE Vice-President. Below, center—D. G. Fink (*left*), 1958 IRE President, greets R. C. Sprague, Chairman of the Board of Sprague Electric Co. and the Federal Reserve Bank of Boston, who was the banquet guest speaker. He spoke on "The Federal Reserve and the Electronics Industry." Below, right—G. L. Haller (*left*) banquet toastmaster, exchanges anecdotes with P. E. Haggerty, spokesman for the 75 new IRE Fellows, who were honored at the annual banquet.

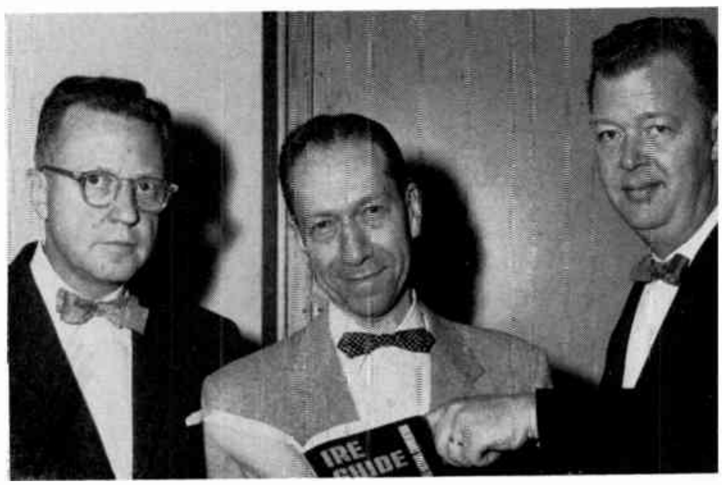
Above, left—Maurice Levy of the Canadian Post Office Department headed a five man team which described during a session the Canadian automation system o postal operations. Above, right—One session attendee was Howard Coonen, Directin Postmaster of New York City, shown here as he addressed an IRE press conference

Left—Ernst Weber, President of Polytechnic Institute of Brooklyn and IRE Direc tor, was the featured speaker at the IRE annual meeting on "The Broad Spectrum."

AND RADIO ENGINEERING SHOW

Crowd Attends

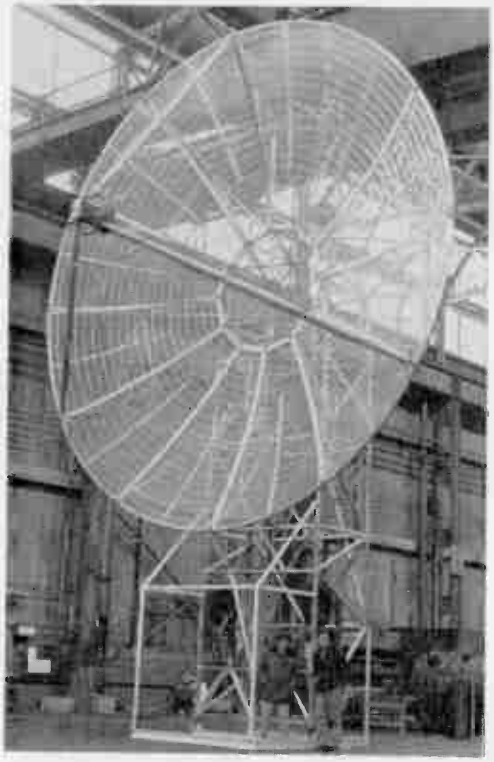




Above—E. W. Herold (*left*), of C. Stellarator Associates, and W. E. Tolles (*right*), Ar borne Instruments Labs., brief the press on the sessions devoted to controlled thermon clear power and medical electronics at a meeting moderated by Lewis Winner, of Brya Davis Publishing Co., and convention vice-chairman of publicity.

Above—Among the many interesting papers were those given by (*left to right*) E. A. Sack of Westinghouse on electroluminescent displays, Gordon Teal of Texas Instruments on engineering education, and Seville Chapman of Cornell Aeronautical Lab. on space flight propulsion. Right—P. C. Goldmark, President of CBS Laboratories, demonstrated for the first time a stereophonic disk that can also be played on standard long-playing phonographs for monaural listening.

Exhibits at the Show



Above—A mobile antenna for use at frequencies between 200 and 4000 mc, has a 28-foot diameter reflector which is sectionalized for storage within its mounting tower. Retractable wheels in the tower can be used for towing.

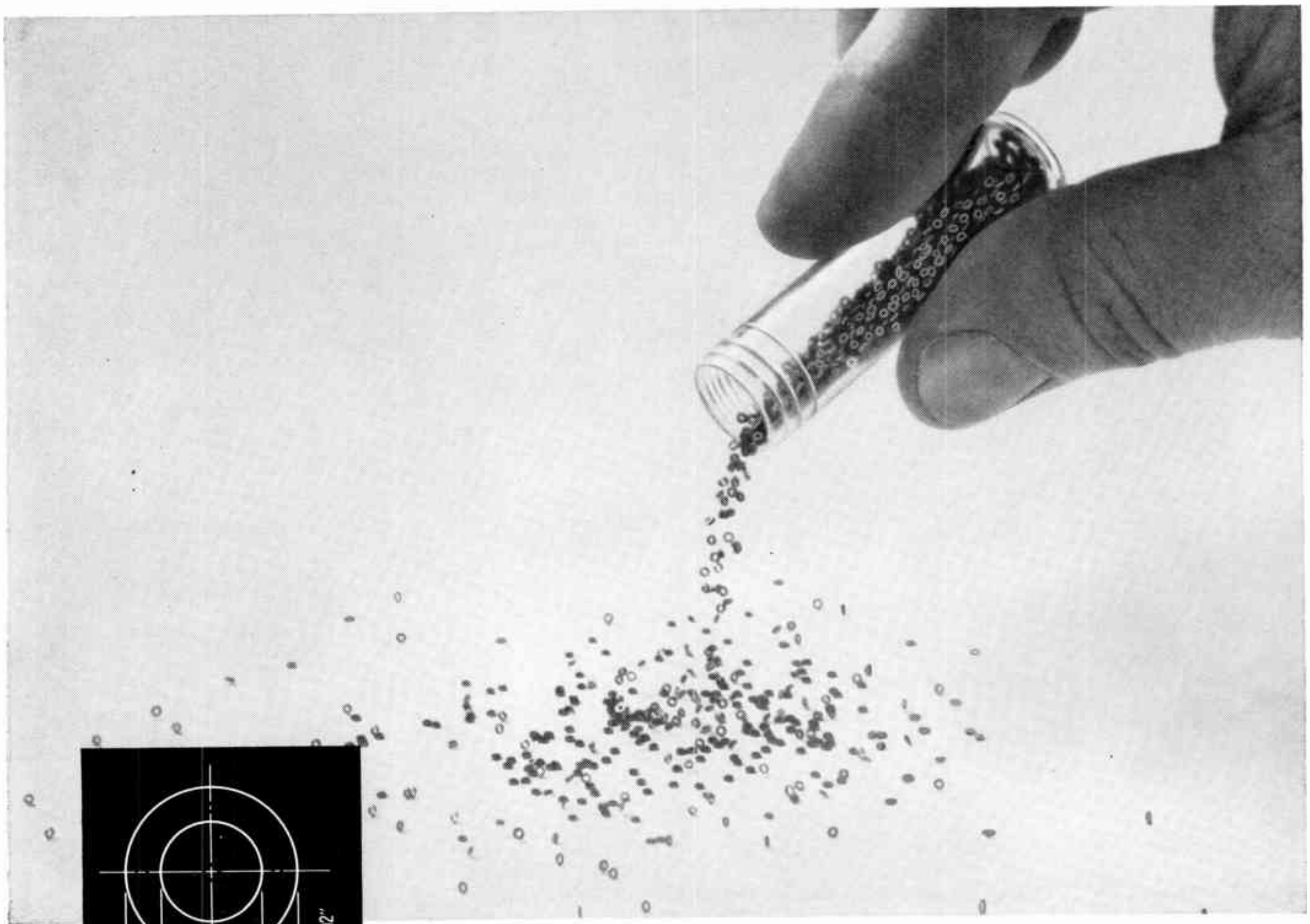
Above—The operation of MOBIDIC, a mobile, completely transistorized, general-purpose digital computer on wheels, is explained to these two members of the United States Army Signal Corps.

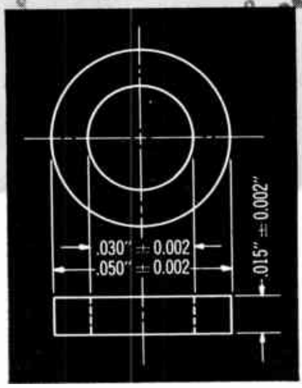
Above—The electronic spectroanalyzer determines the identity of chemicals by the quantity of infrared waves which are absorbed. Major units are a spectrophotometer, a recording device, a "library," and a high speed electronic computer. It is useful in medical electronics and nuclear radiation analysis.

Left—A submarine homing torpedo has been developed for the U. S. Navy.

Below—The Administrative Committee of the Professional Group on Broadcast Transmission Systems held their annual breakfast meeting just prior to the IRE National Convention. Seated (*left to right*) are: R. N. Harmon, T. E. Howard, P. B Laeser, C. H. Owen, chairman, and Carl Smith. Standing (*left to right*) are: Dana Pratt, Lewis Winner, *ex-officio*, R. J. Rockwell, George Town, O. W. B. Reed, Jr., *ex-officio*, W. L. Hughes, and G. W. Hagerty, secretary-treasurer.

NEW! From the Laboratories of General Ceramics





50 MIL O.D. Memory Cores for Transistorized High Speed Memories

These new 50 mil O.D. cores are now available in General Ceramics S-4, the material that has proven so successful in such vitally important systems as the SAGE computer. Switching time is less than one microsecond with 550 ma full drive. At recommended operating conditions, the "ONE" output voltage is greater than 60 millivolts; the "ZERO" output voltage is less than 6 millivolts. Cores are provided in two quality levels, to .015 AQL and to 6.5 AQL. Dimensions are .050" O.D., .030" I.D. and .015" in height, all with tolerances of ± .002". General Ceramics has designed and built special equipment for core testing to insure that each unit meets established electrical properties. 50 mil O.D. cores are supplied in production quantities in two quality levels. Parts are shipped according to MIL Specification 105A to 0.015 AQL or 6.50 AQL. For complete information on this core write General Ceramics Corporation, Keasbey, New Jersey, for Bulletin 326; address Dept. P.

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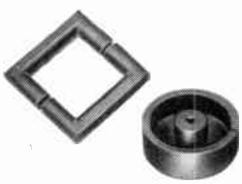
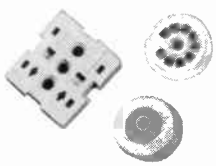
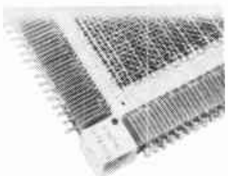


FERRAMIC CORES



PRECISION STEATITES



MAGNETIC MEMORY PLANES

"ADVAC" HIGH TEMPERATURE SEALS

SOLDERSEAL TERMINALS

IRE News and Radio Notes

Wayne University Joins NEC

The 1958 National Electronics Conference, the 14th Annual Forum on Electronic Research, Development, and Application, will be held at Chicago, Ill., at the Hotel Sherman, on October 13–15.

Wayne State University of Detroit, Michigan has been added as a participant, and is the ninth institution of higher learning to join the ranks of the NEC.

Sponsors of the conference are the IRE, American Institute of Electrical Engineers, Illinois Institute of Technology, and Northwestern and Illinois Universities.

Participants in the conference, in addition to Wayne, are Notre Dame, Purdue, Michigan State, Michigan and Wisconsin universities, Electronic-Industries Association, and Society of Motion Picture and Television Engineers.

Proceedings Now Available

Proceedings for the Tenth Southwestern IRE Conference are available now. They cost $3.00 for registrants, and $3.50 for separate orders. They are reproduced in the same format as the *IRE National Conference Record* and the *WESCON Record*. The issue will be copyrighted to protect the authors. Copies of the proceedings may be ordered from the San Antonio Section magazine, *The Analog*, Box 7948, Univ. Station, Austin 12, Tex., or from the registration chairman, Tenth SWIRECO, Box 55, San Antonio 6, Tex. The proceedings will carry the full text as prepared by the author of every paper given in the technical sessions, the winning paper of the Region 6 Student Paper Contest, and an abstract of all student papers entered in the Student Paper Contest.

MISCELLANEOUS IRE PUBLICATIONS

The following issues of miscellaneous publications are available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, New York, at the prices listed below:

Publications	Price per Copy
Component Symposia	
Proceedings of the 1954 Electronic Components Symposium, May 4–6, 1954, Washington, D. C.	$4.50
Proceedings of the 1957 Electronic Components Symposium, May 1–3, 1957, Chicago, Ill.	5.00
Electronic Computer Conferences	
Proceedings of the Joint AIEE-IRE-ACM Western Computer Conference, February 11–12, 1954, Los Angeles, Calif.	3.00
Proceedings of the Joint AIEE-IRE-ACM Western Computer Conference, March 1–3, 1955, Los Angeles, Calif.	3.00
Proceedings of the Joint AIEE-IRE-ACM Western Computer Conference, February 7–9, 1956, San Francisco, Calif.	3.00
Proceedings of the Joint AIEE-IRE-ACM Western Computer Conference, February 26–28, 1957, Los Angeles, Calif.	4.00
Proceedings of the Joint AIEE-IRE-ACM Eastern Computer Conference, December 10–12, 1951, Philadelphia, Pa.	3.50
Proceedings of the Joint AIEE-IRE-ACM Eastern Computer Conference, December 8–10, 1954, Philadelphia, Pa.	3.00
Proceedings of the Joint AIEE-IRE-ACM Eastern Computer Conference, November 7–9, 1955, Boston, Mass.	3.00
Proceedings of the Joint AIEE-IRE-ACM Eastern Computer Conference, December 10–12, 1956, New York, N. Y.	3.00
Proceedings of the Joint AIEE-IRE-ACM Eastern Computer Conference, December 9–13, 1957, Washington, D. C.	3.00
Magnetic Amplifiers Conference	
Proceedings of the Conference on Magnetic Amplifiers, April 5–6, 1956, Syracuse, N. Y.	4.00
Reliability and Quality Control in Electronics Symposia	
Proceedings of the National Symposium on Quality Control and Reliability in Electronics, November 12–13, 1954, New York, N. Y.	5.00
Proceedings of the Second National Symposium on Quality Control and Reliability in Electronics, January 9–10, 1956, Washington, D. C.	5.00*
Proceedings of the Third National Symposium on Reliability and Quality Control in Electronics, January 14–16, 1957, Washington, D. C.	5.00
Proceedings of the Fourth National Symposium on Reliability and Quality Control in Electronics, January 6–8, 1958, Washington, D. C.	5.00
Telemetering Conference	
Proceedings of the 1953 National Telemetering Conference, May 20–22, 1953, Chicago, Ill.	2.00

* IRE member rate—$3.00.

Calendar of Coming Events and Authors' Deadlines* 1958

PGMTT Symp., Stanford Univ., Stanford, Calif., May 5–7

Western Joint Computer Conf., Ambassador Hotel, Los Angeles, Calif., May 6–8

Nat'l Aero. Elec. Conf., Dayton Biltmore Hotel, Dayton, Ohio, May 12–14

Spring Assembly of Radio Tech. Comm. for Marine Services, Ben Franklin Hotel, Phil., Pa., May 13–15

IEE Convention on Microwave Values, Savoy Place, London, England, May 10–23

PGPT Symp., Hotel New Yorker, New York City, June 4–6

Int'l Automation Exposition & Congress, Coliseum, New York City, June 9–13

Nat'l Soc. of Prof. Engrs. Annual Mtg., Chase-Park Plaza Hotel, St. Louis, Mo., June 11–14

PGMIL Convention, Sheraton-Park Hotel, Wash., D. C., June 16–18

Nat'l Summer Mtg., Inst. of Aero. Sciences, Ambassador Hotel, Los Angeles, Calif., July 8–11

Spec. Tech. Conf. on Nonlinear Mag. and Mag. Amplifiers, Hotel Statler, Los Angeles, Calif., Aug. 6–8

Elec. Radio & Standards Conf., Univ. of Colo. and NBS, Boulder, Colo., Aug. 13–15 (DL*: Apr. 15)

Int'l Conf. on Semiconductors, Univ. of Rochester, Rochester, N. Y., Aug. 18–22

WESCON, Ambassador Hotel and Pan-Pacific Audit., Los Angeles, Calif., Aug. 19–22 (DL*: May 1, R. C. Hansen, Microwave Lab., Hughes Aircraft Co., Culver City, Calif.)

Int'l Conf. for Analog Computations, Strasbourg, France, Sept. 1–7

Int'l Cybernetics Conv., Namur, Belgium, Sept. 3–10 (DL*: June 1, Secretariat, 13, rue Basse-Marcelle, Namur, Belgium)

Int'l Nuclear Elec. Conf. UNESCO House, Paris, France, Sept. 16–20

Nat'l Symp. on Telemetering, Americana Hotel, Miami Beach, Fla., Sept. 22–24

Indus. Elec. Conf., Rackham Mem. Audit., Detroit, Mich., Sept. 24–25

PGEWS Symp., New York City, Oct. 2–3

IRE Canadian Convention, Exhibition Park, Toronto, Can., Oct. 8–10

Nat'l Electronics Conf., Hotel Sherman, Chicago, Ill., Oct. 13–15 (DL*: May 1, L. W. Von Tersch, Elec. Eng. Dept., Mich. State Univ., E. Lansing, Mich.)

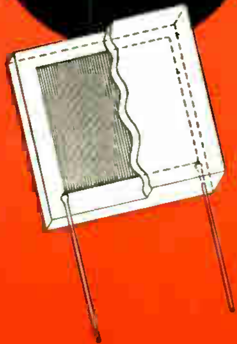
PGCS Symp. on Aero. Communications, Hotel Utica, Utica, N. Y., Oct. 20–22

Nat'l Simulation Conf., Dallas, Tex., Oct. 23–25 (DL*: June 25, D. J. Simmons, Rte. 8, Box 447, Ft. Worth, Tex.)

* DL = Deadline for submitting abstracts.

(Continued on page 20A)

NEW CTS



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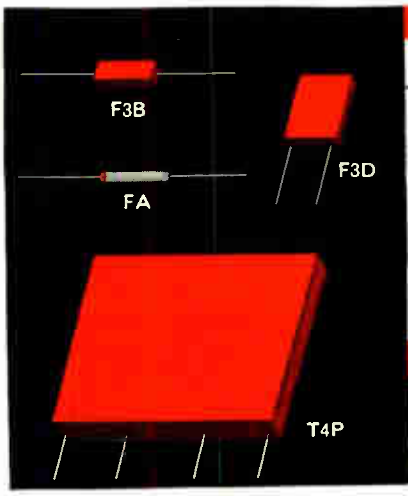

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0.25	3/4" x 3/16" x 1/2"	0.5-100,000	F3D
0.25	1" x 3/16" x 1/4"	0.5-100,000	H3B
0.5	1" x 3/16" x 3/8"	0.1-100,000	H3C
0.5	1" x 3/16" x 1/2"	0.1-200,000	H3D
0.75	1" x 3/16" x 3/4"	0.1-300,000	H3F
1.0	1" x 3/16" x 1"	0.1-400,000	H3H
1.5	1-1/2" x 3/16" x 1"	0.1-600,000	L3H
2.0	2" x 3/16" x 1"	0.1-1,000,000	P3H
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0.5	1" x 3/8"	0.1-250,000	HC
1.0	1-1/4" x 1/2"	0.1-1,000,000	JD

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OTHER EXPORT
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Calendar of Coming Events and Authors' Deadlines*

(Continued from page 18A)

EIA-IRE Radio Fall Meeting, Sheraton Hotel, Rochester, N. Y., Oct. 27–29

East Coast Aero. & Nav. Elec. Conf., Lord Baltimore Hotel and 7th Regiment Armory, Baltimore, Md., Oct. 27–29 (DL*: May 15, W. A. Scanga, Aircraft Armaments, Inc., Cockeysville, Md.)

1958 Electron Devices Meeting, Shoreham Hotel, Washington, D. C., Oct. 30–Nov. 1 (DL*: Aug. 1, S. T. Smith, Code 5240, Naval Research Lab., Wash. 25, D. C.)

Conf. on Elec. Tech. in Medicine & Biology, Minneapolis, Minn., Nov. 12–14

Atlanta Section Conf., Atlanta-Biltmore Hotel, Atlanta, Ga., Nov. 17–19

NEREM, Mechanics' Bldg., Boston, Mass., Nov. 19–20

Acoustical Soc. of Amer., Chicago, Ill., Nov. 20–22

Elec. Computer Exhibition, Olympia, London, Eng., Nov. 28–Dec. 4

Eastern Joint Computer Conf., Bellevue-Stratford Hotel, Philadelphia, Pa., Dec. 3–5

Nat'l Symp. on Global Comm., Colonial Inn, Desert Ranch, St. Petersburg, Fla., Dec. 3–5 (DL*: Aug. 1, M. R. Donaldson, Elec. Commun., Inc., 1501 72nd St., N., St. Petersburg, Fla.)

PGVC Annual Mtg., Hotel Sherman, Chicago, Ill., Dec. 4–5

Mid-Amer. Elec. Convention, Mun. Audit., Kansas City, Mo., Dec. 9–11 (DL*: Aug. 1, Wilbert O'Neal, The Vendo Co., 7400 E. 12th St., Kansas City, Mo.)

1959

Rel. & Qual. Control Nat'l Symp., Bellevue-Stratford Hotel, Phil., Pa., Jan. 12–13

Transistor-Solid State Circuits Conf., Univ. of Pa., Philadelphia, Pa., Feb. 12–13

IRE Nat'l Convention, Coliseum and Waldorf-Astoria, New York City, Mar. 23–26

Nuclear Cong., Cleveland, Ohio, Apr. 5–10

SW Regional Conf., Dallas, Tex., Apr. 16–18

Nat'l Aero Elec. Conf., Dayton, Ohio, May 4–6

Seventh Region Conf., Univ. of New Mex., Albuquerque, N. M., May 6–8

Elec. Components Conf., Ben Franklin Hotel, Phil., Pa., May 6–8

WESCON, San Francisco, Calif., Aug. 18-21 (DL*: May 1, R. C. Hansen, Microwave Lab., Hughes Aircraft Co., Culver City, Calif.)

Int'l Congress on Acoustics, Stuttgart, Germany, Sept. 1–8 (DL*: May 15, I.E. Zwicker, Stuttgart N, Breitscheidstr. 3)

Nat'l Electronics Conf., Hotel Sherman, Chicago, Ill., Oct. 12–14

East Coast Aero & Nav. Conf., Baltimore, Md., Oct. 26–28

1959 Electron Devices Meeting, Shoreham Hotel, Washington, D. C., Oct. 29–31

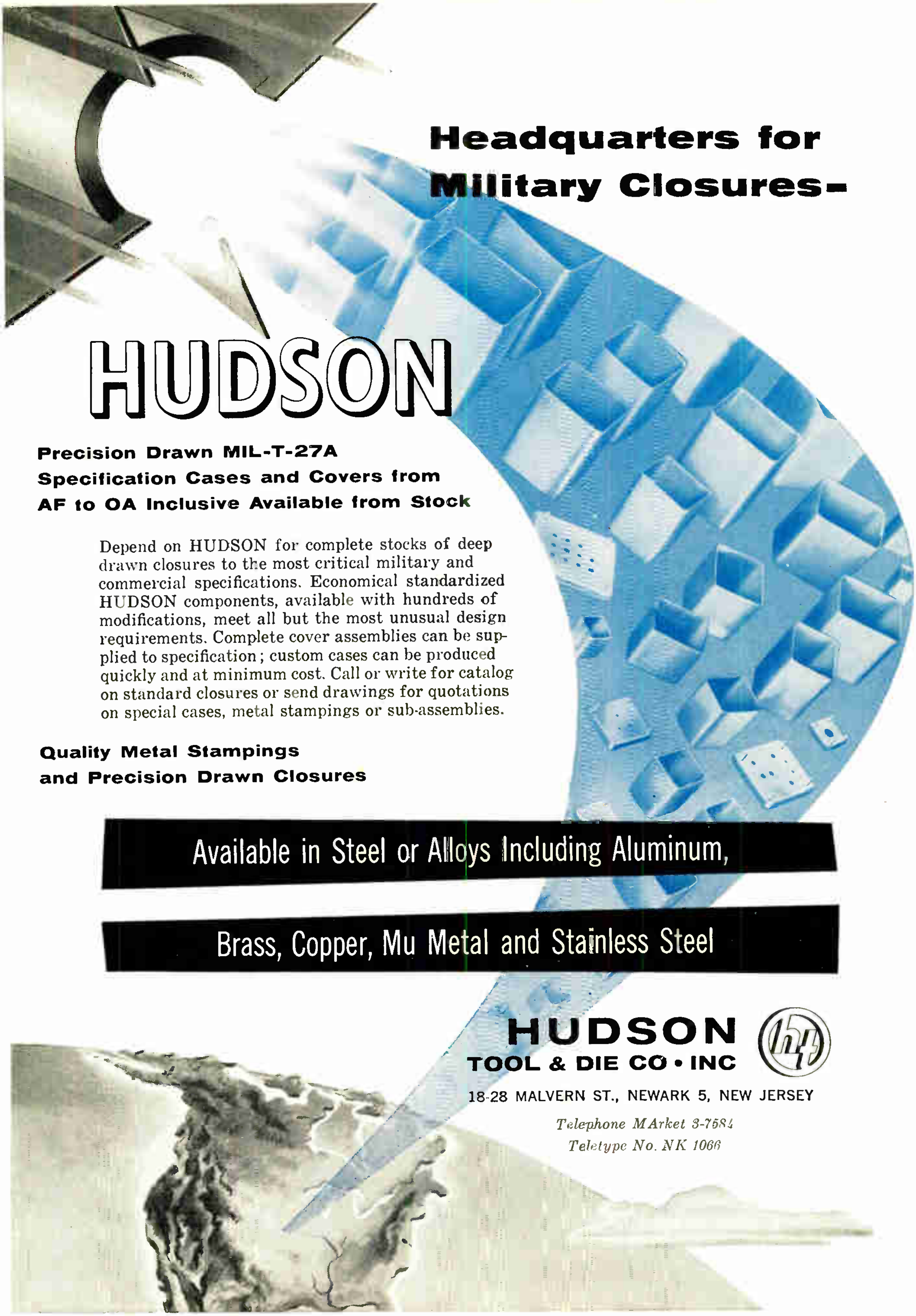
Nat'l Automatic Control Conf., Hotel Sheraton, Dallas, Tex., Nov. 4–6

* DL = Deadline for submitting abstracts.

At the recent annual banquet of the Washington Section, IRE President D. G. Fink delivered the dinner address and presented eight Fellow awards. Three Section members also received distinguished service citations, and six student members were awarded certificates. The picture above shows, from left to right: N. H. Street of George Washington Univ., R. J. Lindsey, Jr. and Wilbert Mason from Howard Univ., and George Abraham, Section Vice-Chairman of the Student Affairs Committee, who made the presentations of the six student awards.

Getting ready for the 1958 WESCON are (*left to right*): Gerald Goldenstern, registration vice-chairman; Fred MacKenzie, arrangements vice-chairman; Louis Holland, program committee member; Spencer Varian, arrangements chairman; and W. E. Peterson, convention director. WESCON is scheduled for Aug. 19–22 at Los Angeles.

About 1000 engineers and scientists attended the recent Reliability & Quality Control Symposium. Symposium program chairman, Clifford Ryerson (*left*); Maj. Gen. F. L. Ankenbrandt (*center*), keynote speaker; and M. M. Tall (*right*), general chairman, discuss the successful program of speakers, including one from England.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Library of Congress Changes Names of Two Publications

Effective immediately, the names of two Library of Congress monthly publications, which serve as keys to new literature coming into this country from the U.S.S.R. and East Europe, will be changed from "Lists" to "Indexes." The purpose of the change is to indicate to scientists and other scholars not familiar with the monthlies that they are more than mere "booklists" and that they contain information about the content of new Russian and East European books and articles.

The *Monthly List of Russian Accessions*, which enters its 11th year of publication in April 1958, becomes the *Monthly Index of Russian Accessions*. The Government Printing Office sells it for $12 a year ($15 abroad).

The *East European Accessions List*, published since 1951, becomes the *East European Accessions Index*. The GPO sells it for $10 a year ($12.50 abroad).

The change in the names of the two bibliographies does not mean a change in the editorial content of either one. It should also be noted that they contain information *about* the content of books and articles, *not* translations of entire books and articles.

Each index gives in English a monthly account of new material in a variety of subject fields as received from the U.S.S.R. and East Europe by the Library of Congress and by other American research libraries. The translation of all titles of books and articles into English and the elaborate subject guides in English permit a researcher who has no command of Russian or East European languages to identify easily the material important to him.

AIEE Holds Magnetics Meeting

The AIEE-sponsored 1958 Special Technical Conference on Nonlinear Magnetics and Magnetic Amplifiers, convening at the Hotel Statler, Los Angeles, Calif., Aug. 6–8, 1958 will present a technical program divided into four sessions. This will include technological and theoretical aspects of nonlinear magnetics and magnetic amplifiers, computer applications, special purpose devices and applications, and "new frontiers" in this ever-expanding field. Displays by forty manufacturers in the magnetics field will be an additional feature of the program.

For the first time through a technical conference, the AIEE will offer a scholarship in the name of the author of the best judged technical paper, presented to the university of his choice.

International Conference on Acoustics Meets in Stuttgart

The Third International Congress on Acoustics will be held Sept. 1–8, 1959 in Stuttgart, Germany. It is sponsored by the International Commission on Acoustics and arranged by the "Verband Deutscher Physikalischer Gesellschaften" together with the "Nachrichtentechnische Gesellschaft des Verbandes Deutscher Elektrotechniker" and the "Verein Deutscher Ingenieure."

The technical sessions will be held in Stuttgart's "Liederhalle" and in the lecture theaters of the "Technische Hochschule Stuttgart."

The program will cover the entire field of physical and technical acoustics with some preference to noise and vibration control, electroacoustics, architectural acoustics and physical acoustics. Invited papers are planned on the following subjects: acoustical methods in solid state physics, sound propagation in gases under low pressure, turbulence and sound, cavitation, investigation of molecular structure by ultrasonics, physiological acoustics, psychological acoustics, stereophony, room acoustics, building acoustics, noise control, information theory, and speech.

In addition to the technical program, participants will enjoy a combined technical and sightseeing excursion to Munich as well as bus trips to the environs of Stuttgart.

Modern acoustical and electrical measuring devices will be shown in a special industrial exhibition in connection with the congress.

Alien participants, especially those from overseas, will be given the opportunity to visit German research institutes before or after the congress.

Further information can be obtained from Dr.-Ing. Eberhard Zwicker, Stuttgart N, Breitscheidstr. 3. Announcements of contributed papers, which are accepted until May 15, 1959, must be sent to the same address.

1958 IRE President D. G. Fink (*center*) chats with A. G. Clavier (*left*) and Henri Busignies (*right*) on a recent tour of the IRE Northern New Jersey Section. Mr. Clavier and Mr. Busignies are IRE Fellows and vice-president and technical director, and president, respectively, of the Federal Telecommunication Laboratories.

7th Region Conference Emphasized Electronics Progress

The Seventh Region IRE Conference and Trade Show opened at the Hotel Senator, Sacramento, Calif., April 30–May 2.

Technical sessions on military electronics, communications, modern engineering management, industrial electronics, instrumentation, microwave development, electronic devices, and computers revolved around the conference theme of "Accelerated Progress in Electronics." The sessions and the trade show were held in the Merchandise Mart Building of the California State Fair grounds. Student papers were also presented at an evening session on May 1.

Highlights of the conference were a cocktail party on April 30 and an all-industry luncheon on the following day. The luncheon speaker was Lewis Dunn, Executive Vice-President and Chief Engineer, Space Technology Labs., Div. of Ramo-Wooldridge Corp., who spoke on "What Future Demands Will Space Technology Make on Electronic Engineers?"

International Cybernetics Convention Meets in Belgium

The Second International Cybernetics Congress will meet at Namur, Belgium, Sept. 3–10, 1958. The congress will meet on Sept. 3, and adjourn briefly for two days to visit the Brussels Fair before reconvening again on Sept. 8.

Subjects to be dealt with at the congress will include information, automatic machinery, automation, economic and social effects of automation, cybernetics and the social sciences, and cybernetics and biology. The languages used will be French and English. Papers will be read during the mornings, and group discussions on these papers will gather during the afternoons. Publication of convention proceedings is also planned.

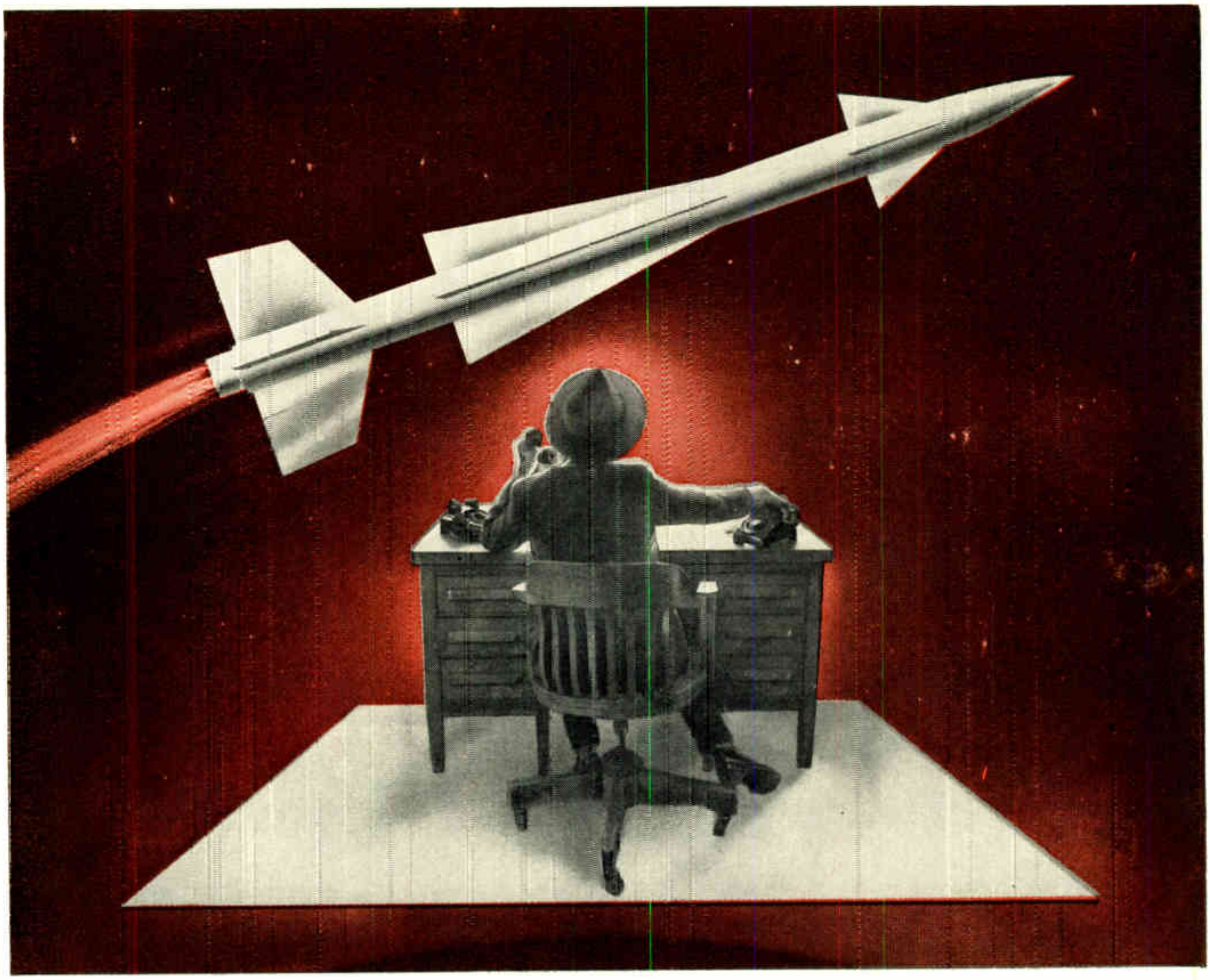
Those wishing to attend the congress should contact the Secretariat of the International Association for Cybernetics, 13, rue Basse-Marcelle, Namur, Belgium. The admission fee for members of the International Association for Cybernetics is 200 frs.; for non-members, 400 frs. Authors of papers will be admitted free.

Authors of papers are requested to send titles and summaries of their papers to the Secretariat no later than June 1, 1958. Each summary should not exceed in length one typewritten page.

The organizing committee of the congress is made up of: G. R. Boulanger, president; René Close, Louis Couffignal, John Diebold, W. G. Walter, all members; and Josse Lemaire, permanent delegate.

Professional Group News

A special issue of the Transactions of the Professional Group on **Circuit Theory**, tentatively scheduled for March, 1959, will be dedicated to papers on sequential transducers. This relatively new study forms a theoretical basis for the understanding of the capabilities of switching circuits and digital



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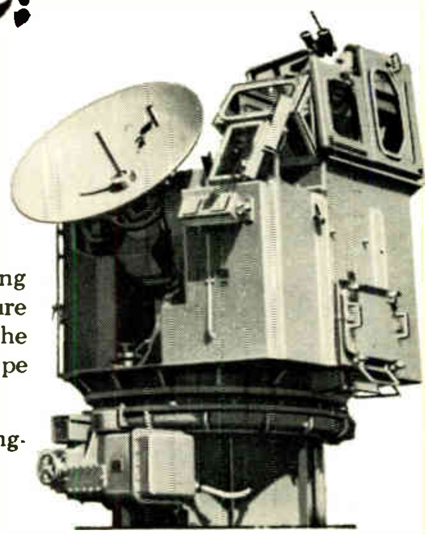

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computer control, and coding circuits. Suitable topics for papers submitted for consideration will include: algebra for the terminal description of binary filters; studies of the memory requirements of sequential circuits; interrelationships between finite-state transducers and turing machines; linear and nonlinear recursive functions and sequence filters; internal circuit redundancy for purposes of terminal error-corrections; and application of old and new synthesis methods to sequential circuits using transistors, magnetic cores, and superconductive elements. Emphasis will be placed on general theoretical techniques rather than the design of specific equipment.

All decisions by the Editorial Committee as to the suitability of papers submitted for publication in the "Sequential Transducers" issue will be made by November 1, 1958. Papers initially submitted in extended abstract form (about 1000 words) by July 15, 1958 will be given priority in consideration. In any case the deadline for papers, complete with standard abstract and drawings suitable for reproduction, will be December 1, 1958. The preparation of manuscripts should follow the suggestions in "Preparation and Publication of I.R.E. Papers," PROC. I.R.E., November, 1954, pp. 1604–1605.

Please address all correspondence about the "Sequential Transducers" issue of the PGCT TRANSACTIONS to: D. A. Huffman, Department of Electrical Engineering, Massachusetts Institute of Technology, 77 Massachusetts Ave., Cambridge 39, Mass.

1958 National Conference on Aeronautical Electronics

BILTMORE HOTEL, DAYTON, OHIO, MAY 12–14, 1958

MONDAY, MAY 12

9:00–12 NOON

Electronics and Space Flight

Moderator: Frank Lehan, Ramo-Wooldridge Corp.

Radio Astronomy and Outer Space, F. T. Haddock, Univ. of Michigan.

United States Satellite, Eberhardt Rechtin, Jet Propulsion Lab., California Inst. of Technology.

Mid-Course Guidance Techniques for Space Vehicles, A. D. Wheelon, Ramo-Wooldridge Corp.

New Techniques Required for Navigation, Guidance and Control of Orbiting Vehicles, R. L. Berg and F. H. Kierstead, Goodyear Aircraft Corp.

Ground-to-Space Propagation Phenomena, E. R. Moe, Lockheed Aircraft.

Hypersonic Vehicle Antenna Design, A. S. Dunbar, Lockheed Aircraft.

Air Safety

Moderator: D. S. King, Civil Aeronautics Administration.

Time Shared Indicators for Displaying Non-Synchronous Data from Two Radars, R. L. Sorenson, Civil Aeronautics Administration, Technical Development Center.

High Altitude Automatic Flight Checking on the Federal Airways System, H. I. Metz, Civil Aeronautics Administration.

How Firm Is our Foundation, K. L. Brannon, Civil Aeronautics Administration.

Development of ATC Procedures for Civil Jet Aircraft, P. T. Astholz and T. K. Vickers, Civil Aeronautics Administration, Technical Development Center.

Airborne Collision Avoidance Systems, F. C. White, Air Transport Association.

Electronic Components I

Engineers' Club—West Room

Moderator: H. V. Noble, Electronics Components Lab., Wright Air Development Center.

Astro Electronic Component Parts, R. C. Marshall, Electronic Components Lab., Wright Air Development Center.

Behavior of Electronic Components in Noise Fields, G. W. Kamperman and J. J. Baruch, Bolt, Beranek and Newman.

High Frequency Transistor Performance, Neil DiGiacomo, Electronic Components Lab., Wright Air Development Center.

Miniature High Frequency Transistorized Ferri-Inductive Devices, E. Abbot, Emerson Radio & Phonograph Corp. and Capt. C. K. Greene, Electronic Components Lab., Wright Air Development Center.

2–5 Amperes Silicon Power Transistor, H. W. Henkels and T. P. Nowalk, Westinghouse Electric Corp.

Thermal Design I

Engineers' Club—East Room

Moderator: Walter Robinson, Consulting Engineer.

Experimental Results and Application Considerations for Integrating Evaporative Cooling into Electronic Equipment, P. J. Zukauskas, and E. K. Cameron, Martin Co.

A 5KW Airborne Evaporative Cooled Power Amplifier, D. W. Smith, Martin Co.

Electronic Cooling by Simultaneous Heat and Mass Transfer, A. R. Saltzman, B. T. Plizak, L. F. Tomko, and J. Nycum, Johnsville Naval Air Development Center.

Cooling Techniques for Electronic Devices Operating at 300–500°C Ambient Temperature, J. P. Welsh, Cornell Aeronautical Lab.

Thermal Design and Evaluation of Airborne Integrated Communication Navigation Systems, A. H. Schroeder and C. P. Toussaint, Collins Radio Co.

Compact Expendable-Refrigerant System with Maximum Coolant Utilization, F. E. Carroll, United Aircraft Products.

Communication Systems

Biltmore Ballroom

Moderator: Tom Rogers, Communication Lab., Air Force Cambridge Research Center.

A Digital Data Transmission System using Phase Modulation and Correlation Detection, F. A. Losee, Hughes Aircraft Co.

Improved Communications Through Synchronous Detection, R. J. Lutze, General Electric Co.

Auroral Zone-to-Air-Ground Communication Problems, W. R. Vincent, R. L. Leadabrand, and A. M. Peterson, Stanford Research Inst.

Air-to-Ground Meteor-Burst Communications, A. M. Peterson, Stanford Research Inst. and Irwin Roth, Stanford Res. Inst.

Current Problems in Ionospheric Scatter Communications, Ross Bateman, Page Communications Engineers, Inc.

Design of Point-to-Point Communication Systems for 400 to 600 Mile Paths, W. E. Morrow, Jr., Lincoln Lab., M.I.T.

AFTERNOON

2:00–3:15 P.M.

Panel Discussion A: Electronics and Space Flight

Biltmore Ballroom

Moderator: F. W. Lehan; Panel Members: F. T. Haddock, Eberhardt Rechtin, A. D. Wheelon, R. L. Berg, F. H. Kierstead, E. R. Moe, and A. S. Dunbar.

3:45–5:00 P.M.

Panel Discussion B: Communication Systems

Moderator: Tom Rogers; Panel Members: F. A. Losee, R. J. Lutze, W. R. Vincent, R. L. Leadabrand, A. M. Peterson, Ross Bateman, W. E. Morrow, Jr., and Irwin Roth.

2:00–5:00 P.M.

Navigation I—Doppler

Engineers' Club—Auditorium

Moderator: J. L. Dennis, Consulting Engineer.

Back-Scattering Characteristics of Land and Sea at X-Band, J. P. Campbell, General Precision Lab.

A Low Noise CW Doppler Technique, K. C. M. Glegg, Canadian Marconi Co.

AN/APN-78, A Doppler Navigator for Helicopters, N. L. Laschever, Lab. for Electronics, Inc.

Continuous Wave Doppler Navigation, J. J. Bowden, Ryan Aeronautical Co.

Antenna Transmitter-Receiver, H. C. Bussey, C. A. Zielinski, General Elec. Co.

Approach and Landing

Arranged and Presented by the Institute of Aeronautical Sciences

Moderator: C. R. Bryan, Flight Control Lab., Wright Air Development Center.

Approach and Landing Problems, H. K. Fletcher and F. O. Powell, Bell Aircraft Co.

Automatic Approach and Landing Flare Control, E. R. Buxton, Autonetics Div., North American Aviation, Inc.

Airport Surface Detection Equipment at Idlewild, J. E. Woodward, Airborne Instruments Lab.

Components II

Engineers' Club—West Room

Moderator: A. H. Dicke, Electronic Components Lab., Wright Air Development Center.

A Rugged, Permanent Magnet Focused 1-Watt Traveling Wave Amplifier at S Band, P. E. Hargrove, Sylvania Electric Prod., Inc.

A Medium Power, High Gain, X-Band Traveling-Wave Tube for Use in High Performance Aircraft, N. M. Gutlove, Sperry Gyroscope Co.

A Rugged C-Band Traveling Wave Tube, Laird Haas, Eitel-McCullough, Inc.

Electron Tube Parameter Variability as a Restriction on Circuit Design, R. S. Whitlock and R. C. Radeleff, Electronic Components Lab., Wright Air Development Center.

Recent Developments and Display Techniques for Aircraft Display, R. W. Deichert, Allan B. DuMont Labs.

Thermal Design II

Engineers' Club—East Room

Moderator: J. R. Welsh, Cornell Aeronautical Lab.

Parameters for Rating Heat Transfer Fluids, D. R. Fairbanks, Raytheon Manufacturing Co.

Design Techniques for Weight Reduction Within the Evaporative-Gravity Cooling System, Melvin Mark, Consulting Engineer, Mr. Stephenson and C. E. Goltsos, Raytheon Manufacturing Co.

Some Properties of Two Fluorochemicals Used as Immersion Coolants for Airborne Electronic Equipment, J. R. Shackleton, Hughes Aircraft Co.

Correlations of Thermal Performance Data of Forced Air Cooled Airborne Electronic Equipment, Walter Robinson, Consulting Engineer.

Analytical Approaches to Electronic Equipment Forced Cooling Problems, B. W. Randolph, Hughes Aircraft Co.

Electron Tube Thermal Performance, R. L. Santin, General Electric Co.

Circuits

Moderator: Constantine Houpis, United States Air Force Inst. of Technology.

Video Pulse Reshaping Circuitry with Application in Purse Train Decoding Equipments, M. H. Murphy and R. W. Griggs, Packard-Bell Electronics Corp.

Pulse Spectrum Control Techniques for Close Channel Spacing, H. B. Scarborough and J. Hobbs, Federal Telecommun. Labs.

Effect of Video Bandwidth on Receiver Sensitivity, C. R. Ammerman and W. L. Blair, Haller, Raymond and Brown, Inc.

Some Practical Design Aspects of Multiple-Tuned Circuits, J. H. Griswold, General Electric Co.

Synchronous Integrating Circuit, L. E. Tryon, Raytheon Manufacturing Co.

6:45 P.M.

BILTMORE HOTEL BALLROOM

Annual Banquet

TUESDAY, MAY 13

9:00–12 NOON

Air Traffic Control I

Engineers' Club—Auditorium

Moderator: V. I. Weihe, General Precision Equipment Corp.

The Airways Modernization Board Five-Year Program, Col. C. B. Fisher, Airways Modernization Board.

Operational Aspects of the Data Processing and Display System, R. F. Link, Airways Modernization Board.

The Data Processing and Display Technical Program, R. A. Finkler, General Precision Lab.

Data Acquisition Aspects of the AMB Five-Year Program, W. N. Pike, Airways Modernization Board.

Communications for the Five-Year Program, Nathan Marchand, Marchand Electronics Labs.

Telemetering

Engineers' Club—East Room

Moderator: N. M. Nichols, Univ. of Michigan.

Advances in Airborne Telemetering Equipment, K. M. Uglow, Consulting Engineer.

Techniques for Interplanetary Telemetering, Hans Scharla-Nielson, Radiation, Inc.

Future Requirements for Range Telemetering, C. H. Smith, Autonutronic Systems, Inc.

Ground-Air-Data Transmission and Air Traffic Control, I. H. Bowker, Communication and Navigation Lab., Wright Air Development Center and R. G. Ferguson, Air Research and Development Command.

Computers I

Engineers' Club—West Room

Moderator: Eldred Nelson, Ramo-Wooldridge Corp.

The Central Computer in the Man-Machine System, R. L. Terry, Litton Industries.

Transac, Transistorized Airborne Computer, G. L. Hollander, Philco Corp.

A Military Aircraft Marshalling System, H. J. Bramson, Hughes Aircraft Co.

CP-209 Digital Computer Flight Test Program, R. E. Berri, Librascope, Inc.

A Digital High Speed Coordinate Conversion System, Max Palevsky, Packard Bell Computer Corp.

Radar

Biltmore Ballroom

Moderator: L. R. Cutrona, Univ. of Michigan.

Synthesis of Comb Filters, Warren White, Airborne Instruments Lab.

Bistatic Radars and Forward Scattering, K. M. Siegel, Univ. of Michigan.

Application of Pulsed Doppler to Airborne Radar Systems, W. W. Maguire, Hughes Aircraft Co.

Electronic Antenna Scanning, H. R. Senf, Hughes Aircraft Co.

Application of Precision Tracking Radar to Location Control and Data Transmission for an Unmanned Observation Platform, D. K. Barton, Radio Corp. of America.

Detector Range of CW & Pulse Doppler Radars, J. J. Bussgang, P. Nesbeda, and H. Safran, Radio Corp. of America.

Operations Research

Moderator: Max Astrachan, United States Air Force Inst. of Technology.

An Electronics Engineer's View of Operations Research, J. F. Digby, The Rand Corp.

Operations Research in Life Testing, L. G. Johnson, General Motors Corp.

Operations Research Based on Simulation for Training, W. C. Biel, System Development Corp.

The Monte Carlo Method as a Decision Aid in Airways Modernization, Alfred Blumstein, Cornell Aeronautical Lab.

AFTERNOON

2:00–5:00 P.M.

Forum: Patterson High School Auditorium

Subject: Air Traffic Control in the Jet Age.

Moderator: Mr. Pyle, Civil Aeronautics Administration.

WEDNESDAY, MAY 14

9:00–12 NOON

Air Traffic Control II

Engineers' Club—Auditorium

Moderator: Nathaniel Braverman, Airways Modernization Board.

Digital Simulation of Air Traffic Control Systems, G. W. Bond, K. S. Gale, and C. J. Moore, Armour Research Foundation, Illinois Inst. of Technology.

Mathematical Analysis of Some Phases of

Control of Air Traffic Flow, R. S. Miller, Franklin Inst. of Pennsylvania.

The Dead Reckoning Computer in Air Traffic Control, Paul Weinschel, Ford Instrument Co.

Communication Capabilities of the Air Traffic Control Radar Safety Beacon, Kurt Merl, Ford Instrument Co.

Universal Signalling System for Common Airspace Control, B. H. Baldridge and E. V. Hogan, General Electric Co.

Civil Avionic System Design

Biltmore Hotel Ballroom

Moderator: W. T. Carnes, Jr., Aeronautical Radio, Inc.

The Development of Operational Requirements for Civil Avionic Systems, D. S. Little, American Airlines.

System Concepts in Avionic System Design, H. K. Morgan, Bendix Aviation Corp.

The Design of Airborne Equipment for Avionic Systems, George Church, Bendix Radio, Bendix Aviation Corp.

The Airline Contribution to System Development, B. F. McLeod, Pan American World Airways, Inc.

The Integration of Avionic Systems in the Modern Commercial Aircraft, R. H. Jerome, Douglas Aircraft Co.

Communication and Navigational Aid Standards for International Civil Aviation, C. C. E. Bellringer, International Civil Aviation Organization.

Computers II

Engineers' Club—West Room

Moderator: C. C. Goldman, Aeronautical Research Lab., Wright Air Development Center.

A Simple, Transistorized Device for Precise Digital-to-Analog Conversion, William Hiatt, General Electric Co.

A Direct Frequency to Parallel Binary Converter Using Operational Digital Computing Circuits, Edward Ostroff, Lab. for Electronics.

The Analytic Versus the Spring Flight Path Program Computer, W. L. Harmon, Weapons Guidance Lab., Wright Air Development Center.

A Coasting Time Computer, Donald Colbert, Electronic Communications, Inc.

The ASN-7 Navigation Computer—Past, Present and Future, S. I. Frann, and J. A. Quinn, Ford Instrument Co.

Antennas

Engineers' Club—East Room

Moderator: T. C. Tice, Antenna Lab., Ohio State Univ.

An S-Band Slotted Waveguide Antenna, R. Fratila, Melpar, Inc.

Tri-Scanner Antenna, Jesse Butler, Sanders Associates, Inc.

Omni-Directional Reflecting Lens, E. M. Lipsey, United States Air Force Inst. of Technology.

Microwave Wire Antennas, J. A. Kuecken, Avco Manufacturing Corp., Crosley Div.

Scimitar Antenna and Its Application to Modern Aircraft, E. M. Turner, Aerial Reconnaissance Lab., Wright Air Development Center.

Reliability

Moderator: F. J. Ruther, Electronic Computers Lab., Wright Air Development Center.

The Basis and Use of a Method for Predicting Reliability, J. H. Hershey, Bell Tel. Labs.

Consideration of Deterioration Effects in Equipment Reliability Improvement Programs, S. R. Scott, Aeronautical Radio, Inc.

Engineering Organization for Reliability Control, C. M. Ryerson, Radio Corp. of America.

Proof of Reliability, F. J. Ruther, Electronic Components Lab., Wright Air Development Center, and L. D. Smith, General Electric Co.

Graphical Estimation of Reliability, B. E. Phillips, Martin Aircraft Co.

AFTERNOON

2:00–3:15 P.M.

Panel Discussion C: Civil Avionic System Design

Biltmore Ballroom

Moderator: W. T. Carnes, Jr.; Panel Members: D. S. Little, H. K. Morgan, G. Church, B. F. McLeod, R. H. Jerome, and C. E. Bellringer.

3:45–5:00 P.M.

Panel Discussion D: Reliability

Moderator: F. J. Ruther; Panel Members: J. H. Hershey, S. R. Scott, C. M. Ryerson, B. E. Phillips.

2:00–5:00 P.M.

Navigation II

Engineers' Club—Auditorium

Moderator: Sven Doddington, Federal Telecommunication Labs.

Airborne Earth Rate Directional Heading Reference, D. J. Atwood, Dynatrol Corp.

The Correlation Air Navigator, A Vertically Beamed Radar, F. R. Dickey, Jr., General Electric Co.

A Mobile TACAN Beacon; Its Description and Uses, Anthony Casabona, Federal Telecommunication Labs.

Synchronous Distance Measurement, N. E. McIver, Bendix Radio Corp.

Semi-Automatic Radar Guidance System for Manned Aircraft, Vefik Basman, R. C. Tabor, and R. E. Scott, Univ. of Michigan.

Navigation Methods and Systems, T. P. Haney, Westinghouse Electric Co.

Military System Management

Moderator: P. R. Murray, Air Research and Development Command.

Dynamics and Weapons Systems Management, R. H. Widmer, Convair.

New Challenges for the Systems Manager, H. W. Merrill, Martin Co., Denver, Div.

The Key to Weapons System Management Communication, R. E. Galer, Temco Aircraft Corp.

Management of the SNARK Missile Weapon System, George Douglas, Northrop Aircraft, Inc.

Contractor Management of Military Weapon Systems, G. Stoner, Boeing Airplane Co.

The Weapon System Partnership, J. G. Beerer, North American Aviation, Inc.

Avionic Equipment

Engineers' Club—West Room

Moderator: Ray Nordlund, Weapons Guidance Lab., Wright Air Development Center.

A Missile Command Receiver, J. F. Clemens, Avco Manufacturing Corp.

The Advantages of Dynamic Problem Simulating Test Equipment, J. S. Williams, Radio Corp. of America.

Design Considerations for Automatic Test Equipment, Ronald G. Matteson and J. A. Huie, Stromberg-Carlson Div., General Dynamics. Corp.

Static Power Supply Design for Emergency Low Voltage Operation, D. L. Theurer, Magnavox Co.

Production Techniques

Moderator: D. E. Noble, Motorola.

Changes Reach Further than you Think, D. B. Dobson, Radio Corp. of America.

Why the Air Force Requirement for Burn-In and Shake-Down of Data Link Equipment, V. J. Carpantier, Communication and Navigation Lab., Wright Air Development Center.

Correlation of Production Failure Reduction and Field Performance Programs, N. S. Barnes and D. C. Meeks, General Electric Co.

A Systematic Method for the Preparation of Complicated Wiring and Cabling Drawings, E. J. Weber, Martin Company, and G. E. Alexander, Martin Co.

Plating of Magnesium, A. C. Dwyer, Raytheon Manufacturing Co.

Simulators

Engineers' Club—Auditorium

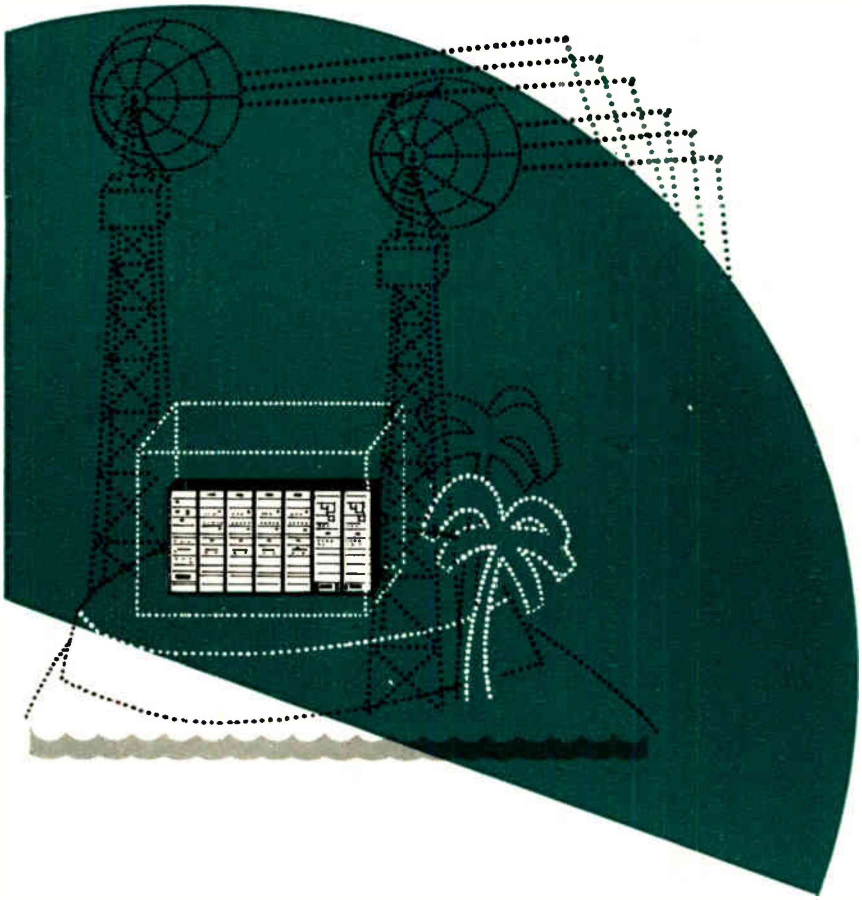
Moderator: V. S. Haneman, United States Air Force Inst. of Technology.

Flight Simulators and Their Uses, D. H. Lipscomb, Nortronics, Northrop Aircraft Co.

The Force Feel Regulator, Mitchell Streicher, Electronics Communications, Inc.

A General Purpose Artificial Feel System for Human Factors Research, F. C. Williams, Hughes Aircraft Co.

Radar Simulator in Optimization of the Detection Range Performance of a Search Radar, B. S. Bylinkin, Autonetics Div., North American Aviation.

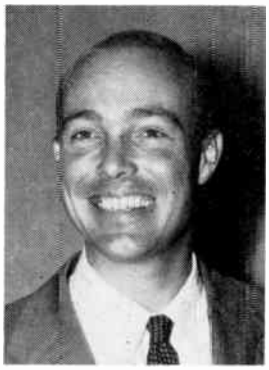


Second National Conference on Production Techniques

HOTEL NEW YORKER, NEW YORK CITY, N. Y., JUNE 4–6, 1958

THEME: "PREVIEWING ELECTRONIC DESIGN TRENDS"

W. D. NOVAK
Gen. Precision Lab.
General Chairman



R. L. SWIGGETT
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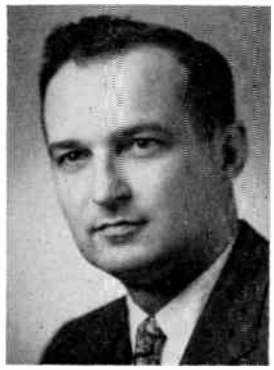
J. W. TRINKAUS
Sperry Gyroscope Co.
Program

F. COLLINGS
Radio Corp. of America
Local Arrangements



W. E. VANNAH
Control Eng. Magazine
Publicity

The Second National Conference on Production Techniques, to be held at the Hotel New Yorker, New York City, June 4–6, 1958, will be sponsored by the IRE Professional Group on Production Techniques. Conference chairmen are: W. D. Novak, General Chairman; J. W. Trinkaus, Program; R. L. Swiggett, Exhibits; F. Collings, Local Arrangements; and W. E. Vannah, Publicity.

A special feature of the conference will be a session on electronic developments abroad, attended by speakers from England, Sweden and Russia, and also reports from France and Germany.

Registration opens at 9 A.M., June 4, on the mezzazine of the Hotel New Yorker. Inspection trips, each limited to fifty persons, will be made to Dumont Labs. in E. Paterson, N. J. and Western Electric Co. in Kearny, N. J. Tickets for these trips, to be taken on June 4 and June 6, can be had at the registration desk from W. Ellwood, Inspection Trips Chairman. Plans for a cocktail party and banquet are also now underway, and tickets for the luncheon on June 5 are available at $3.75 per person.

Advance mail registrations can also be secured from F. Collings, Local Arrangements Chairman, RCA, Moorestown, N. J.

A dozen exhibits, with space for yet a few more, will be on display at the Hotel New Yorker during the course of the conference. Exhibitors will be able to demonstrate their products and services between sessions through the courtesy of Gen. Precision Lab. via a closed-circuit television system of the projection screen type to this highly selective audience. A TV camera will be at the disposal of all speakers who wish to augment their talks with live demonstrations of their equipment.

Papers presented at this conference will later be made available in the TRANSACTIONS.

JUNE 5, 9:30 A.M.–12 NOON

Tomorrow's Components

Session Co-Chairmen, F. Collings, Radio Corp. of America, and J. Brothers, Philco Corp.

Modular Dimensioning of Electronic Parts for Mechanized Assembly, W. Lane, United States Army Signal Engineering Labs.

Future Components For Mechanization, E. Plesser, Philco Corp.

Design for Tommorrow's Capacitor Components, A. Tiezzi, Sprague Electric Co.

Circuit Miniaturization, E. Leyonmark, Photocircuits Corp.

LUNCHEON

Philosophical Treatment of Conference Theme

Session Chairman: E. Martin, Sperry Gyroscope Co.

Com. G. W. Hoover, Weapons and Systems, Air Branch, ONR.

2:30–5:00 P.M.

Developments Abroad

Session Co-Chairmen: I. Weir, General Electric Co., and R. Geshner, Stromberg-Carlson Co.

The Change-Over in the Production Techniques When Introducing Printed Circuits, W. Flack, Radio and Allied Industries, Ltd., Slough Bucks, England.

Electronic Production and Engineering Techniques in France, N. Wunderlich, International Electronics Corp.

Germany's Training of Workers and Standardizations in Design and Production of Electronic Equipment, H. Ruehlemann, Elco Corp.

Electroforming of Millimetric Waveguide Components, A. Morrison, Mullard Research Labs., Surrey, England.

Title to be announced, Col. E. Kiessling, USAF Andrews Field, Washington, D. C.

Title to be announced, M. Eisler, England.

Title to be announced, M. Westberg, Sweden.

A speaker from Russia and his paper will be announced.

JUNE 6, 9:00 A.M.–11:30 A.M.

Designing with New Materials

Session Co-Chairmen: L. Schwartz, Remington Rand Univac, and A. Middleton, Sandia Corp.

Tuf-Plate Hole for Printed Circuits, George Geddy, Photocircuits Corp.

Designing with Poly-Strip Flat Wire Cable, A. Pugh, International Resistance Corp.

The Impact of Transistors on Military Electronic Design, A. Jacobsen, Motorola Inc.

Micro-Modules, Components and Materials Requirements, V. Kublin, U. S. Army Signal Engineering Labs.

1:00–4:00 P.M.

Production Tool Advances

Session Co-Chairmen: R. Marder, Stavid Engineering, Inc., and J. Carr, Melpar, Inc.

Microminiaturization Techniques, M. J. Finkelstein, M-F Electronics.

Punched Card Controlled Component Insertion Machine, K. H. Hazel, IBM.

Evolution of a Small Lot System for the Production of Electronic Equipment, C. P. Cardani, United Shoe Machinery Corp.

Mechanized Assembly Equipment for Small Cup Type Speakers, Nils Lindhe, RCA.

New Possibilities for Automatically Molding Small Parts Economically, J. L. Hull, Hull Corp.

Automation of Floated Gyro Drift Measurements, J. G. Nelson, Minneapolis Honeywell Regulator Co.

Inspection Trips

Chairman: W. Ellwood, Bell Telephone Labs.

JUNE 4, 1:00—5:00 P.M.

Dumont Laboratories, Inc.,
East Paterson, N. J.

JUNE 6, 8:30 A.M.–1:00 P.M.

Western Electric Co., Kearny, N. J.

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Second National Convention on Military Electronics

JUNE 16, 17, 18, 1958, SHERATON PARK HOTEL, WASHINGTON, D. C.

The Second National Convention on Military Electronics, sponsored by the IRE Professional Group on Military Electronics, will be held at the Sheraton Park Hotel, Washington, D. C., June 16–18, 1958. The convention theme will be "Missiles and Electronics—1958" 'Ad Astra Per Aspera.' The speaker at the convention banquet, on the evening of June 17, will be W. M. Holaday, Guided Missile Program, Department of the Defense, who will address the assembly on "Progress in the Guided Missiles Program."

Brig. Gen. W. B. Larew (Ret.) is President of the Executive Committee planning the convention. With him on the committee are: P. D. Foote, W. M. Holaday, R. Adm. W. D. Irvin, Lt. Gen. J. M. Gavin, R. Adm. Rawson Bennett, Lt. Gen. C. S. Irvine, Adm. Arleigh Burke, R. Adm. W. E. Cleaves (Ret.), Capt. C. L. Engleman (Ret.), and A. H. Schooley, advisors; Morton Gale, secretary; L. D. Whitelock, exhibits; Henry Randall, finance; D. J. McLaughlin, technical; R. H. Cranshaw, arrangements; B. J. Goldfarb, Transactions; and George Rappaport, public relations.

MONDAY, JUNE 16

2:00–5:00 p.m.

SPACE VEHICLE ELECTRONICS (CONFIDENTIAL)

Sponsored by the Air Research and Development Command.

Moderator, To be announced.

Approaches to Microwave Tube Design for Outer Space Application, J. R. Hayes, Electronic Components Lab., WADC, Wright Patterson Air Force Base.

A Ground Controlled System for Firing a Third Stage of Vanguard, A. B. Bligh, J. J. Fleming, and D. H. Gridley, U. S. Naval Research Lab.

Accelerometer-Integrator and Digital Velocity Meter, R. M. Zehr, D. D. Lenhart, and R. J. Arthur, Bell Aircraft Corp.

Design for Space-Vehicle Control System Reliability, R. O. Anderson, Flight Control Laboratory, WADC, Wright Patterson Air Force Base.

Radiation Characteristics of Slot Arrays on Conical Surfaces, M. G. Chernin, H. E. Shanks, and R. E. Plummer, Hughes Aircraft Co., and R. F. Goodrich, R. W. Kleinman, A. L. Maffett, C. E. Schensted, and K. M. Siegel, Univ. of Michigan.

Nuclear Radiation Effects on an Electronic System, Nathan Ehrlich, Bell Tel. Labs.

RELIABILITY TECHNIQUES AND PROCEDURES

Moderator, to be announced.

A New Role for Reliability Stress Analysis—Systematizing the Attack on Maintainability Problems, H. L. Wuerffel and D. I. Troxel, Radio Corporation of America.

A Method for Demonstrating Missiles Recurrently with Development, J. H. Yuch, Polaris Project, Lockheed Missile Systems Div., and Maj. R. G. Harris, USMC.

On the Meaning of Mean Time to Failure, Trevor Clark, Westinghouse Electric Corp.

Reliability Insight, M. L. Furst, ARMA Div., American Bosch Arma Corp., and Denis Krusos, Garden City, N. Y.

Techniques for the Evaluation of Guided Missile Reliability, R. E. Fontana and C. E. Parker, U. S. Naval Air Missile Test Center.

DATA HANDLING SYSTEMS AND DATA LINKS

Moderator: R. M. Maiden, Head Integration Techniques Section, Naval Research Lab.

A Data Handling Input-Output System for RAYDAC, J. P. Harvey, U. S. Naval Air Missile Test Center.

A Comparison of Binary Data Transmission Systems, J. G. Lawton, Cornell Aeronautical Lab.

Comparative Evaluation and Optimization of Airborne Target Detection Systems, N. S. Potter, Westinghouse Electric Corp.

Operational Analysis as Applied to Optimizing Interceptor Weapon System Subsystem Performance Requirements, R. Clanton and A. Schwartz, Westinghouse Electric Corp.

A Graph and Digit Printer with Analog Input, S. D. Warner, Convair,

COMPONENTS FOR THE MISSILE ENVIRONMENT

Moderator, to be announced.

Capacitive Transducers in Missile Guidance Systems, W. W. Breden, General Electric Co.

A Small, Light-Weight Traveling-Wave Tube for Military Applications, H. J. Wolkstein and R. Pekarowitz, Electron Tube Div., RCA.

Ultra High Temperature Electronic "Black Box" Development, W. L. Frisby, Missile and Ordnance Systems Dept., General Electric Co.

High Temperature Printed Circuits, J. J. Kinsella, Physics of Materials Dept., The Haloid Co.

TUESDAY, JUNE 17

9:00–12 Noon

RANGING AND TRACKING (CONFIDENTIAL)

Sponsored by the Air Research and Development Command.

Moderator: A. A. Varela, Melpar, Inc.

A Passive Ranging System, Joseph Star, Fairchild Guided Missiles Div.

AZUSA—Missile Tracking and Impact Prediction, L. G. Chase, Convair.

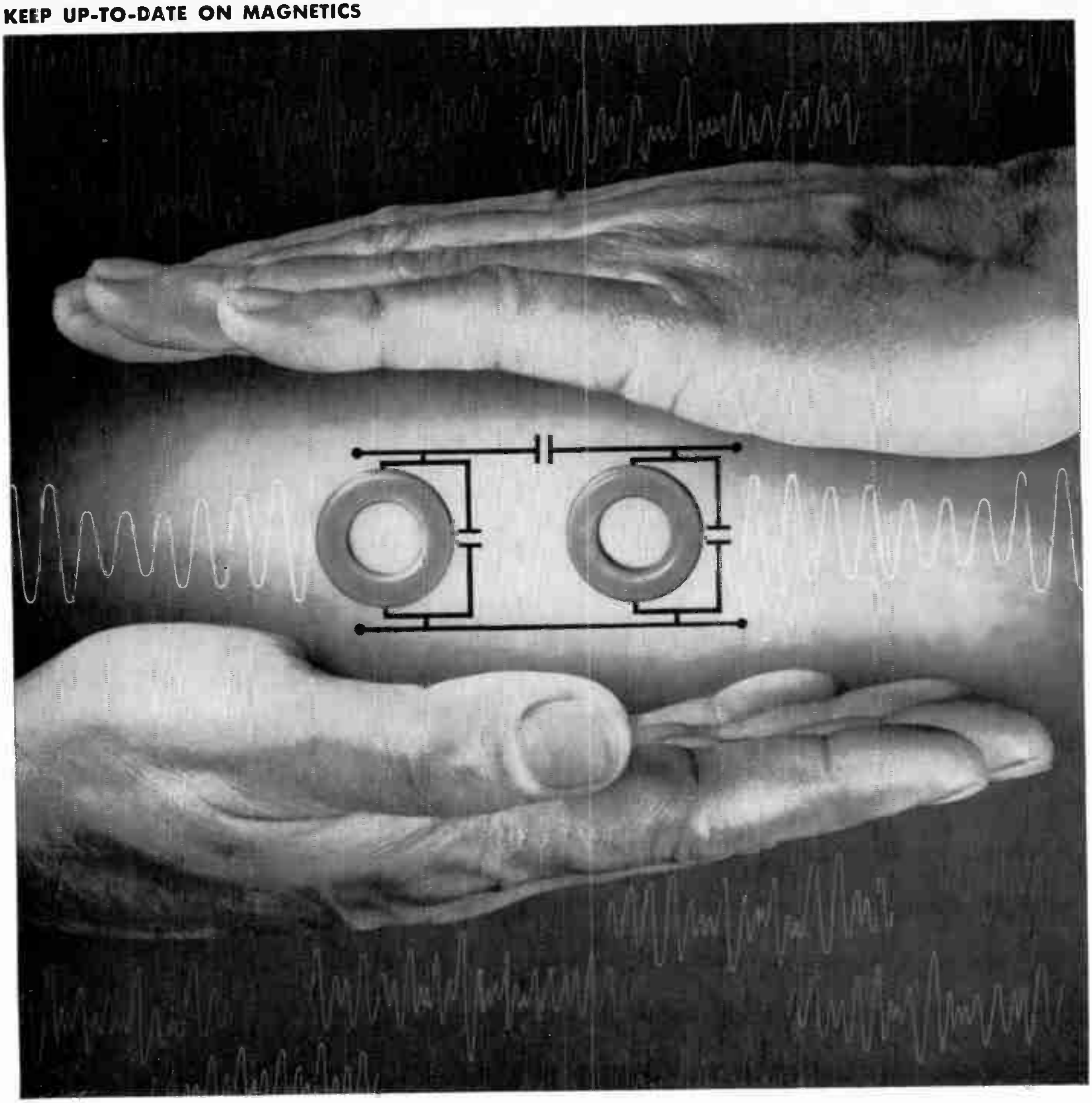
Automatic Drone Tracking System, R. O. Blackert, Motorola, Inc.

A Time-Duplexed Monopulse Receiver, E. J. Downey, R. H. Hardin, and J. Munushian, Hughes Aircraft Co.

Precision Logarithmic I. F. Amplifier, J. S. Brown, General Electric Co.

Redundancy Handling Systems with Spe-

Part of the executive committee of the Second National Convention on Military Electronics sits for its portrait. *Left to right*, they are: Jack Carter, Floyde Percy and R. H. Cranshaw of the arrangements committee; George Rappaport, public relations; W. B. Larew, convention president; B. J. Goldfarb, TRANSACTIONS; Donald McLaughlin, program; Morton Gale, secretary; and J. J. Wynne, arrangements. The Air Research & Development Command is sponsor of five confidential sessions.



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cial Application to the Extraction of Position-Information from Data Supplied from Three Jammed Radar Stations, Lars Lofgren, Swedish Research Institute of National Defense.

Reconnaissance and Infrared Systems

Moderator, to be announced.

Viewfinder, P. T. Kaestner, F. W. Fecker, Inc.

Reconnaissance Auxiliary Data Handling Systems, P. H. Beck, Aerial Reconnaissance Lab., WADC, Wright Patterson Air Force Base.

A Night Television System, C. T. Shelton, Radio Corporation of America.

Infrared Detectors, L. J. Neuringer, Raytheon Mfg. Co.

Aerial Reconnaissance Systems, Fairchild Camera and Instrument Corp.

ASTIRS, Donald Wahl, AVION Div., ACF Industries, Inc.

Reliability Statistics, Cost and Value Analysis

Moderator: C. M. Ryerson, Radio Corporation of America.

Six Million Measurements Made Useful, Craig Walsh and T. C. Tsao, The McGraw Hill Technical Writing Service.

Value Engineering Analysis and Reliability, R. M. Jacobs, Radio Corporation of America.

Reliability for Missile Computer Circuits, R. M. Schultz, Auerbach Electronics Corp.

Inertial Guidance System Reliability Program, E. F. Dertinger, ARMA Div., American Bosch Arma Corp.

Component Failure Statistics for the Planned Reliability of Miniaturized Airborne Digital Computers, E. L. Wong and W. C. Gray, Hughes Aircraft Co.

Determining Total Equipment Costs, A. M. Dame, Naval Aviation Electronics Service Unit, Naval Receiving Station.

Components for Military Electronics

Moderator, to be announced.

Microminiaturization of Internal Electronics—"Microelectronics," T. A. Prugh, Diamond Ordnance Fuze Labs.

Avalanche Noise in FN Junctions, S. Sherr and S. King, General Precision Lab., Inc.

Application Notes on a New UHF Ceramic Triode, M. W. Hamilton, General Electric Co

Considerations for the Reliable Use of Receiving Type Tubes in Class C Circuits, A. Dzik, Radio Corporation of America.

2:00–5:00 p.m.

Infrared Detection and Reconnaissance Systems (Confidential)

Sponsored by the Air Research and Development Command.

Moderator: D. S. Lowe, Naval Ordnance Lab.

IR Aerial Mapping and Telemetering with Various Detectors and Its Application to Drones, W. B. Birtley, J. P. Walker, Jr., and J. B. Cannon, Jr., Haller, Raymond and Brown, Inc.

Principles of Short Range Infrared Devices, R. M. Talley, U. S. Naval Ordnance Lab.

Airborne Target Detection by Ultra Violet Radiation, Harry Kihn and F. G. Durfee, RCA Labs.

Transistorization of the S-4 Infrared Reconnaissance System, A. D. Pirone, M. V. Joyce, and W. W. Chou, Perkin-Elmer Corp.

Traveling-Wave Tube Preamplification for Crystal-Video Intercept Systems, speaker to be announced.

Space Electronics—Guidance and Control

Moderator: A. S. Locke, Vitro Labs.

Interferometric Rocket Guidance, M. J. E. Golay, Philco Corp.

Considerations Influencing the Design of Autopilot Parameters for an Aerodynamically Unstable Missile, J. R. Jacques, Johns Hopkins Univ.

On the Actual Shape of the Earth as Applied to Problems in Satellite Dynamics, P. H. Savet, Technical Consultant, Rand, Arma Div., American Bosch Arma Corp.

Space Vehicle Communications System Design, D. E. Sukhia, The Martin Co.

A Rapid Method for Determination of Times and Paths over the Earth's Surface for Artificial Earth Satellites, R. A. Moore, Swedish Research Institute of National Defense.

Ranging and Tracking

Moderator, to be announced.

A Guidance Radar for Satellites, Space Vehicles and Ballistic Missiles, E. A. Mechler, J. W. Porter, and R. O. Yavne, RCA.

Digital Ranging System, A. J. Lisicky, Radio Corporation of America.

On the Theory of Light Radar, R. H. Meier, Perkin-Elmer Corp.

A Method for Tracking Cooperative Targets, M. G. Spooner, G. E. Richmond, and John Beyer, Cornell Aeronautical Lab., Inc., and T. R. Benedict, Univ. of Wisconsin.

A Double Delay and Subtraction Airborne Clutter Canceller, Karl Solomon, Radio Corporation of America.

Circuitry I

Moderator, to be announced.

A General Discussion of Switching Circuit Synthesis, G. D. Ward, Hughes Aircraft Co.

The Current-Switching Technique for Digital Computer Circuitry, C. M. Campbell, Jr., Burroughs Corp.

Some Design Considerations in the Application of Silicon Transistors to Voltage Mode Digital Circuitry, J. V. B. Cooper and W. K. Mead, IBM Corp.

A High-Speed Transistor Shift Register for Operation up to 135°C, J. L. Robinson, Philco Corp.

Wednesday, June 18

9:00–12 Noon

Data Handling Systems and Data Links (Confidential)

Sponsored by the Air Research and Development Command.

Moderator: Capt. P. Van Leunen, Office of the Chief of Naval Operations, Navy Department.

Data Processing Equipment for the Land Based Talos Defense Unit, G. W. Oberle, Radio Corporation of America.

AN/FST-2 Radar Processing Equipment for Sage, E. W. Veitch, Burrough Research Center.

Radar Recording and Data Handling Using the Ultrasonic Light Modulator, speaker from Fairchild Camera and Instrument Corp.

Airborne Data Link System, R. G. Gabrielson, Motorola, Inc.

A Quantitative Analysis on the Importance of Detection Range in Air Defense, Capt. J. A. Corvi, USMC, U. S. Naval Air Missile Test Center.

In-Flight Measurement of X-Band Attenuation and Phase Jitter on a Liquid-Fuel Rocket, R. L. Scrafford, Motorola, Inc., and Robertson Stevens, Jet Propulsion Lab.

Communications Systems

Moderator: D. C. Trafton, Directorate of Communications, Headquarters, USAF, Pentagon.

Phase Multilock Communication, C. A. Crafts, Robertshaw-Fulton Controls Co.

Graphic Communications, L. M. Seeberger, Radio Corporation of America.

Meteor Burst Propagation Techniques as an Aid to Naval Communications, H. E. Chubb, U. S. Navy Electronics Lab.

Amplitude and Phase Distortion in FM Communication and Tracking Systems Due to Multipath Interferences, H. D. Becker, Cornell Aeronautical Lab., Inc.

Communication by Vibratory Tactile Stimuli, Joseph Hirsch, Joseph Hirsch and Associates.

Inertial Systems

Moderator: O. H. Schuck, Minneapolis-Honeywell Reg. Co.

Problems Involved in Extending Inertial Navigation to Requirements of the Future, C. J. Mundo, Missile Guidance Arma Division. American Bosch Arma Corp.

GYSTATS, M. J. E. Golay, Philco Corp.

ABLE—A Portable, High Accuracy Gyro-Compass, M. E. Campbell, Autonetics.

Inertial System Alignment Study, J. E. Vaeth, The Martin Co.

Circuitry II

Moderator, to be announced.

Design of Voltage-Controlled Audio Oscillators, C. S. Schwarts, Gen. Precision Lab.

A Proposed Differential Pre-Amplifier with Improved Rejection of Common Mode Input, H. C. Talmadge, Jr., U. S. Naval Research Lab.

A Voltage-Sensitive Switch—Part I. Preparation and Properties, K. O. Otley and R. F. Shoemaker, Diamond Ordnance Fuze Lab.

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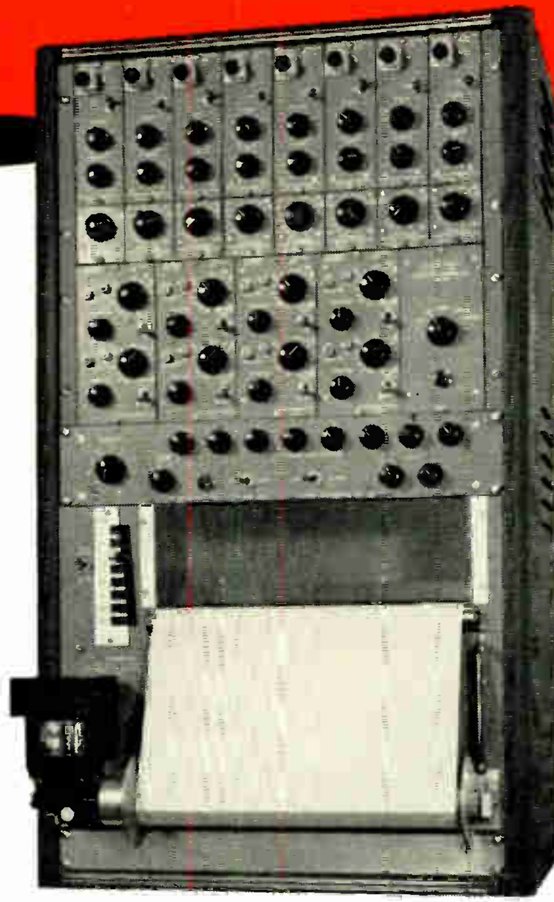

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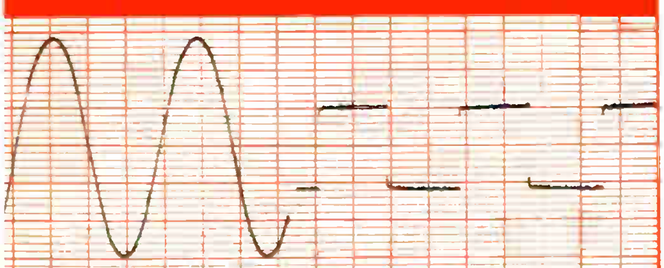

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A Voltage-Sensitive Switch—Part II. Circuits for the Switch, R. F. Shoemaker and P. J. Franklin, Diamond Ordnance Fuze Lab.

2:00–5:00 p.m.

Guidance and Control (Confidential)

Sponsored by the Air Research and Development Command.

Moderator, to be announced.

Discussion and Applications of Adaptive Control Systems, Capt. R. R. Rath, Flight Control Lab., WADC, Wright Patterson Air Force Base.

Terminal Maneuver of a Ballistic Missile, Elliott Cutting, Jet Propulsion Lab., California Institute of Technology.

Digital Mechanization of a Simple Jigsaw Puzzle Game and its Application to Long-Range Guidance, Edwin Shetland, Johns Hopkins Lab.

A Method of Evaluating a Missile Guidance System, W. Hammond and A. Ryder, The Martin Co.

Beam-Rider Guidance, C. F. Chubb, Sperry Gyroscope Co.

TALOS Homing System, H. L. Beazell, Jr., Bendix Aviation Corp.

Missile Environment and Packaging

Moderator, to be announced.

The Optimum Electronic Module Size for Airborne Digital Computers—Large?, Medium?, Small?, J. J. Staller, Arma Div., American Bosch Arma Corp.

Radiant Heating from Rocket Exhausts, D. C. Rohlfs and W. W. Balwanz, U. S. Naval Research Lab.

The Determination of Missile Vibration Environment, M. R. Beekman, U. S. Naval Air Missile Test Center.

A 5 KW Evaporative Cooled Power Amplifier, D. W. Smith, The Martin Co.

Integrated Electronic Systems Using the Modular Design Concept, A. H. Wulfsberg, Collins Radio Co.

Simulation Systems

Moderator: J. I. Leskinen, Naval Training Devices Center.

A Simulator for Accurate Testing of a Doppler Radar System, S. K. Benjamin, S. King, P. M. Levy, and S. Sherr, General Precision Lab., Inc.

Radar Guidance Simulator, D. F. Nieman and G. H. Miller, Univ. of Virginia,

Command Guidance and Data Handling System Simulator for Closed Loop Checkout of PPM-AM Systems in the Microwave Region, L. R. Spillane, Sperry Gyroscope Co.

Radar Land Mass Simulation, S. H. Gross, Fairchild Guided Missiles Division.

Navigational Systems

Moderator: A. W. Coven, U. S. Naval Research Lab.

The Radux-Omega Long-Range Navigation System, M. L. Tibbals and C. J. Casselman, U. S. Navy Electronics Lab.

Military Utilization of a Common ATC System, B. H. Baldridge and E. V. Hogan, General Electric Co.

The RADAN Frequency Tracker—Part I, W. B. Lurie and C. S. Schwarts, General Precision Lab.

The AN/APN-96 Frequency Tracker—Part II, W. A. Seism and W. B. Lurie, General Precision Lab.

Professional Groups†

Aeronautical & Navigational Electronics—Joseph General, 6019 Highgate Dr., Baltimore 15, Md.

Antennas & Propagation—J. I. Bohnert, Code 5250, Naval Research Lab., Washington 25, D. C.

Audio—Dr. H. F. Olson, RCA Labs., Princeton, N. J.

Automatic Control—E. M. Grabbe, Ramo-Wooldridge Corp., Box 45067, Airport Station, Los Angeles 45, Calif.

Broadcast & Television Receivers—H. A. Bass, Avco Mfg. Corp., Arlington, Cincinnati, Ohio.

Broadcast Transmission Systems—C. H. Owen, 7 W. 66th St., N. Y. 23, N. Y.

Circuit Theory—W. H. Huggins, 2813 St. Paul St., Baltimore 18, Md.

Communications Systems—J. W. Worthington, Jr., Dawn Dr., Mounted Route, Rome, N. Y.

Component Parts—R. M. Soria, American Phenolic Corp., 1830 S. 54 Ave., Chicago 50, Ill.

† Names listed are Group Chairmen.

Education—J. D. Ryder, Dean of Engineering, Mich. State Univ., E. Lansing, Mich.

Electron Devices—T. M. Liimatainen, 5415 Connecticut Ave., N.W., Washington, D. C.

Electronic Computers—Werner Buchholz, IBM Product Dev., Poughkeepsie, N. Y.

Engineering Management—C. R. Burrows, Ford Instrument Co., 31-10 Thomson Ave., Long Island City 1, N. Y.

Engineering Writing and Speech—D. J. McNamara, Sperry Gyroscope Co., Great Neck, L. I., N. Y.

Human Factors in Electronics—H. P. Birmingham, U. S. Naval Research Lab., Washington 25, D. C. (acting)

Industrial Electronics—W. R. Thurston, General Radio Co., 275 Massachusetts Ave., Cambridge 39, Mass.

Information Theory—W. B. Davenport, Jr., Lincoln Lab., M.I.T., Lexington 73, Mass.

Instrumentation—F. C. Smith, Jr., Southwestern Industrial Electronics Co., 2831 Post Oak Rd., Houston 19, Tex.

Medical Electronics—L. B. Lusted, M.D., Univ. of Rochester School of Med., 260 Crittenden Blvd., Rochester 20, N. Y.

Microwave Theory and Techniques—W. L. Pritchard, Raytheon Mfg. Co., Newton, Mass.

Military Electronics—W. E. Cleaves, 3807 Fenchurch Rd., Baltimore 18, Md.

Nuclear Science—J. N. Grace, Westinghouse Atomic Power Div., Pittsburgh 34, Pa.

Production Techniques—E. R. Gamson, Telemeter Magnetics Inc., 2245 Pontius Ave., Los Angeles 64, Calif.

Radio Frequency Interference—H. R. Schwenk, Sperry Gyroscope Co., Great Neck, L. I., N. Y.

Reliability and Quality Control—Victor Wouk, Beta Electric Corp., 333 E. 103rd St., New York 29, N. Y.

Telemetry and Remote Control—C. H. Doersam, Jr., 24 Winthrop Rd., Port Washington, L. I., N. Y.

Ultrasonics Engineering—C. M. Harris, Columbia Univ., 632 W. 125 St., N. Y. 27, N. Y.

Vehicular Communications—C. M. Heiden, General Electric Co., Syracuse, N. Y.

Sections*

Akron (4)—H. F. Lanier, 2220—27th St., Cuyahoga Falls, Ohio; Charles Morrill, 2248—16th St., Cuyahoga Falls, Ohio.

Alamogordo-Holloman (6)—Lt. Col. D. H. Vlcek, 2106 Castle Dr., Holloman AFB, N. M.; Taft Nicholson, 2304 Willow Dr., Alamogordo, N. M.

* Numerals in parentheses following section designate region number. First name designates Chairman, second name, Secretary.

Albuquerque-Los Alamos (7)—B. L. Basore, 2405 Parsifal, N.E., Albuquerque, N. Mex.; John McLay, 3369—48 Loop, Sandia Base, Albuquerque, N. Mex.

Atlanta (3)—E. G. Holmes, 309 Pharr Rd., N.E., Rm. 9, Atlanta, Ga.; R. L. Ellis, Jr., 77 Karland Dr. N.W., Atlanta, Ga.

Baltimore (3)—M. R. Briggs, Westinghouse Elec. Corp., Box 746, Baltimore 3, Md.; B. Wolfe, 5628 Stonington Ave., Baltimore 7, Md.

Bay of Quinte (8)—W. D. Ryan, Cavalry House, Royal Military College, Kingston,

(Cont'd

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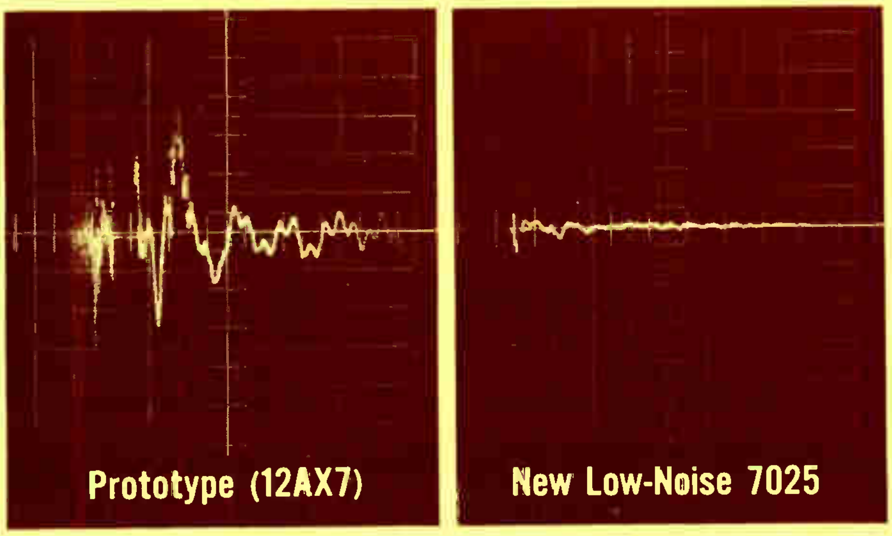
FROM THE RECEIVING TUBE DEPARTMENT OF GENERAL ELECTRIC COMPANY

General Electric Low-Noise 7025 AF-Amplifier Tube Major Step Toward Improved Hi-Fi Reproduction!



Scope Trace at Right Shows Superiority of New 7025 Twin Triode

You can see by comparison the greatly reduced noise output of the new General Electric amplifier tube. A single, identical tap was applied externally to a 12AX7 and to a 7025, both representative tubes from current production. Vertical measurement is plate voltage . . . horizontal measurement is time. Conditions: E_b: 250 v, R_l: 10 K, E_c: −2.5 v.

Modern sound-reproduction techniques put a premium on low background noise. The richness of today's high-fidelity tone calls for circuitry and tubes that reduce hum, microphonics, and other noise to a level approaching silence.

General Electric, long a pioneer in audio research—originator of the famous variable-reluctance cartridge and other basic aids to sound reproduction—now assists circuit designers with an outstanding low-noise amplifier tube, the 7025. This new twin triode promotes hum-free, noise-free reproduction of both disk and tape sound recordings.

In equipment now being designed or in production, the 7025 will directly replace Type 12AX7.

New Snubber Mica Holds Cathode Tight. Special Low-Hum Heater Employed.

The new 7025 features a spring snubber mica applied to the top of the cathode, which exerts a damping effect on any movement of the cathode caused by shock or vibration. This cuts microphonics substantially.

Also, a new tube heater of special design reduces hum by virtually eliminating heater magnetic influences on plate current and consequent hum in the plate circuit.

High-precision General Electric manufacture has been called on to achieve extremely close fits of all tube parts—a third, important factor in low-noise performance.

For best audio, apply the new General Electric 7025 AF-amplifier tube! Complete information about this low-noise twin triode is available from any G-E Receiving Tube office listed on the following page.

Military Equipment Builder Finds G-E 7077 Ceramic Triodes Have Mean Noise Figure Below 5 db!

Using a high-performance test circuit of advanced design, the research laboratory of a large manufacturer of military equipment has found that a sample lot of G-E 7077 RF-amplifier ceramic triodes show the mean noise figure of 4.6 db at 16 db gain. Tubes were operated at 500 megacycles.

The new 7077 is rated at 5.5 db noise at 14.5 db gain, 450 megacycles under power-matched conditions. Therefore, the test performance underscores the tube's suitability for military use, where low noise and high gain are vital.

Intended primarily for communications, radar, and navigation equipment, the new 7077 is a high-mu triode of planar construction. Altitude rating is 100,000 feet. It is economical in price, dependable, and rugged.

Ceramic construction gives the 7077 exceptional heat resistance. The tube is expected to be useful up to 300 C. It is designed for optimum mounting in grounded-grid UHF amplifier circuits. Size is extremely small —less than ½ inch long and wide.

Orders are being accepted now for delivery this year. See page that follows for average characteristics and typical-operation data.

Tear off and keep this sheet for reference. It contains useful tube-application data.

GENERAL ELECTRIC 7077 RF-AMPLIFIER CERAMIC TRIODE

AVERAGE CHARACTERISTICS

Plate Supply Voltage	250	Volts
Resistor in plate circuit (by-passed)	18000	Ohms
Cathode-Bias Resistor	82	Ohms
Amplification Factor	80	
Plate Resistance, approximate	8900	Ohms
Transconductance	9000	Micromhos
Plate Current	6.4	Milliamperes
Grid Voltage, approximate $G_m = 50$ Micromhos	−5	Volts

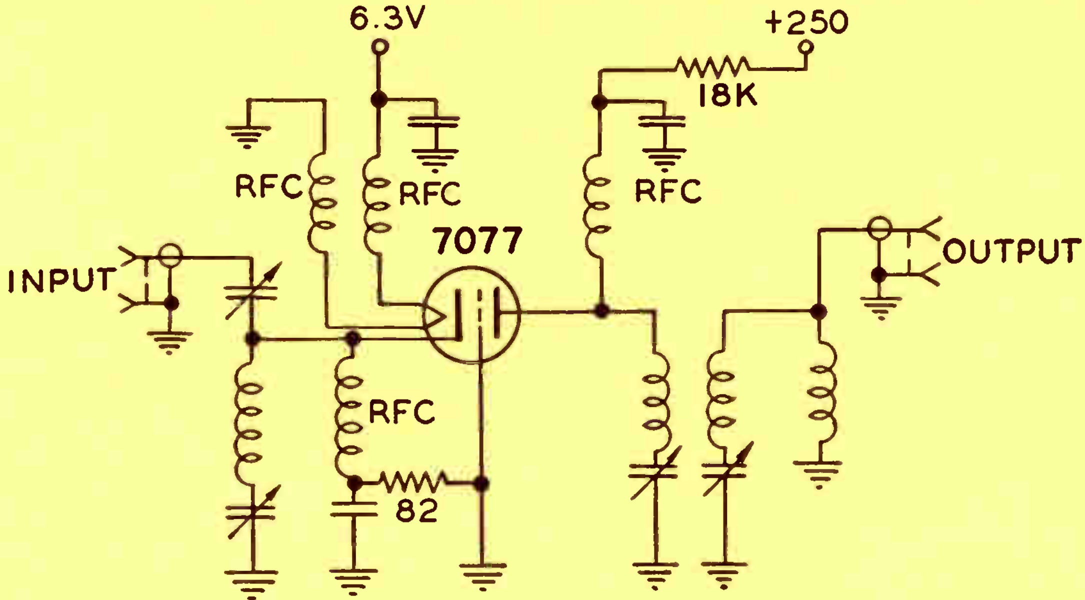
TYPICAL OPERATION

GROUNDED-GRID AMPLIFIER—450 MEGACYCLES

Plate Supply Voltage‡	250	Volts
Resistor in plate circuit (by-passed)‡	18000	Ohms
Cathode-Bias Resistor	82	Ohms
Plate Current	6.4	Milliamperes
Bandwidth, approximate	7	Megacycles
Power Gain, approximate	14.5	Decibels
Noise Figure (Measured with power-matched input, using argon lamp noise source), approximate	5.5	Decibels

‡ Lower supply voltage and a lower value of resistor may be used in the plate circuit with some sacrifice in uniformity of performance.

TYPICAL GROUNDED-GRID AMPLIFIER CIRCUIT USING THE 7077



Disclosure of the foregoing examples of the tube applications does not convey to purchasers of tubes any patent license, nor is it to be construed as recommending the use of such tubes in the infringement of patent claims.

For further information, phone nearest office of the G-E Receiving Tube Department below:

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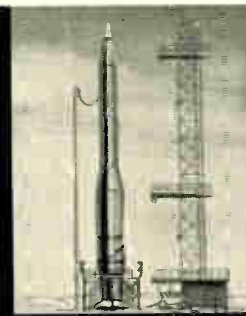

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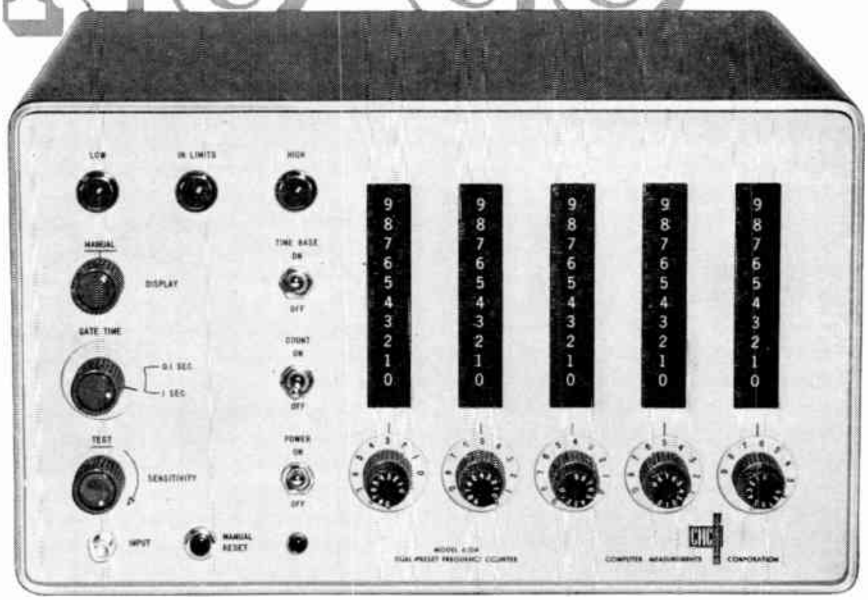
GO-NO-GO



For production testing and process control

CMC *announces the Model 620A*

ELECTRONIC GO-NO-GO GAUGE

Here's an interesting new instrument from CMC with a host of applications in production testing and process control.

The Electronic Go-No-Go Gauge monitors any control or limiting situation which can be stated in terms of frequency. For instance, in the electronics industry relatively unskilled workers can tune oscillator circuits, filter networks, etc. with great accuracy. Frequency stability and comparison checks can be made quickly and easily. In mills and factories producing a continuous flow of goods such as steel, rubber, paper, the device can be used as a material flow controller keeping the output in tune with the input, preventing line buckle and stretch-out. In chemical and petroleum processing, the Model 620A can serve as a pressure or liquid flow regulating indicating system. Wherever motor speed control is a problem, the Model 620A will hold the speed within preselected limits.

How it Works

In operation, the unknown frequency generated by either the unit under test or one of the many types of transducers on the market is applied at the input. Upper and lower frequency limits are selected by setting the control knobs on the front panel. If the unknown frequency falls below the lower limit, a red "low" lamp lights. Equal to or above the higher limit, a red "high" lamp lights. Within either limit, a green "in limits" lamp lights. Relay contact closure for external control occurs at each lamp condition.

Actual input frequency is displayed on decades. Remote visual monitoring can be obtained with CMC's new Inline-Inplane *Readable* Readout. Use of CMC's new fast printer provides a permanent printed record.

Like all CMC instruments, the new Model 620A features unitized construction for structural strength and low weight.

KEY SPECIFICATIONS

FREQUENCY RANGE	1 – 40,000 cps
INPUT REQUIREMENTS	.05v rms: 10 – 40,000 cps .07v rms: 1 – 10 cps Positive Pulse Rise Time: ½ volt or more/sec.
ACCURACY	±1 count ± stability
STABILITY	0.1% (Normal power line stability) Crystal time base optional.
TIME BASES	0.1 sec. and 1 sec. (10 sec. optional)
READOUT	4 digits (5 digits optional)
DISPLAY TIME	Automatic: Continuously variable 0.1 to 10 seconds. Manual: Until reset
INPUT IMPEDANCE	0.5 megohm and 50 mmfd
PRICE	$1120.00 f.o.b. factory

For more information, contact your nearest CMC engineering representative or write to us directly for complete specifications and applications data. Please address Department 435.

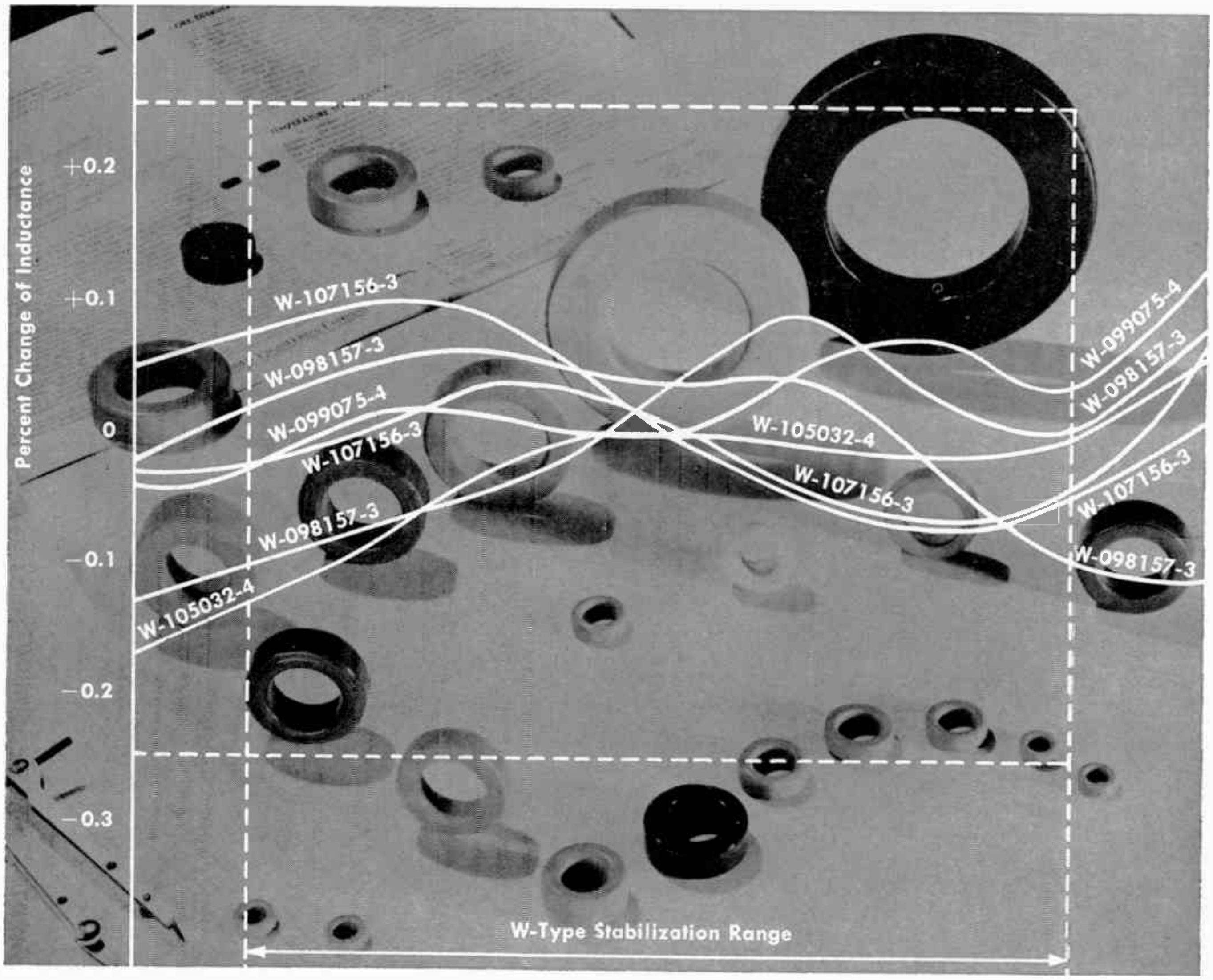
Subsidiary of Hancock Manufacturing Company

Computer-Measurements Corporation

5528 Vineland Avenue • No. Hollywood, California

Phone: STanley 7-0401 • TWX: N HOL 2160

2



ARNOLD offers you the <u>widest</u> selection of Temperature Stabilized MO-PERMALLOY POWDER CORES

Arnold Molybdenum Permalloy powder cores are available with the temperature coefficient of inductance controlled within certain limits over specific temperature ranges. Most core sizes and permeability combinations can be supplied in at least one of the four different types of temperature stabilization available.

For example, most of the popular core sizes are manufactured in the new type of wide range—"W"—stabilized cores whose temperature coefficient of inductance does not exceed 0.5% over the temperature range covered by the MIL-T-27 specification of −55° C to +85° C.

This type of guaranteed maximum change of inductance with temperature, as well as the constancy of permeability with time and flux level, are of particular importance to apparatus and circuit engineers. Many precision military and industrial applications demand the uniform performance and the excellent physical properties found only in Arnold Mo-Permalloy powder cores.

For design flexibility they are furnished in a full range of sizes, up to 5.218" O.D., in four standard permeabilities: 125, 60, 26 and 14. You will find them dependable and easy to use. You will find most sizes and types in stock *now* for immediate shipment.

● Let us furnish your requirements for temperature stabilized Mo-Permalloy powder cores, or any magnetic materials you need, from the most complete line in the industry.

WSW 6999

For more information write for Bulletin PC-104B

Lists complete line of Mo-Permalloy Powder cores . . . available in 25 sizes from 0.260" O.D. to 5.218" O.D. Furnished also with various types of temperature stability from Type "A" unstabilized to Type "W" stabilized over the temperature range of —65° F to +185° F.

ADDRESS DEPT. P-85

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T/I-transistorized 'peeping drones'

see better fly farther

Transistorized radar... and other Texas Instruments "electronic eyes" can peg the shape, location, motion, heat, and magnetic character of "targets of opportunity" ... *relaying this vital data for action in those brief moments that the opportunity exists!* In manned or unmanned reconnaissance aircraft, TI's light, tough and compact electronics save fuel, space and weight while trimming maintenance and logistic problems.

Discussion of this advanced reconnaissance capability can be arranged on short notice. Authorized industrial or military personnel write or wire: Service Engineering Department ...

apparatus division

systems management— reconnaissance, airways control, anti-submarine warfare, anti-missile, countermeasures, airborne early warning, navigation, attack control, missile systems, engine control.

equipments— radar, infrared, sonar, magnetic detection, computers, timers, telemetering, intercom, microwave, optics, detector cells, engine instruments, transformers, time standards, and other precision devices.

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NEWS
New Products

Ultrasonic Analyzer

Probescope Company, Inc., 8 Sagamore Hill Dr., Port Washington, L.I., N.Y., announces a new wideband spectrum analyzer, model SS-500, covering the 75 cps to 600 frequency range.

The model SS-500 will give an instantaneous fourier analysis of high speed vibration, noise, pulse, and harmonics.

This model is suited to the design of jet engines, telemetering, microphonic studies, crystal characteristics and wind turbulence test.

Features include automatic optimum resolution, continuously variable sweep width and center frequency, input overload protection, tube failure indicators, internal frequency calibration and a flat face cathode ray tube with camera mount bezel.

Specifications: (center frequency) to 500 mc; sweep width 2 kc to 200 kc; Resolution 150 cps to 2 kc; full scale sensitivity, 250 microvolts to 250 volts; linear and two decade log voltage scale, and a 60 db, dynamic range.

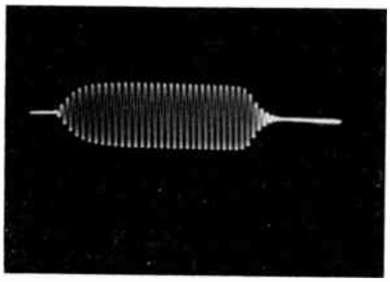
High Power Pulsed Oscillator

The PG-650 oscillator, a variable frequency pulse modulated rf source for application requiring high power output as well as extreme stability is announced by **Arenberg Ultrasonic Laboratory, Inc.,** 94 Green St., Jamaica Plain 30, Mass. Its principal use has been in measuring the various parameters of ultrasonic delay lines

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

whose high initial insertion loss as well as operation at low impedance levels have presented difficulties. The wide band widths covered by delay lines have also meant that the oscillator would have to cover a wide range of frequencies.

With this unit the rf output may be displayed directly on the plates of an oscilloscope, and the output of a delay line (60 db into 50 ohms) can also be shown at rf using only the vertical amplifier of the CRO and no other. A second item is that it has been possible to cascade two delay lines without an intermediate amplifier.

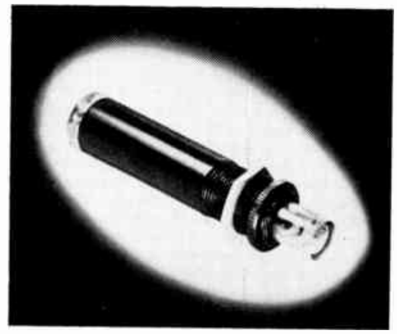


The circuits are stable to the extent that it has been possible to take photographs with 80 second exposure of a 40 mc pulsed sine wave 2 μsec long showing each cycle. The sharp rise and fall time of ½ μsec as well as minimum pulse length of less than 2 μsec has been the best achieved consistent with the high power level.

Additional features that make this self-contained instrument a versatile addition to a well-equipped laboratory, with many uses besides testing delay lines, and in fact outside of ultrasonics, are (1) the incorporation of a trigger generator for internal use which can lock on 60 cycles; (2) a trigger amplifier for synchronizing with an external source; (3) a variable trigger delay circuit for viewing the leading edge of pulses on a CRO display, while adjusting to coincide with suitable time marks; and (4) a continuously variable pulse length control.

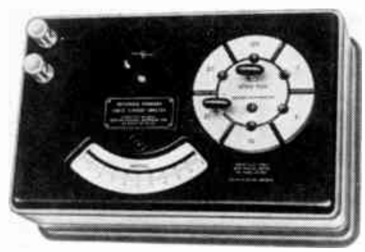
Computer Control Panel Components

A new line of special control panel components has been developed by **Transistor Electronics Co.,** 3357 Republic Ave., Minneapolis 26, Minn. The Echo-Lite is a push button with a NE-2 neon bulb enclosed, the Memo-Lite a subminiature thyratron indicator for transistor circuitry, and the Transistor-Lite a transistor con-



trolled neon lamp circuit. Each is housed in an anodized aluminum tube of ½ inch diameter, 2 inches long. The gold plated terminal connections are for AMP series "53" taper pins. The lenses are molded from clear plastic, but are available in colors as well. The terminal header is nylon, and mylar insulation is used internally. Provisions are made internally for as many as four-½ watt composition resistors.

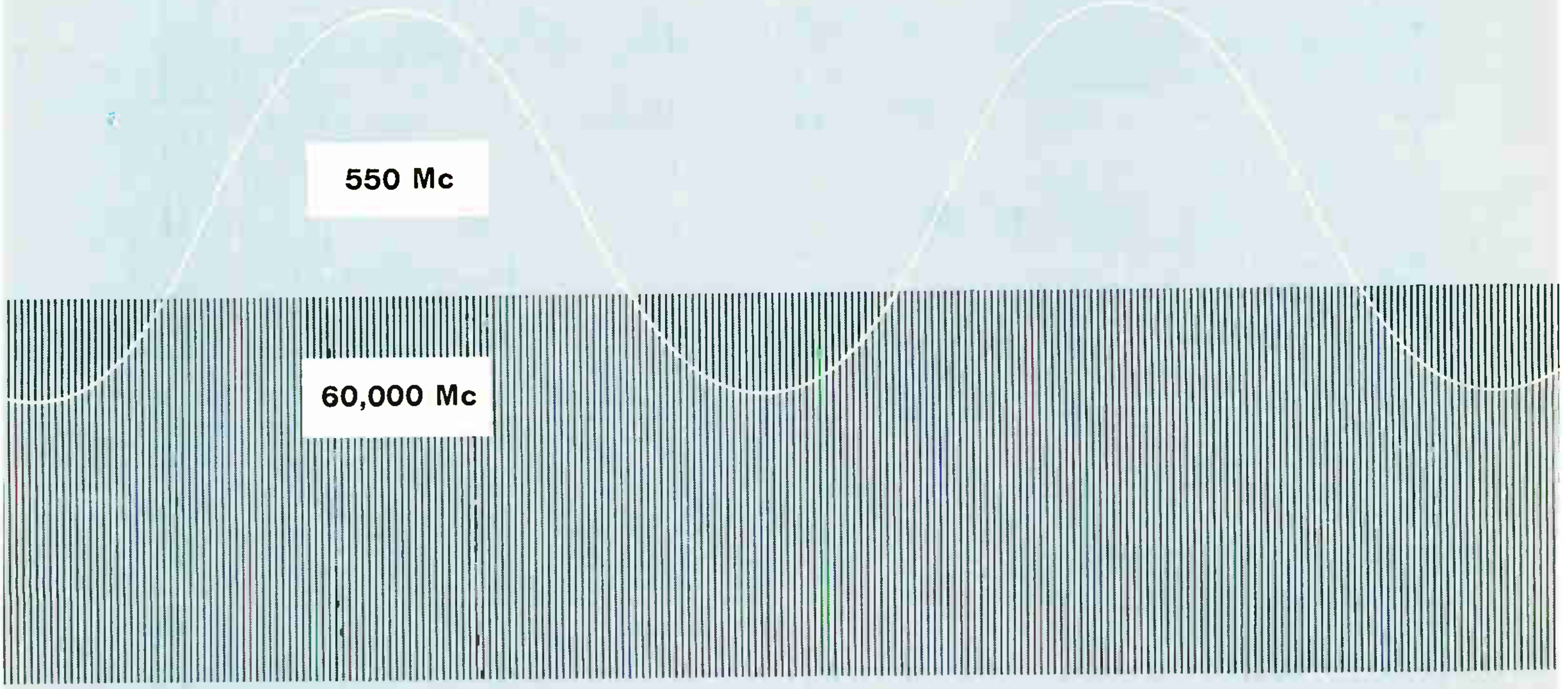
Direct Reading Heavy Current DC Ammeter



A new heavy current precision dc ammeter which is capable of accurate measurement of a wide range of currents at a moderate cost is announced by **Sensitive Research Instrument Corp.,** New Rochelle, N.Y. Self contained and portable, this instrument eliminates external shunts and leads. Ranges are selected by means of positive-fitting tapered plugs and holes. The movement features Diamond Pivots and shock mounted jewels for ruggedness and repeatability. Operation: Direct Currents. Accuracy: is 0.2 per cent. Ranges: are a choice of combination #1: 0–1/5/10/20/50/100 amperes; or combination #2: 0–1. 5/3/7.5/15/30/75 amperes. The scale is 6.3 inches, hand-drawn, and mirrored; Comb. #1: 100 divisions, Comb. #2: 150 divisions. The type has a double-diamond pivoted, permanent magnet. Sensitivity is 50 millivolts. Temperature coefficient is negligible between 20 and 30°C. Period is 1 second.

(Continued on page 130A)

Raytheon — World's Largest Manufacturer of Magnetrons and Klystrons



RAYTHEON REFLEX KLYSTRONS

from 550 to 60,000 Megacycles

More than 70 Raytheon reflex-type klystrons for local oscillator, signal generator and transmitter applications.

Raytheon produces more reflex klystrons than all other manufacturers in the world combined . . . one important reason why Raytheon klystrons have established a matchless record for reliability and proved performance in thousands of installations.

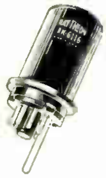
Equipment designers are welcome to call on our Application Engineer Service. Write for consolidated data booklet presenting comprehensive characteristics of the *complete* line of Raytheon klystrons, magnetrons and special tubes. There is no cost, or obligation.

3 TYPICAL RAYTHEON REFLEX KLYSTRONS

RK-5721 — Velocity variation oscillator designed for use with a coaxial cavity in CW or pulsed operation over the 4290 to 11,000 Mc range for signal generator and special local oscillator applications.

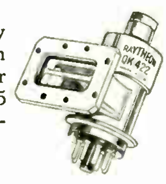
Heater Input @ 0.58 A	6.3 V
Reflector Voltage Transit Mode	2¾ cycles
Frequency Range	4290-8340 Mc
DC Resonator Input @ 20 mA	1000 Vdc
DC Reflector Voltage	-50 to -625 V
Electronic Tuning (Half Power) Frequency Change	12 Mc min.
Reflector Modulation Sensitivity (8340 Mc)	0.1 Mc/volt
Power Output (Average CW)	160 mW

RK-6116 — A ruggedized thermally tuned oscillator of the integral cavity type designed for CW operation in the 8500 to 9600 Mc range with an average power output of 30 mW.



Heater Input @ 0.52 A	6.3 V
Tuner Heater Current	0.80 A
Frequency Range	8500-9660 Mc
Resonator Input @ 25 mA	300 Vdc
Reflector Voltage (max. Po @ 8550 to 9660 Mc)	-60 to -145 Vdc
Thermal Tuning Time 8500-9660 Mc	2 seconds
Electronic Tuning Range @ 9080 Mc	100 Mc
Power Output 8500-9660 Mc	26 to 34 mW

QK-422—A mechanically tuned velocity variation oscillator designed for CW operation in the 7125 to 8125 Mc range in microwave relay systems.



Heater Input @ .44 A	6.3 V
Frequency Range	7125 to 8125 Mc
DC Resonator Input @ 32 mA	300 Vdc
DC Reflector Voltage (max. Po @ 7125 to 8125 Mc)	-130 to -210 Vdc
Power Output 7125 to 8125 Mc	100 mW min.
Electronic Tuning (to half power points) @ 7600 Mc	25 Mc min.
Modulation Sensitivity @ 7600 Mc (10 V pk. to pk. mod. volt.)	.5 Mc/V min.

RAYTHEON MANUFACTURING COMPANY

RAYTHEON

Excellence in Electronics

Microwave and Power Tube Division, Section PT-06, Waltham 54, Mass.

Regional Sales Offices: 9501 W. Grand Avenue, Franklin Park, Illinois. 5236 Santa Monica Blvd., Los Angeles 29, California

Raytheon makes: Magnetrons and Klystrons, Backward Wave Oscillators, Traveling Wave Tubes, Storage Tubes, Power Tubes, Miniature and Sub-Miniature Tubes, Semiconductor Products, Ceramics and Ceramic Assemblies

IRE People

The Federal Communications Commission named **R. J. Renton** (A'29–M'45), now Assistant Chief Engineer in charge of the Technical Research Division, to be Associate Chief Engineer. Mr. Renton has served the FCC and its predecessor, Federal Radio Commission, in various field and headquarters engineering capacities since 1929. He was a member of or delegate to various international engineering conferences and, for several years, chairman of the North American Regional Broadcasting Engineering Conference. From 1951 to 1956 he was United States Supervisor of the Conelrad program. Born in Boston in 1905, he received his B.E.E. degree from Northeastern University in 1927.

❖

A. G. Skrivseth (A'42–M'45–SM'48) was also named by the FCC to be Acting Chief of the Technical Research Division. Mr. Skrivseth was born in Thor, Iowa, in 1915. He was graduated from Montana State University in 1939 with a B.A. degree in physics and mathematics, and then was a graduate assistant at Cornell University until 1941. In the latter year he joined the FCC as an assistant radio inspector. Since then he has filled a variety of commission engineering posts, was a delegate to the International Radio Consultative Committee conference at Warsaw in 1956, and is author or co-author of several technical reports issued by the commission.

❖

J. F. Jenkins, Jr. (SM'54) has been appointed Director of Development at Interstate Electronics Corporation in Anaheim, Calif. He is responsible for the direction of major development projects at IEC, including all phases of development work performed under prime contract to the government for missile test and evaluation instrumentation. He came to IEC as Assistant Director of Development early in 1956 and in 1957 was appointed member of the Board of Directors.

J. F. Jenkins, Jr.

Prior to his association with Interstate Electronics Corp., Dr. Jenkins directed and participated in numerous advanced research and development projects. He was project engineer for design, development, and construction of a complete Askania phototheodolite range at Edwards AFB. He participated as a senior staff physicist in the establishment of a transistor study group to apply transistors to guided-missile systems. Other activities included prime responsibility in a project to obtain data on high-energy protons and pions at balloon altitudes, extensive research work in high-altitude physics, and research in vibration problems attending the use of electron tubes in guided missiles.

Dr. Jenkins received his Ph.D. degree in physics from the University of Maryland in 1954. He is a member of the American Physical Society, Washington Philosophical Society, Sigma Xi, Phi Kappa Phi, and Sigma Pi Sigma.

❖

Page Communications Engineers, Inc. recently announced the appointment of **Edwin Dyke** (S'41–M'46–SM'48) as an assistant chief engineer.

Edwin Dyke

Over the past twelve of his seventeen years in the field, Mr. Dyke has directed microwave-relay design and planning for systems in Canada, Alaska, South America, Europe, Australia, and Asia.

Mr. Dyke had been assistant director of engineering at Collins Radio Co. in Dallas, Tex., since 1954, and previously spent eight years with Motorola, Inc., Chicago. He contributed to the Navy's first microwave radar while with the Bureau of Ships from 1941–44, and later, during two years with Lear, Inc., Grand Rapids, Mich., he was a group leader for military automatic direction finder design.

He holds numerous patents for microwave relay, antennas, diversity, and automatic tuning; his design is the most widely installed 6000 mc/s microwave relay in this country. The CAA recently purchased Mr. Dyke's microwave relay design for installation in some 140 stations throughout the U. S., which involved an outlay of approximately $7 million for this equipment.

He is a graduate of the University of California at Berkeley, and of the University of Chicago School of Business Administration. His memberships include the Armed Forces Communications and Electronics Association, Petroleum Electrical Suppliers Association, and the Aircraft Owners and Pilots Association.

❖

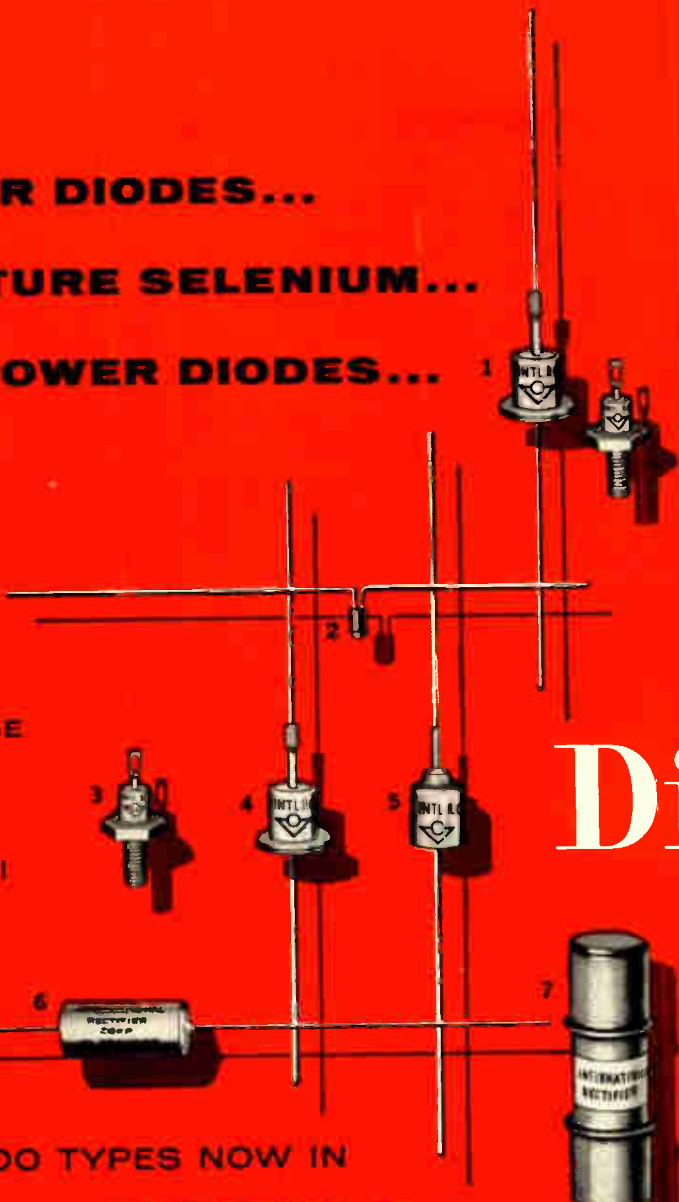
The R. W. Johnson Company recently announced an expansion of its consulting engineering services to industry and government, and a change of location to 9372 Hillview Rd., Anaheim, Calif. The firm specializes in test-range instrumentation, industrial electronics, new project planning and organization, and technical pro-

(Continued on page 48A)

ZENER DIODES...

SUB-MINIATURE SELENIUM...

SILICON POWER DIODES...



A COMPLETE RANGE FROM 50 TO 16,000 VOLTS PIV!

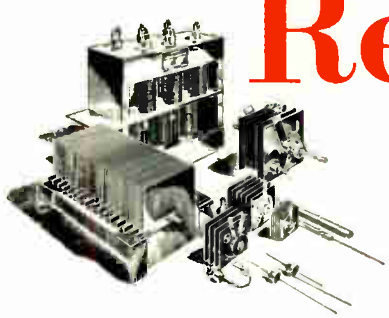
OVER 700 TYPES NOW IN FULL PRODUCTION AT

1. **Silicon Zener Diodes** Pigtail Type: 1.0 Watt Rated. Stud Mounted: 3.5 Watt Rated Series in each type. 3.9 volts to 30 v. in 10% voltage steps. Hermetically sealed. Bulletin SR-155.
2. **Selenium Sub-Miniature Diodes** High inverse resistance. Ideal for bias supplies computer matrices. Output voltages: 20 to 160 v.; output currents: 100 microamperes to 11 ma. Complete series. Bulletin SD-1B.
3. **Stud Mounted Silicon Power Diodes** Style T. For power supply and mag amp applications. PIV ratings from 50 to 600 v. at 800 ma dc output. All-welded, hermetically sealed. Bulletin SR-135C.
4. **Silicon Power Diodes—Pigtail** PIV range: 50 to 600 v. at 300 ma dc output current. For power supply and mag amp application. All-welded, hermetically sealed. Complete series. Bulletin SR-132E.
5. **High Voltage Silicon Power Diodes** Style J. Hermetically sealed. A complete series with PIV ratings from 600 to 1000 v. at 125 ma dc output current. Bulletin SR-136E.

Diodes

6. **High Voltage Silicon Power Diodes** Style K. Hermetically sealed. PIV from 600 to 1200 v. at 100 ma dc output current. Bulletin SR-144A.
7. **High Voltage Cartridge Types** For tough miniaturization problems! A series with ratings from 1000 volts PIV at 100 ma to 16,000 volts PIV at 45 mm (Half-Wave). Hermetically sealed. Bulletin SR-139B.

International Rectifier Corp.



The diodes listed are typical of the wide selection available at International Rectifier to solve your rectification problem...*with excellent reliability!* Your letterhead inquiry will bring the bulletin you specify and—if you include the details of your project—a recommendation stating the diode best suited to your need. The illustration at left suggests the scope of our complete line of selenium, germanium and silicon rectifiers for all dc needs from microwatts to megawatts, literally the widest range in the industry.

EXECUTIVE OFFICES: EL SEGUNDO, CALIFORNIA • PHONE OREGON 8-6281 • CABLE RECTUSA • NEW YORK AREA OFFICE: 132 EAST 70TH ST., PHONE TRAFALGAR 9-3330
CHICAGO AREA OFFICE: 205 W. WACKER DR., PHONE FRANKLIN 2-3888 • NEW ENGLAND AREA OFFICE: 17 DUNSTER ST., CAMBRIDGE, MASS., PHONE UNIVERSITY 4-6520

WORLD'S LARGEST SUPPLIER OF INDUSTRIAL METALLIC RECTIFIERS • SELENIUM • GERMANIUM • SILICON

IRE People

(Continued from page 46A)

posals. **R. W. Johnson** (S'42–A'45–M'45–SM'49), Chief Engineer of the firm, is known in the test-range instrumentation and industrial electronics fields, and until recently was Director of Development at Interstate Electronics Corp. in Anaheim. Mr. Johnson is a Registered Professional Engineer, and is the author of many papers on management and technical subjects.

❖

H. E. Hockeimer (M'52) has been promoted to manager of the Field Engineering department of Philco Corporation's Government and Industrial Division.

H. E. Hockeimer

Mr. Hockeimer has been with Philco since 1947, and has served in various field and headquarters assignments. He joined the G and I Division in 1951 as a project engineer on Philco's early microwave installations and has been assistant manager of field engineering since 1955.

A native of Winzig, Germany, Mr. Hockeimer attended high school in Breslau and studied electronics at the Breslau Technical Institute. He also studied industrial management and electronics at New York University after completing an electronics course at the RCA Institute.

He worked for the American Military Government in Germany from 1945 to 1946 and served as a communications instructor with the New York National Guard from 1947 to 1951.

❖

Cascade Research Division, Monogram Precision Industries, Inc. announces the appointment of **A. J. Thompson** (M'55) as a research and development engineer.

A. J. Thompson

Before joining the Los Gatos, California, microwave component company, Mr. Thompson was a senior electronics research physicist at Eitel-McCullough, Inc., having been with Eimac since 1952.

A native of Pasco, Wash., Mr. Thompson majored in physics, graduated with a B.A. degree from the College of Pacific in 1952. Prior to this he spent two and one-half years in the U. S. Navy.

In his new post, Mr. Thompson will be working on traveling-wave tubes and backward-wave oscillators.

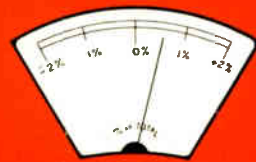
(Continued on page 50A)

YOU CAN DEPEND ON

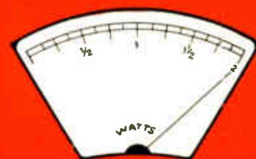
DALOHM ®

PRECISION DEPOSITED CARBON RESISTORS

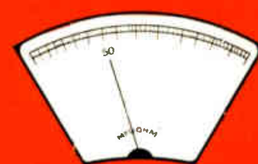
HI-TEMPERATURE
−55° C. to +150° C.



HI-PRECISION
±1%



HI-WATTAGE
¼ to 2 Watts



HI-RESISTANCE
1 Ohm to 50 Megohms

DCH Resistors take VIBRATION up to 2500 cps...yet retain 100% reliability!

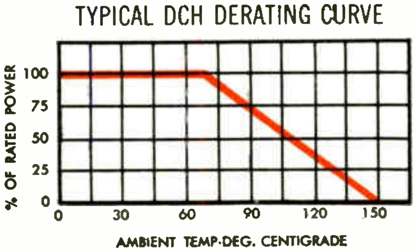
Severe vibration is only one of many tough parameters that DALOHM DCH Deposited Carbon Resistors meet with 100% reliability.

All DALOHM Deposited Carbon Resistors surpass the extremes of resistor requirements for their respective types, at the same time providing a wide margin in precision, miniaturization and reliability.

Look at these over-all parameters and see how DALOHM Deposited Carbon Resistors can help you meet your critical design problems.

- Precision tolerance: 1%
- Powered at ⅛, ¼, ½, 1 and 2 watts.
- Resistance range from 1 ohm to 50 megohms.
- Surpass requirements of MIL-R-10509B.
- Temperature coefficient: 140 ppm/degree C. to 500 ppm/degree C.
- Voltage coefficient: 0.002% or less per volt.

TYPICAL DCH DERATING CURVE



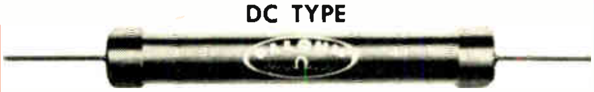
A TYPE FOR EVERY APPLICATION

DCH TYPE

Completely sealed in newly developed, non-hydroscopic ceramic; gives absolute protection from humidity, salt spray, shock, vibration and other adverse environmental conditions; 8 sizes from 9/32 x .155 to 2¼ x .400; power, resistance and tolerance as listed at left.

Request Bulletin R-27

DC TYPE



Silicone sealed providing maximum protection from abrasion, moisture, salt spray and other environmental conditions; 8 sizes from 9/32 x 3/32 to 2-1/16 x 5/16; power, resistance and tolerance as listed at left.

Request Bulletin R-24

MC TYPE

Latest development in deposited carbon; molded in plastic for complete protection from moisture; salt spray, shock, vibration and all environmental conditions; 3 sizes from ¾ x ¼ to 2¼ x ⅜; powered at ½, 1 and 2 watts; resistance and tolerance as listed at left.

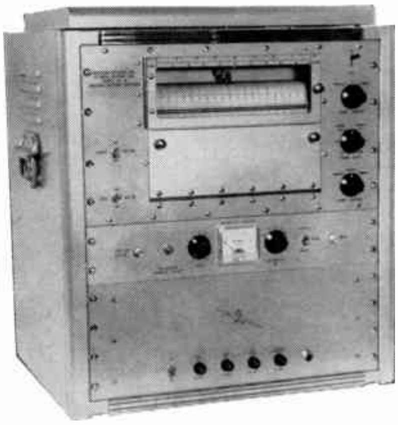
Request Bulletin MC

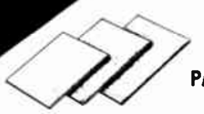
JUST ASK US

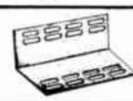
DALOHM line includes a complete selection of precision wire wound, power and precision deposited carbon resistors. Also trimmer potentiometers, precision wire wound and deposited carbon; and collet fitting knobs. Write for free catalog.

If none of DALOHM standard line meets your need, our engineering department is ready to help solve your problem in the realm of development, engineering, design and production. Just outline your specific situation.

DALE PRODUCTS, INC.
1302 28th Ave.
Columbus, Nebr., U.S.A.







IRE People

(Continued from page 48A)

G. L. Haller (A'28–M'36–SM'43–F'50) 51, has been elected a Vice-President of the General Electric Company, it was announced recently.

G. L. HALLER

Dr. Haller is General Manager of General Electric's Defense Electronics Division, a position he has held since 1956.

Dr. Haller was born in Pittsburgh, Pa., on May 8, 1907, graduated from Mercersburg Academy in 1924 and received the following degrees from Pennsylvania State University: B.S. in E.E. 1927; E. E. 1934; M.S. in Physics 1935, and Ph.D. in Physics 1942.

He was a radio engineer with Westinghouse Electric and Manufacturing Company from 1927 to 1929 and an audio engineer with E. A. Myers and Sons in Pittsburgh from 1929 to 1933, before returning to Penn State as a graduate assistant.

He remained at the university until 1935 when he became a radio engineer for the War Department at Wright Field where he served until 1942.

From 1942 to 1946 he was on active duty in the Signal Corps, later the Air Force, attaining the rank of Colonel in 1945. He received the Legion of Merit for the development of the Air Force radar countermeasures program.

In 1946 he became assistant dean of the College of Chemistry and Physics at Pennsylvania State University and a year later was appointed dean. He held that position until 1954.

During the period 1946 to 1954 he also helped found, and served on the Board of Directors of the consulting firm of Haller, Raymond and Brown, Inc.

In 1954, after serving as a part-time consultant to the General Electric Company for several years, he joined the company as manager of its Electronic Laboratories, with headquarters at Syracuse, N. Y.

He was appointed General Manager of the Company's Defense Electronics Division in 1956.

Dr. Haller is a member of the Army Electronic Proving Ground Advisory Council, the Technical Advisory Panel on Electronics for the Assistant Secretary of Defense, and Chairman of the Electronics Division of the American Ordnance Association.

Dr. Haller also is a fellow in the American Physical Society, an associate fellow in the Institute of Aeronautical Sciences, a member of the American Institute of Electrical Engineers, the American Society for Engineering Education, and the Franklin Institute, and a life member of the Air Force Association.

He is a Registered Professional Engineer in the State of New York.

(Continued on page 53A)

How US SEMCOR assembles a Silicon Diode with ALL STAINLESS STEEL FUSITE TERMINAL

1. Terminal as supplied by Fusite. V-24 glass and all stainless steel.

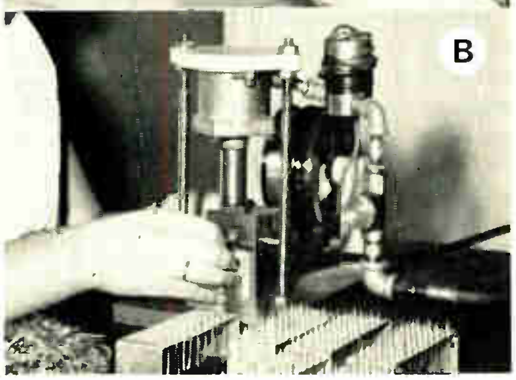
2. Inside end of straight wire electrode is crimped to a hollow open top shell. (See photo A)

3. Fraction of an inch of gold wire and aluminum wire are inserted and crimps folded over tight. (See photo B)

4. Precision bit of silicon crystal is fused with heat to ends of wire inserts.

5. Assembly is enclosed in miniature can and projection welded closed.

A



B

This application is typical of the hundreds of electronic components whose continuing operation is assured by the safety factor of a terminal with electrodes fused into the glass. The resulting rigidly fixed position of the electrode guards against damage to the finished assembly through movement of an electrode depending only on compression for its position in the glass.

Robert A. Rutherford, Vice President of U. S. Semiconductor Products, permits us this direct quotation.

"The Fusite header provides us with a very satisfactory solution to the problem of the lead wires turning in the header. The fused glass to steel has solved this very troublesome problem. The stainless steel material also provides excellent corrosion resistance. Aside from receiving a superior product from Fusite, we have also received very excellent service and a great deal of cooperation from both the company and their representative."

Test samples of any style terminal available on request. Stainless steel available on most Fusite Standard Headers.

Write Dept. G-2

INSULATE ELECTRICALLY · GLASS TO METAL · FUSITE TERMINALS · PROTECT PRODUCT PERFORMANCE · SEAL HERMETICALLY

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MAIN CHARACTERISTICS

VERSATILITY
Nine Available Plug-In Preamplifiers—Wide Band, Dual Trace, Low Level, Differential, and others for specialized applications.

HIGH PERFORMANCE
DC to 30 MC with fast-rise plug-in units.
DC to 24 MC with dual-trace plug-in unit.
0.02 μsec/cm to 15 sec/cm sweep range.

EASY OPERATION
24 Calibrated Direct-Reading Sweep Rates.
Sweep Magnification—2, 5, 10, 20, 50, and 100 Times.
Preset Triggering—Eliminates triggering adjustments in most applications.
Single Sweep Operation—Lockout-Reset Circuitry for one-shot recording.

HIGH WRITING RATE
250 cm/μsec. 10-kv accelerating potential assures bright trace for operation in single-sweep applications, and with low sweep repetition rates.

TYPE 543 PRICE, without plug-in units.......... **$1200**

Type 53/54K Fast-Rise Unit...................... **$125**
Type 53/54C Dual-Trace Unit.................... **$275**
Type 53/54R Transistor Test Unit.................. **$300**

Prices f.o.b. factory.

Please call your Tektronix Field Engineer or Representative for complete specifications and, if desired, to arrange for a demonstration at your convenience.

***ADD SWEEP LOCKOUT** to your Tektronix Type 531 and 541 Oscilloscopes—order Modification Kit*
K531 Sweep Lockout, Tek. 040-118....... $25
for Type 532
K532 Sweep Lockout, Tek. 040-147....... $25

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IRE People

(Continued from page 50A)

Herbert Harris, Jr. (S'37–A'41–M'55) has been appointed Manager of Sperry Gyroscope Company's Air Armament Division.

H. Harris, Jr.

Mr. Harris joined Sperry in 1942 as an assistant project engineer. Since then he advanced steadily to project engineer, senior project engineer, section head and, in 1948, was made head of the special weapons engineering department. In 1951 he became engineering director for air armament and, in 1957, chief engineer for the Air Armament Division.

He received his B.S.E.E. degree from Rensselaer Polytechnic Institute in 1938 and his M.S.E.E. degree the following year. He then worked as a research assistant at Massachusetts Institute of Technology's Center of Analysis while completing most of the requirements for his Ph.D. degree.

Ten patents have been filed in Mr. Harris' name and others are pending. He has written material for publication on servo-mechanisms and a wartime report on servos for the National Defense Research Council, in addition to many papers for various technical publications. He is also a member of the American Institute of Electrical Engineering.

❖

The appointment of **Peter Humeniuk** (SM'52) as Manager—Engineering of General Electric's Television Receiver Department, at Syracuse, N. Y. has been recently revealed.

Mr. Humeniuk is a native of Hamilton, Ontario and graduated from the University of Toronto in 1943 with a Bachelor of Applied Science degree in electrical engineering.

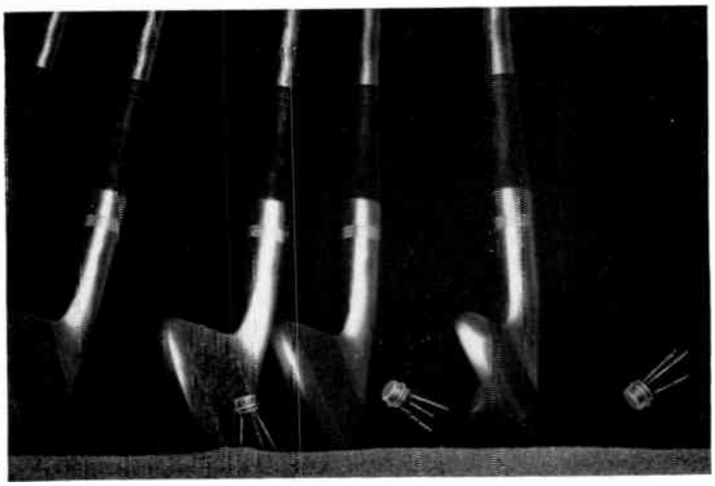
As an undergraduate, Mr. Humeniuk worked at Defense Industries, Ltd. in Ajax, Ontario as a technical assistant to the electrical maintenance superintendent and at the Ford Motor Company of Windsor, Ontario in both structural drafting and tool and die repair. Following his graduation, Mr. Humeniuk worked in the Research Enterprises of Toronto in the radar approvals laboratory and also as a laboratory demonstrator at the University of Toronto. In 1944, he joined the Canadian General Electric Co. as a radio engineer, later becoming a television engineer.

In 1948 he joined Philips of Canada as a television engineer heading up their electrical television subsection and spent six months in their main plant at Eindhoven, Holland, in TV engineering.

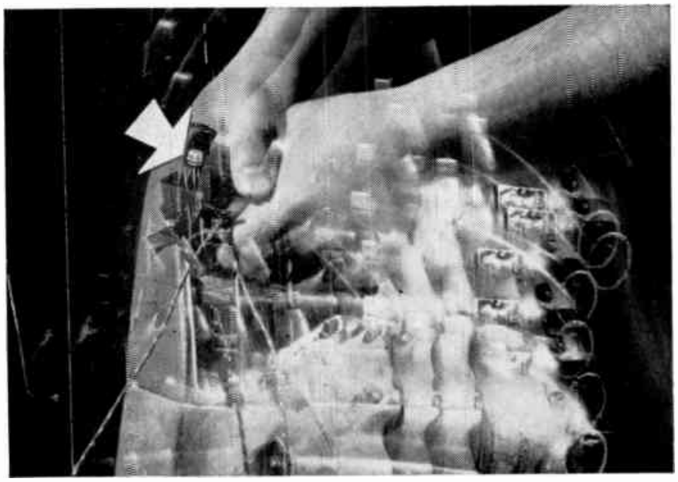
In 1950, he returned to the Canadian General Electric Company as manager of engineering for both radio and television. Two years later he was appointed manager

(Continued on page 56A)

tremendous impact and vibration



"GOLF CLUB TEST" General Electric transistors with Fixed-Bed Mounting have been struck full force with a No. 2 Iron. After traveling forty yards, tests showed they still worked perfectly.



"JACKHAMMER TEST" Another G-E transistor with Fixed-Bed Mounting was taped to a pneumatic drill, which was then operated for ten minutes. When the transistor was removed, tests showed it still worked perfectly.

Ceramic disk guards against major causes of transistor failure

In General Electric's new Fixed-Bed Mounting, critical elements of the transistor are welded flat on a disk of ceramic. Thus any impact must be great enough to damage the disk itself before transistor failure can occur. In conventional methods of manufacture, impact need only penetrate the transistor's metal case in order to damage the standard upright header.

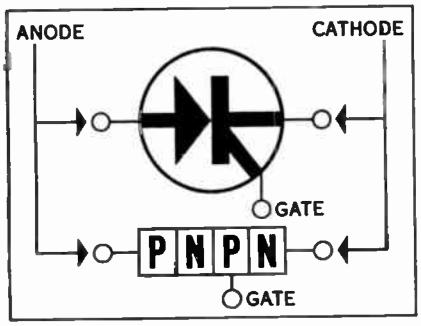
Because of their many suspended parts, standard upright headers are also subject to inertial stress at a number of points. General Electric's Fixed-Bed Mounting eliminated *all but one* of those parts—the suspended aluminum emitter lead. And this is provided with enough slack to absorb inertial stress, with connection points so securely welded that the unit withstands far more than the military centrifuge test of 20,000 G's.

To eliminate thermal stress, the coefficient of expansion of G.E.'s ceramic disk has been made equal to that of the semiconductor metal. Previously, enough "play" had to be allowed to absorb alternate expansions and contractions, thereby reducing the strength and stability of the unit.

The Fixed-Bed Mounting's electrical elements lie flat, in close contact to the transistor case, providing greater heat conduction out through the case. Therefore, the fixed-bed construction cuts down junction temperature, making it possible to double the power dissipation of the same transistor made with upright-header construction.

Fixed-Bed Mounted units have exceeded all standard shock, centrifuge and temperature-cycling tests. General Electric's unijunction transistor (see below) now has this feature.

New G-E Controlled Rectifier rectifies and controls current up to 5 amperes at 300 v.



The controlled rectifier is a four-layer silicon device with a "gate" to which a signal can be applied to control forward current. It can handle more than one kw of power.

NEED A FEW SEMICONDUCTORS IN A HURRY? Check your local G-E distributor *first*. You'll find his delivery, service facilities and prices are hard to beat.

General Electric's new silicon controlled rectifier acts like a thyratron. In the reverse direction, it's a standard rectifier. But it will also block forward current until either a critical breakover voltage is exceeded or a signal is applied to the third lead. Then it switches to a conducting state and acts as a forward-biased silicon rectifier.

The controlled rectifier can be actuated by a little as 15 mw. Breakdown occurs at speeds approaching a microsecond, after which voltage across the device is so low that current is determined by the load. This enables the user to control a large anode-to-cathode current with an extremely small amount of power, or to switch power from high impedance to low impedance in microseconds.

Applications include replacement of relays, thyratrons, magnetic amplifiers, power transistors and conventional rectifiers. Sample quantities of the controlled rectifier are now available. Prices will be sent on request.

GENERAL ELECTRIC

IRE People

(Continued from page 53A)

of engineering and manufacturing, radio and television and, in 1955, manager of that company's radio and television plant operation. This latter post he held until this recent appointment.

He is a member of the Association of Professional Engineers of Ontario and has been actively engaged in the Engineering Alumni Association of the University of Toronto.

❖

O. M. Dunning, Vice-President in charge of the Engineering & Production Division of Airborne Instruments Laboratory, Mineola, N. Y., has announced the appointment of **W. F. Woodbury** (A'44–M'45) to the position of Assistant to the Vice-President. Mr. Woodbury was formerly the sales manager of the government and commercial division of Hazeltine Electronics.

W. F. Woodbury

Mr. Woodbury is a member of the Washington Engineers Club, American Ordnance Association, Armed Forces Communications and Electronics Association, and the American Society of Naval Engineers.

❖

The appointment of **E. M. Baldwin** (M'54) as general manager of Fairchild Semiconductor Corporation, Palo Alto, Calif., has recently been announced. Dr. Baldwin will also serve as a vice-president and member of the board of directors of the corporation.

E. M. Baldwin

The company, which was formed in October, 1957, is presently setting up to manufacture a line of high-frequency silicon diffused transistors.

Dr. Baldwin has an extensive background in large-volume production of semiconductor devices. He has been associated with the Hughes Aircraft Company, Products Group, Semiconductor Division, for the past five years, most recently as manager of product engineering. Prior to joining Hughes, Dr. Baldwin was assistant director of the Nuclear Research Center at Carnegie Institute of Technology. He received both the M.S. and D.Sc. degrees in physics from Carnegie Institute of Technology.

Dr. Baldwin is a member of the American Physical Society and served as Chairman of the IRE Professional Group on Electron Devices for the Los Angeles Chapter.

(Continued on page 58A)

...application problems solved by ESC custom-built delay lines

In the past three years alone custom-built ESC delay lines have solved 6,704 different military and industrial application problems...so it's a good bet that we are already experienced in designing a unit very close to what you need. As the first company exclusively devoted to the manufacture of delay lines, we pioneered in the elimination of costly overspecification and brought economical custom-building to the delay line field.

To insure strict adherence to your specifications, the prototype unit is subjected to comprehensive testing and the results of these tests are submitted to you. On this laboratory report are included submitted electrical requirements, photo-oscillograms (which indicate input and output pulse shape and output rise-time), the test equipment used, and an evaluation of the electrical characteristics of the prototype.

Whatever the application, the odds are 6,704 to 1 that ESC can design and build precisely the delay line you need—and do it easily, efficiently and exactly as specified! Write today for complete technical data.

exceptional employment opportunities for engineers experienced in pulse techniques

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534 Bergen Boulevard, Palisades Park, New Jersey

IRE People

(Continued from page 56A)

L. J. Boss (A'44–M'50) has been appointed to the recently created post of Engineering Administrator for Lynch Carrier Systems, Inc., manufacturers of carrier, telephone and multiplexing equipment. He will serve in the capacity of business manager for the Engineering Department.

L. J. Boss

The position of Engineering Administrator has been established to relieve engineers of the administration of record keeping, information flow and coordination of engineering work with other departments.

Mr. Boss came to Lynch from the San Francisco office of the Philco Corporation where he was West Coast regional manager for the government and industrial division during the past eight years.

Born in R. I., Mr. Boss completed his education in electrical engineering at the University of R. I. He began in the East with electric power utilities and after a period of self-employment, he became sales manager of mobile communication for the Federal Telephone & Radio Corporation.

In addition to industrial activities he did important pioneering work in photoelectric photometry under the supervision of Harvard College Observatory and the American Association of Variable Star Observers.

❖

The appointment of **J. R. Saliba** (M'39–SM'43) as sales engineer for Du Mont oscilloscopes and associated electronic test instruments is announced by Allen B. Du Mont Laboratories, Inc.

Mr. Saliba will be part of the home office sales engineering force responsible for providing technical back-up and information for customers and the field sales force. He will help set up technical training programs in the field and in the plant, and assist in all phases of the instrument division program.

Immediately prior to his new appointment, Mr. Saliba served with the U. S. Army Ordnance Corps as a second lieutenant. Before military service, he was associated with Presto Recording Co. and Avion Division, ACF Industries, both in Paramus, N. J.

Mr. Saliba received his master's degree in business administration from the Harvard Graduate School of Business Administration, after he was graduated from Massachusetts Institute of Technology with a bachelor's degree in electrical engineering.

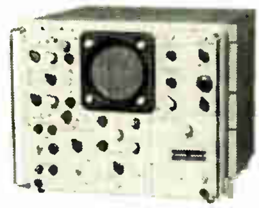
He holds membership with the American Ordnance Association and Phi Mu Delta.

(Continued on page 60A)

THE NEW BIG FOUR

by Lavoie labs

LA-260 Oscilloscope . . .

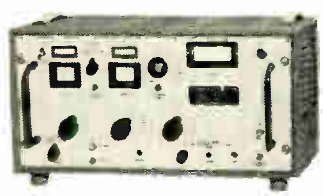


the first CRO designed to military requirements with plug-in single or dual trace vertical preamplifiers. New technique permits all d-c supply voltage regulation to better than 0.1% . . . including d-c filament voltage. Flat 5" CRT increases viewing area, screen visibility from greater distances. Improved electron optics assure brighter high-speed pulse traces.

LA-80 Electronic Counter . . .

high reliability and wide frequency range are featured in this superior designed counter. Eight place, in-line read outs.afford clear, sharp digits, visible at any angle. Other features include MIL spec design, temperature insensitivity, wide time interval range and simplified circuitry.

LA-70 Frequency Meter . . .



generates and measures frequencies from 10 KC to 3000 MC with 1 x 10^{-6} accuracy . . . particularly suitable for VHF receiver measurements in mobile service . . . weighs only 42 pounds. Oscillator is stabilized by use of thermostatically controlled 1 MC precision quartz crystal. Stability over six months — 1 part 10^6.

LA-90 $10x^{-9}$ Frequency Standard . . .

based on new approach to precise crystal oven regulation to provide (1) long term temperate life (2) excellent stability over wide ambient temperatures (3) elimination of permanent temperature shifts found in thermostat or thermistor devices, and (4) use of JAN tubes and magnetic beam switching tubes for reliability.

For complete information write to:

Lavoie Laboratories, Inc.

MORGANVILLE, NEW JERSEY

IRE People

(Continued from page 58A)

J. E. Ebert (A'43–SM'53) has been appointed to the newly-created post of Chief Microwave Engineer of the F-R Machine Works, Inc. He will have engineering responsibility for the Microwave Components and Microwave Test Equipment that sells under the tradename of FXR.

J. E. Ebert

Mr. Ebert joined the company in 1953 as the assistant chief electrical engineer. His work since that time has been primarily connected with the design and introduction of an extensive commercial line of precision microwave test equipment.

Since 1942 Mr. Ebert has been associated with the test equipment field in various engineering and supervisory positions. During World War II he was a staff member of the Microwave Research Institute of the Brooklyn Polytechnic Institute, then engaged in special development work for the Radiation Laboratory of M.I.T. Until 1953 Mr. Ebert was an instructor in the evening school of the Brooklyn Polytechnic Institute.

Mr. Ebert received his B.E.E. degree from the Brooklyn Polytechnic Institute. He has written several papers on uhf and microwave measurement techniques and equipment. In addition, he holds more than a dozen patents in the United States, Canada and Great Britain. Mr. Ebert is a member of Sigma Xi, Eta Kappa Nu, and Tau Beta Pi.

❖

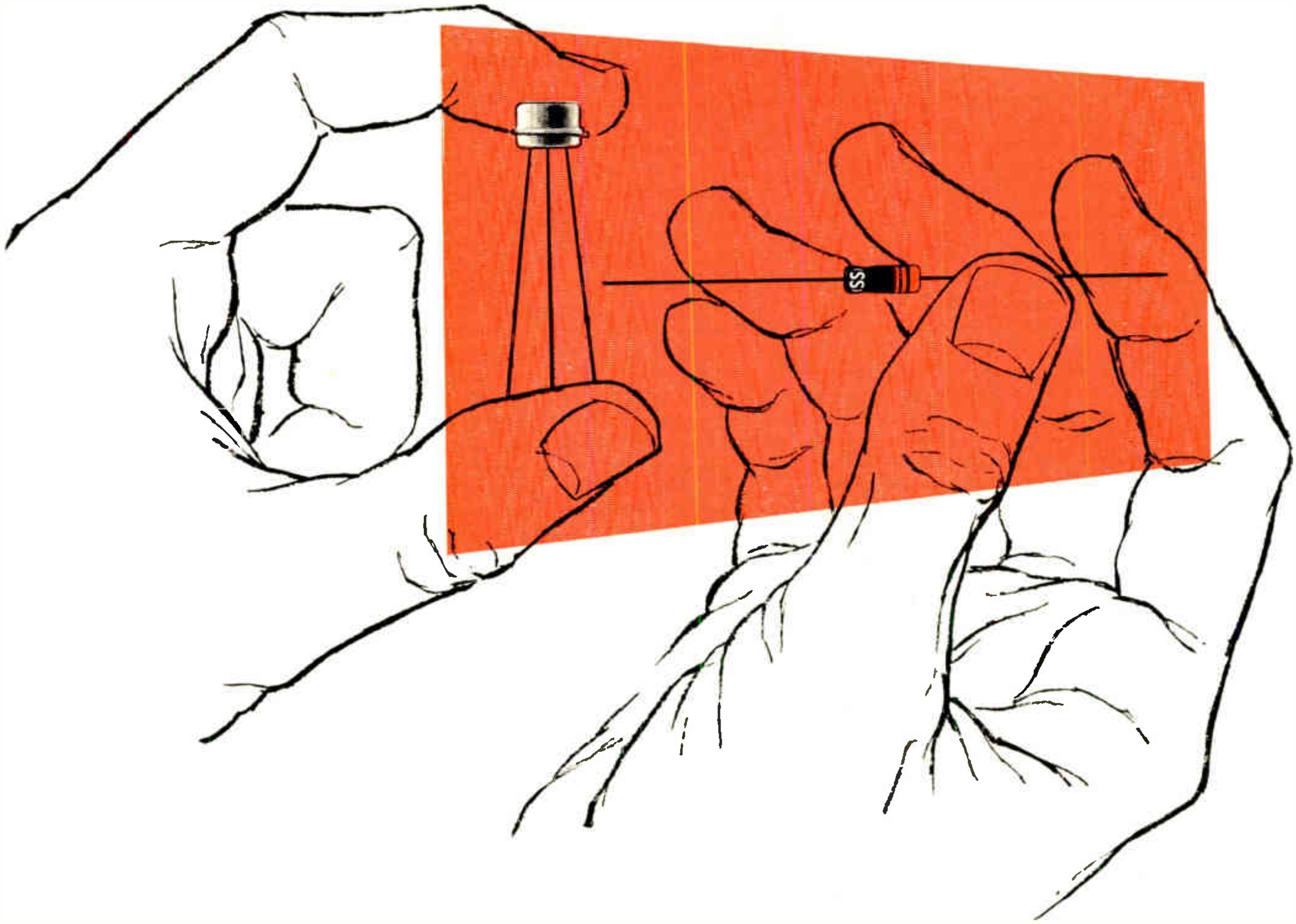
The U. S. Naval Research Laboratory has announced the appointment of **A. H. Schooley** (A'35–SM'47–F'54) as the new Associate Director of Research for Electronics.

A. H. Schooley

Mr. Schooley recently resumed his position as Superintendent of NRL's Electronics Division after a year's leave of absence. During this time, he served under the terms of the Mutual Assistance Pact as an advisor to the Brazilian Navy in matters related to the establishment of a Brazilian Naval Research Institute in Rio de Janeiro.

He has been with the Naval Research Laboratory since 1940, becoming the first superintendent of the Electronics Division when it was formed in 1954. "For development of precision ranging circuits for fire-control radar," Mr. Schooley was awarded the Distinguished Civilian Service Award,

(Continued on page 62A)



ANNOUNCING Sperry Silicon Semiconductor Devices

High-temperature diodes and transistors now in production

The Sperry Semiconductor Division of Sperry Rand Corporation is now making available to military and commercial manufacturers a new line of silicon devices. Performance proven, these high-quality diodes and transistors have been employed in many Sperry Rand systems which had to meet stringent military and commercial specifications.

SILICON DEVICES NOW IN PRODUCTION

- **High-conductance diodes for general purpose applications. 100, 200 and 400 ma types (rated current at 1.0 v). Working voltage up to 300 volts. Subminiature glass package.**
- **High-current switching diodes. Switches ½ amp. in less than 0.8 μsec. Reverse voltage up to 200 volts. Subminiature glass package.**
- **Ultra-fast computer diodes for all computer requirements. Working voltage up to 200 volts. Subminiature glass package.**
- **High-speed computer transistors. Total switching time typically less than ½ μsec. Very low saturation resistance. JETEC-30 case.**

Write for data sheets on all these new production items. We also welcome inquiries on any applications calling for special silicon semiconductor devices.

SPERRY SEMICONDUCTOR DIVISION
Sperry Rand Corporation
South Norwalk, Connecticut

ADDRESS ALL INQUIRIES: Marketing Department, Great Neck, N. Y., or Sperry Gyroscope offices in Brooklyn, Cleveland, Seattle, San Francisco, Los Angeles, New Orleans, Boston, Baltimore, Philadelphia.

IRE People

(Continued from page 60A)

the Navy's highest civilian award, in 1945. He has done considerable research in oceanography as it affects radar propagation, and is also active in advancing engineering management.

A native of Terril, Iowa, Mr. Schooley received a Bachelor of Science degree in electrical engineering from Iowa State College and his Master of Science degree in the same field from Purdue University. Before coming to NRL, he was with RCA Radiotron, where he designed and built the first miniature radio tubes, now widely used by the electronics industry.

A registered professional engineer in the District of Columbia and a member of Sigma Xi, Mr. Schooley has several patents in the field of electronics and is the author of numerous technical articles. He is presently serving as chairman of the Washington, D. C. IRE Section.

❖

A. C. Petrasek (SM'52) has joined the Applied Science Corporation of Princeton (ASCOP) as southwestern sales manager of the electronics company's industrial products line.

A. C. Petrasek

Mr. Petrasek will direct the sales engineering of ASCOP's new industrial telemetering and supervisory control system in the southwestern states. The company's district office is at 4918 Greenville Ave., Dallas, Tex.

For the last four years, Mr. Petrasek was with Collins Radio, first as sales manager for microwave equipment in Dallas and later as sales manager for Collins Radio of Canada Ltd. Previously, he was engaged in sales engineering of world wide communications projects for General Electric International.

Mr. Petrasek has been an amateur radio operator (W5IXI) for 25 years. A retired Army infantry captain, he was a communications instructor at Ft. Benning, Georgia.

He is a member of the Petroleum Electric Supply Association and the Armed Forces Communications and Electronics Association.

❖

C. H. Knowles (M'55) has joined the Semiconductor Division of Motorola Inc. In his new post Knowles's responsibility will be for development and exploratory production of video vhf and uhf transistors.

Mr. Knowles comes to Motorola from the Bell Telephone Laboratories at Murray Hill, N. J. There he supervised exploratory developmental efforts in high frequency transistors. He was also re-

(Continued on page 64A)

ONE INTEGRATED SOURCE for Ceramic-to-Metal Seals

Standard types of Alite high voltage bushings are available in various sizes and configurations.

In *all* phases of planning for ceramic-to-metal seals—from design to finished assembly—you can rely on ALITE for the know-how and "*do*-how" required to produce highest quality ceramic-metal components for critical applications.

High alumina Alite is the ideal material for making rugged, high performance hermetic seals and bushings. It has superior mechanical strength, high temperature and thermal shock resistance, plus reliable electrical characteristics. Our complete high temperature metalizing and bonding facilities assure delivery of the finest seals available—mass-spectrometer tested for vacuum-tightness.

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IRE People

(Continued from page 62A)

sponsible for the Bell Laboratories production engineering in high frequency transistors at the Western Electric Company, Laureldale, Pa.

A graduate of Alabama Polytechnic Institute, Mr. Knowles also attended Vanderbilt University.

❖

The appointment of **C. L. Hogan** (SM'54) Professor of Applied Physics at Harvard University to the post of Manager of Motorola's Semiconductor Div. at Phoenix, Ariz. was recently made known by D. E. Noble, executive vice-president in charge of the Communications and Industrial Electronics, Military and Semiconductor Divisions.

Dr. Hogan will complete the university year at Harvard and will move to Phoenix in June.

Dr. Hogan's twelve years' of civilian experience embraces his experiences as research engineer for Anaconda Copper Mining Company, instructor at Lehigh University, member of technical staff of Bell Laboratories, and Professor of Applied Physics at Harvard University. He has written extensively, holds two patents, and acted as consultant for several industrial laboratories. He is an authority on the microwave properties and application of ferrites.

Dr. Hogan is a member of the American Physical Society, Technical Panel on Magnetic Materials (Materials Advisory Board for the Department of Defense), and Subcommittee on Magnetic Materials (AIEE) Advisory Council of the Department of Electrical Engineering of Princeton University, and the Second vice-president of Harvard Engineering Society.

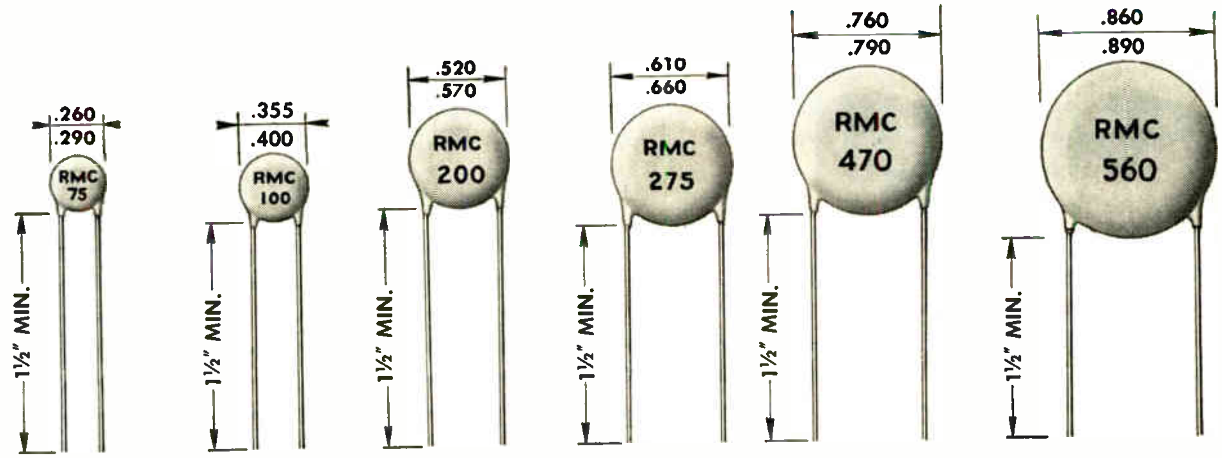
❖

Sperry Gyroscope Company has recently announced the formation of a new Countermeasures Division. The new division, currently weapon system division for production and development of B-52 electronic countermeasures, also will be responsible for other company ECM programs.

N. L. Winter (M'47) heads the new Countermeasures Division. As manager of an extensive and autonomous organization within the Sperry Gyroscope family, Mr. Winter will be in full charge of company ECM engineering, manufacturing, sales, purchasing and service activities.

RELY ON RMC

for TC Capacitors



TC	1/4 Dia.	5/16 Dia.	1/2 Dia.	5/8 Dia.	3/4 Dia.	7/8 Dia.
P-100	1- 3 MMF	4- 9 MMF	10- 20 MMF	—	—	—
NPO	2- 13	14- 30	31- 69	70- 85MMF	86-115 MMF	116-150 MMF
N- 33	2- 13	14- 30	31- 56	57- 62	63-100	101-150
N- 75	2- 15	16- 30	31- 56	57- 68	69-125	126-150
N- 150	2- 15	16- 30	31- 67	68- 75	76-140	141-175
N- 220	3- 15	16- 30	31- 75	76-100	101-140	141-175
N- 330	3- 15	16- 30	31- 75	76-100	101-150	151-190
N- 470	3- 20	21- 51	52- 80	81-120	121-200	201-240
N- 750	5- 30	31- 68	69-150	151-220	221-300	301-375
N-1500	10- 51	52-120	121-200	201-270	271-470	471-560
N-2200	20- 75	76-150	151-200	201-300	301-680	—

TYPE C DISCAPS meet all specifications of the EIA standard RS-198. These temperature compensating DISCAPS are rated at 1000 V.D.C. to provide a higher safety factor than other standard or mica capacitors.

Constant production checks assure that all specifications on temperature characteristics are met. Another phase of RMC quality control consists of a 100% test for capacities.

Over the years leading manufacturers have relied on RMC for quality of product and maintenance of delivery schedules. Write today on your company letterhead for information.

SPECIFICATIONS

LIFE TEST: As per EIA-RS-198

POWER FACTOR: Over 10 MMF less than .1% at 1 megacycle. Under 10 MMF less than .2% at 1 megacycle

WORKING VOLTAGE: 1000 V.D.C.

TEST VOLTAGE (FLASH): 2000 V.D.C.

CODING: Capacity, tolerance and TC stamped on disc

INSULATION: Durez phenolic-vacuum waxed

INITIAL LEAKAGE RESISTANCE: Guaranteed higher than 7500 megohms

AFTER HUMIDITY LEAKAGE RESISTANCE: Guaranteed higher than 1000 megohms

LEADS: No. 22 tinned copper (.026 dia.)

TOLERANCES: ±5% ±10% ±20%

These capacitors conform to the E.I.A. specification for Class 1 ceramic capacitors.

The capacity of these capacitors will not change under voltage.

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GENERAL OFFICE: 3325 N. California Ave., Chicago 18, Ill.
Two RMC Plants Devoted Exclusively to Ceramic Capacitors
FACTORIES AT CHICAGO, ILL. AND ATTICA, IND.

Section Meetings

Akron

"Documentation Retrieval," J. W. Perry, Western Reserve University; Joint meeting with PGEC; 2/18/58.

Alamogordo-Holloman

"Component Reliability," J. K. Sprague; "Selected Aspects of Quality Control," B. Hecht; "New Products for Military Applications," C. Killen, Sprague Electric Co.; 2/19/58.

Albuquerque-Los Alamos

"Electronics—A Little History and A Look Ahead," J. Sprague, C. G. Killen, and Bernard Hecht, Sprague Electric Co.; 2/17/58.

Atlanta

"Space Travel," V. Crawford, Georgia Institute of Technology; 2/28/58.

Baltimore

"Simplifying Linear and Non-Linear Systems," S. J. Mason, M.I.T.; 2/12/58.

Bay of Quinte

"Radar Weather Information Display System," T. W. R. East, McGill University; 3/11/58.

Beaumont-Port Arthur

Visit Facilities of Air Force Base, H. S. Williams, Jr., USAF; 2/25/58.

Buffalo-Niagara

"The Role of Electronics in U. S. Naval Aviation," W. F. Cleaves, Bendix Radio Corp.; 3/12/58.

Central Florida

"Effects of Nuclear Weapons," J. C. Clark, Convair; 1/16/58.

Film—"The Challenge of Outer Space," W Von Braun, L. M. Orman, Army Ballistic Missile Agency; 2/19/58.

Cincinnati

"Sources of Radio Interference in Automotive Electrical Systems," B. H. Short, General Motors; 2/18/58.

Dallas

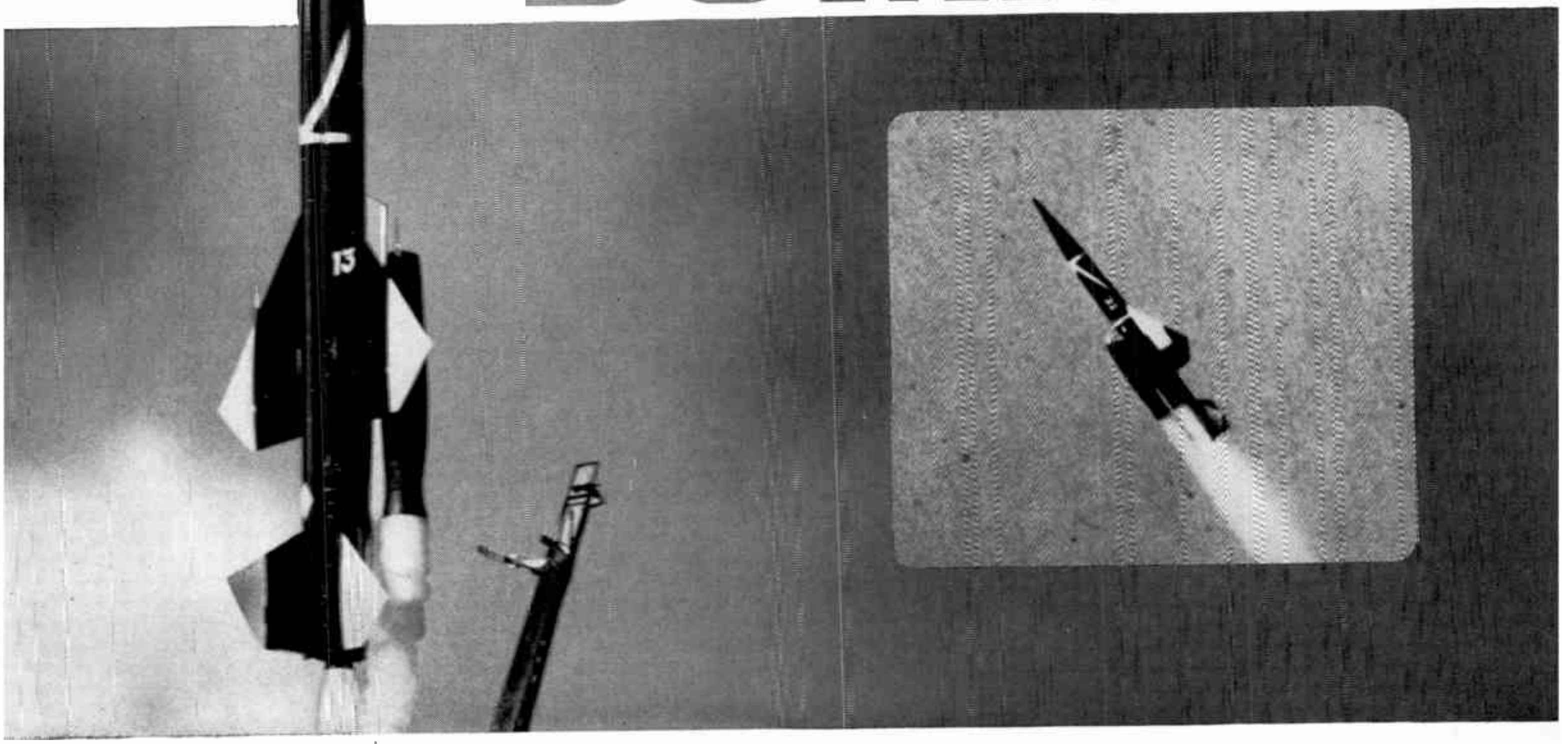
"Problems and Progress in Computer Process Control," T. M. Stout, Ramo-Wooldridge Corp.; 2/4/58.

Panel: "The Role of the Electronics Department in Modern Aircraft and Guided Missile Companies," R. C. Blaylock, Chance Vought; B. Kelly, Bell; I. N. Palley, Temco; F. W. Davies, Convair; Moderator: W. R. Hedeman, Texas Instruments; 3/4/58.

(Continued on page 68A)

Preserver of Peace...

BOMARC



Official U.S. Air Force Photo

It tracks down an enemy at 300 miles

Described as the most potent of all ground-to-air defense missiles, the Bomarc pilotless interceptor, designed by Boeing, stands poised for the destruction of any "enemy" bomber within a 200-300 mile range. Its booster rocket has the power to hurl it more than 60,000 feet straight up; then, powered by two ramjet engines, it hurtles by electronic instinct to its target at up to 3 times the speed of sound. For this guardian of our homes and way of life, RCA has been privileged to supply important advance components of the guidance system.

RCA Tmk(s) ®

RADIO CORPORATION of AMERICA

DEFENSE ELECTRONIC PRODUCTS

CAMDEN, NEW JERSEY

Section Meetings

(*Continued from page 66A*)

Denver

"Microwave Generation & Amplification by Atomic and Molecular Processes," R. C. Mockler, National Bureau of Standards; 2/21/58.

Detroit

"900 Megacycles for Mobile Communications," C. Willyard, Motorola, Inc.; Joint meeting with PGVC; 2/21/58.

Emporium

"Ceramic Stacked Mount Receiving Tubes," R. Slinkman, Sylvania Electric Prod. Inc.; 2/18/58.

Florida West Coast

"Data Link Transmission Systems," C. K. Law, RCA; 2/19/58.

Fort Huachuca

"Digital Data Concepts," R. E. Marquand, Radiation, Inc.; 2/24/58.

Hamilton

Film Entitled—"The Strange Case of the Cosmic Rays," 3/10/58.

Hawaii

"A Practical Approach to Selecting a Loudspeaker Enclosure," D. Pang, U. S. Naval Shipyard; 1/8/58.

"Between the Atmospherics," G. Reber; 2/12/58.

"Psychology and the Electronics Engineer," F. Mason, Pearl Harbor Naval Shipyard; 3/12/58.

Houston

"Oceanographic Instruments during the IGY," H. J. McLellan, Texas A & M College; 2/18/58.

"Radio Waves from Outer Space," J. W. Findley, National Radio Astronomy Observatory; 3/4/58.

Huntsville

"Wave Analysis Problems in Guided Missile Research," T. L. Greenwood, Army Ballistic Missile Agency; 2/25/58.

Israel

"Can Energy be Generated by Controlled Thermonuclear Fusion," G. Schmidt, Technion Institute Haifa; 1/21/58.

Ithaca

"Infra red," A. H. Canada, General Electric Co.; 2/13/58.

Kansas City

"Symposium on Millimicrosecond Pulse Techniques," N. S. Nahman, C. Womack, E. J. Martin Jr., Univ. of Kansas Research Foundation; 1/14/58.

First Annual Award Banquet, Honoring J. C. Shipman, speaker—W. H. Graham; 2/11/58.

Long Island

New Transistor Applications—"New Highs in Transistors," J. Angell; 2/11/58.

"Missile Guidance—Radar Guidance," Messrs. Records, Hill, Muller, Revercomb, General Electric Co.; 2/13/58.

"Radio Navigation and Command Guidance," W. Palmer and R. Davis, Sperry; 2/20/58.

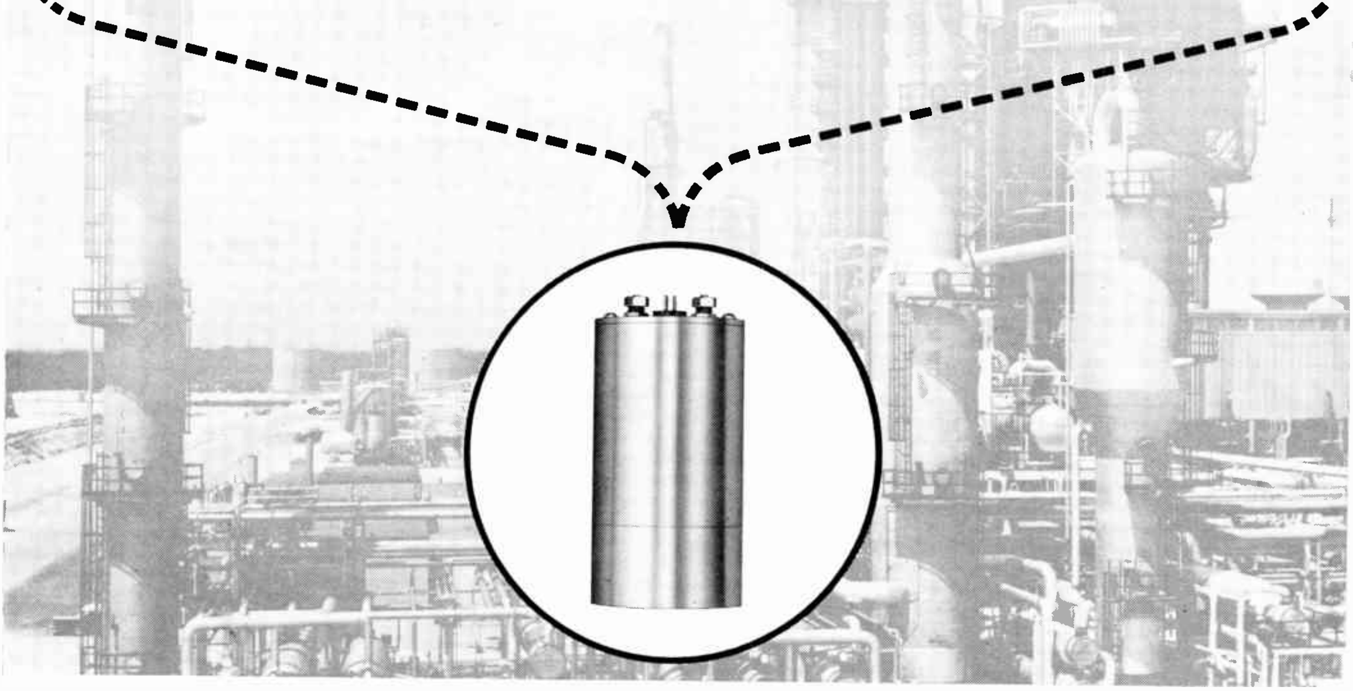
"Infra red Guidance," J. Sanderson; 2/27/58.

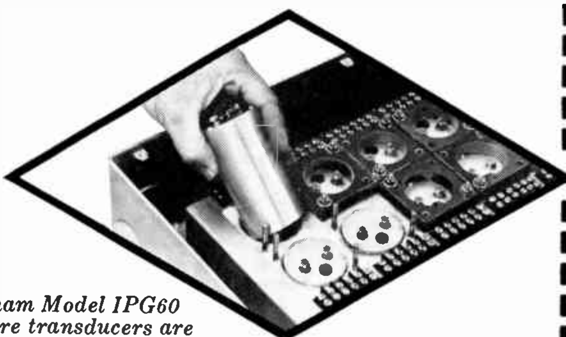
"Guidance System Evaluation," J. C. Fletcher; 3/6/58.

(*Continued on page 71A*)

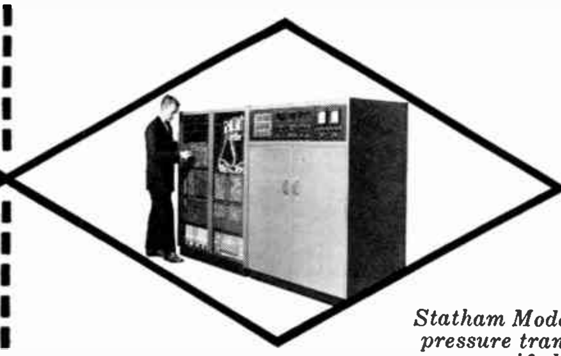
IPG60 PRESSURE TRANSDUCERS

THE RELIABLE HEART OF PROCESS CONTROL





Statham Model IPG60 pressure transducers are suitable for a variety of applications requiring accurate pressure measurements.



Statham Model IPG60 pressure transducers are specified for the Beckman 112 Data Processing Computer.

The evolution of industry toward automation has created a demand for precision data handling systems capable of monitoring and controlling process variables. Statham Model IPG60 pressure transducers are specified in the Beckman 112 Data Processing Computer — the first system of its type to meet successfully the stringent requirements of the process industry. The output of this highly accurate pressure transducer provides the Beckman system with vital information signals required for successful process control. Accuracy within ±¼% makes the Model IPG60 suitable for use in a wide variety of pressure measurement requirements. Complete data on the Model IPG60, or information on other Statham instruments, are available upon request.

OUTPUT: Approximately 35 millivolts full scale at 14 V. excitation
PRESSURE MEDIA: Non-corrosive fluids

statham instruments, inc.

of puerto rico

254 Carpenter Road, Hato Rey, Puerto Rico • Cable Address: STATRICO

A subsidiary of Statham Instruments, Inc., Los Angeles, Calif.

RANGE: 0 to +15 psig.
NON-LINEARITY & HYSTERESIS: Not more than ±0.25% of full scale

Section Meetings

(Continued from page 68A)

MILWAUKEE

"The Effects of Research," F. J. Larsen: Joint meeting Engineers' Society of Milwaukee, and affiliated Engineering Societies; 2/20/58.

"Correlation Analysis," V. C. Rideout. University of Wisconsin; 3/4/58.

NORTH CAROLINA

"Basic Principles of Analog Computers," H. A. Owens. Duke University; 2/28/58.

NORTHERN ALBERTA

"Preservation of Wave Shapes in Television," A. St. Marie, Canadian Broadcasting Corp.; 2/14/58.

NORTHWEST FLORIDA

"Communications Theory and the Test Problem," J. C. O'Brien, Associated Missile (AMF) 1/28/58.

"High Gain Telemetry Antennas," C. H. Hoeppner, Radiation Inc.; 2/19/58.

"Microwave Amplification by Stimulated Electronic Radiation," R. F. Simons, Airborne Instruments Lab.; 3/4/58.

OKLAHOMA CITY

"Microwave Radio Application to Protective Relaying," L. G. Walker, Motorola Inc.; 2/18/58.

"The Road Ahead in Data Reduction," H. P. T. Corley, U. S. Air Force; 3/11/58.

OMAHA-LINCOLN

"Rockets, Satellites, and Space Travel," W. J. Conner, Jr., and B. O'Brien, General Electric Co; 2/27/58.

OTTAWA

"Trend of Overseas Telecommunication Development," R. G. Griffith, Canadian Overseas Telecommunications Corp.; Joint meeting with Engineering Institute of Canada, AIEE, IRE; 2/27/58.

"Cunning Circuits,"—A Panel Discussion, N. F. Moody, Moderator; Panel Members: A. C. Hudson, E. H. Hugenholtz, M. Levy, J. Loutit; 3/6/58.

PHILADELPHIA

"Management of a Navy R & D Activity," S. B. Spangler, U. S. Naval Air Dev. & Material Center; 3/5/58.

PITTSBURGH

"The Theory and Application of Solion Units," D. W. Kuester, Naval Ordnance Labs.; 2/10/58.

"Transistor Versus Magnetic Amplifiers," A. G. Milnes, Carnegie Institute of Technology; Donald G. Fink, IRE President, presented Fellow award to R. A. Ramey at dinner meeting on Feb. 25; 2/17/58.

PRINCETON

"Natural Computers—Reliable Response of Unreliable Components," W. S. McCulloh, M.I.T.; presentation of Fellow Citation to R. W. Peter; 2/13/58.

REGINA

"Video Testing Techniques in Television Broadcasting," A. Ste. Marie, Canadian Broadcasting Corp.; 2/11/58.

SAN ANTONIO

"The Solion Cell—A New Circuit Device," N. N. Estes, Texas Research Associates; 1/15/58.

(Continued on page 72A)

Section Meetings

(Continued from page 71A)

Seattle

"Digital Differential Analyzers," D. Thompson, Testco; 2/5/58.

"The Education and Scientific Challenge we Face Today," Senator Jackson, U. S. Senator from Washington; Joint meeting with IRE, WSPE, ASME, AIEE, AICHE, ASM, ISA, and ASCE; 2/14/58.

"A 100 Channel Pulse Height Analyzer, Theory and Circuit Design," G. Utting, Technical Measurements Co.; 3/3/58.

Shreveport

Panel Discussion On the Explorer, A. M. Randolph, R. W. Bains, H. Porter, E. Gordon; E. D. Nuttall, United Gas Labs.; 3/4/58.

South Bend-Mishawaka

"The Bendix-Missile Plant Computing Facility," J. R. Voss, Bendix Aviation Corp.; 1/30/58.

Tokyo

"Engineering Educational Situation in the United States," J. D. Ryder, Michigan State Univ.; 1/18/58.

Toledo

"Germanium The Magic Metal,"—Tape Talk by General Electric Co.; 1/16/58.

Guided tour through the Sun Oil Company; 2/13/58.

Toronto

"Recent Developments in Video Tape Recording Equipment," J. E. Detlor, Ampex Corp. of Canada; 3/3/58.

Twin Cities

"Noise in Transistors," A. Van Vliet, Univ. of Minn.; 2/17/58.

Data Transmission: Present and Future, A. B. Convey, American T & T Co.; 3/11/58.

Vancouver

"Radio Astronomy," L. Doherty, National Research Council; 1/20/58.

"An Introduction to Information Theory," F. Schrack; "Electronic Counter-Measures," G. Sage; "Radio-Telephone Communications,"; University of British Columbia; annual students' night; 2/17/58.

Washington

Panel Discussion of "Transition Problems in the Development and Integration of Equipment into Presently Operating Two-Way Mobile Radio Systems," C. B. Plummer, FCC, J. P. Fogarty, Bendix, W. J. Weisz, Motorola, S. F. Meyer, DuMont Labs., J. R. Neubauer, RCA, N. Shepherd, General Electric; Joint meeting with PGVC; 3/3/58.

Western Massachusetts

"Missile Guidance," C. E. Gallager, G.E.; 2/11/58.

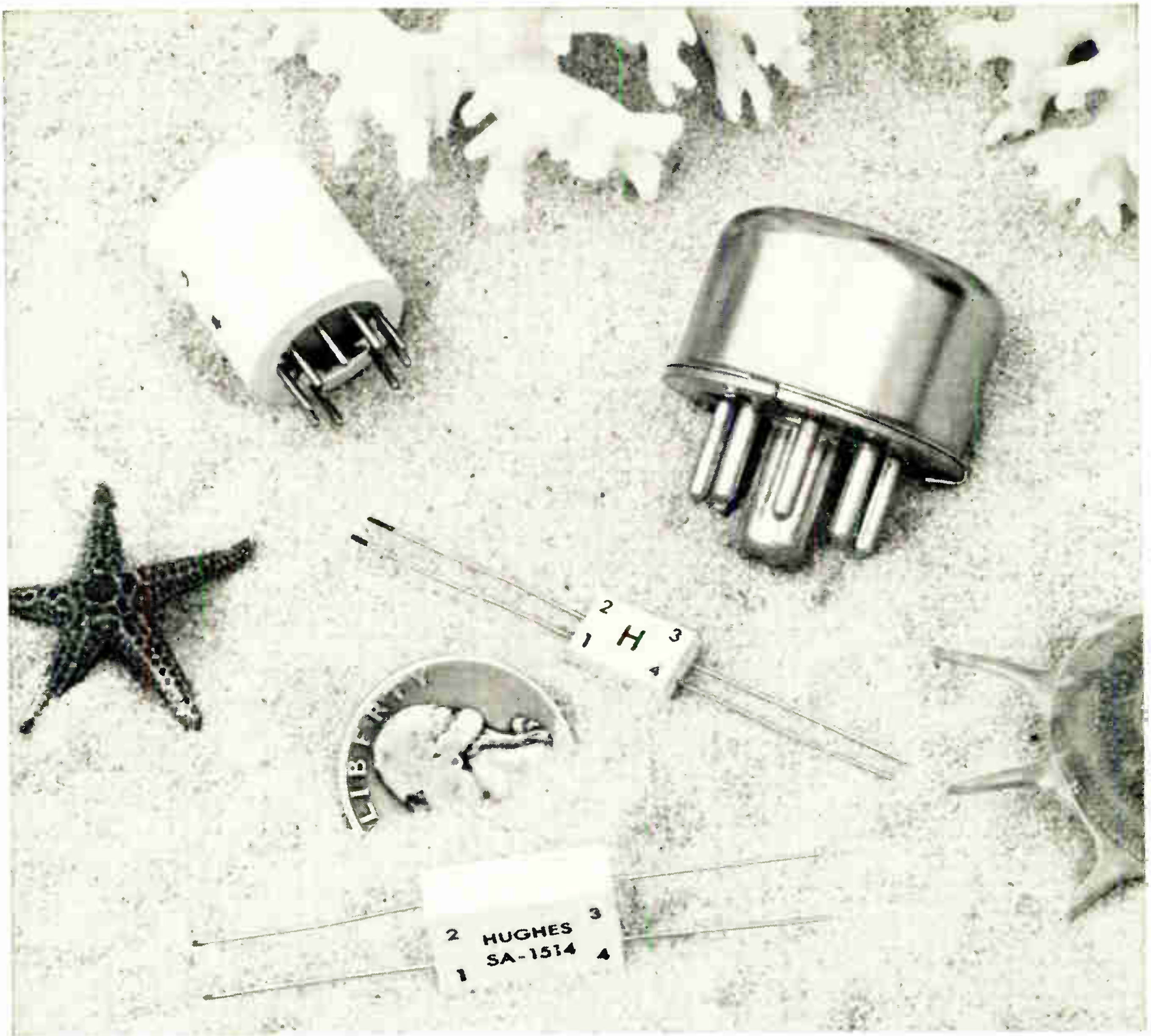
Wichita

"Ionosphere Probing by Radio," R. L. Schrag, Univ. of Wichita; 2/26/58.

Williamsport

Discussion of Section's activities, W. H. Bresee, Presiding Officer; 2/13/58.

(Continued on page 74A)



SPECIAL PACKAGING FOR DIODE ASSEMBLIES

At Hughes, the technique of multiple unit packaging has been perfected to an extent never before achieved. Now specific circuit configurations can be housed in any one of four Hughes packages. Each has its own advantages but all offer one prime advantage—convenience.

Many individual parts are reduced to a single component—solving a spare parts problem for matched units by eliminating the chance they will become separated; circuit design and installation are simplified; and space problems are minimized by the unusual compactness of Hughes multiple unit packaging.

With these features Hughes combines the ability to adjust to a wide range of individual requirements, while providing a completely satisfactory assembly.

ENCAPSULATED PAIR
.468" x .312" x .2"

ENCAPSULATED QUAD
.75" x .5" x .25"

These two Hughes solder-in units can house either matched or unmatched diodes in a variety of different ways.

METALLIC OCTAL SOCKET

ENCAPSULATED 9-PIN MINIATURE

Both are plug-ins intended primarily as direct tube replacement, and both can be adapted to contain special circuit configurations.

Applications: Full wave rectifier—bridge rectifier—modulator—demodulator—phase detector—and many others.

For literature, write: HUGHES PRODUCTS, Semiconductor Division, International Airport Station, Los Angeles 45, Calif.

HUGHES SEMICONDUCTORS

Creating a new world with ELECTRONICS

HUGHES PRODUCTS

Section Meetings

(Continued from page 72A)

SUBSECTIONS

Charleston

"Tropospheric Scatter Communications," R. D. Chippe, Federal Tel. & Tel. Labs.; Joint meeting with AFCEA; 1/15/58.

"Decisions for Americans," R. Bennet; 2/20/58.

Kitchener-Waterloo

"Proposed Revision of Methods A-343 for AC Magnetic Testing," D. C. Dieterly, Armco Steel Co.; 2/24/58.

"The Acoustical Design of the Stratford Festival Theatre," R. H. Tanner, Northern Electric Co. Ltd.; 3/10/58.

Lancaster

"Research, An Investment in the Future," C. T. Pearce; Westinghouse Electric Corp.; 1/16/58.

Las Cruces-White Sands

"Current Trends in Component Development," J. K. Sprague; "Military Applications and New Products," C. G. Killen; "Quality Control," B. Hecht, Sprague Electric Co.; 2/20/58.

"The Communication Theory Model and Economics," by S. Bagno; speaker: I. L. Carbine, New Mexico College of Agriculture and Mechanic Arts; 3/11/58.

Lehigh Valley

"Adventures in Electronics," C. Hoyler, R.C.A.; 2/24/58.

Nashville

"See Where You Can't Be," M. M. Haertig, G.E. Co.; Election of officers; 2/12/58.

Northern Vermont

"Medical Electronics Research," F. Sichel, University of Vermont; 2/24/58.

Orange Belt

"Systems Engineering," J. Byrne; "Coexistence, of the Transistor is Here to Stay," N. J. Krilanovich, Motorola Research Lab.; 1/28/58.

"Contemporary Infra-Red Detecting Equipment," R. W. Powell, Aerojet General Corp.; 2/25/58.

Pasadena

"Stereo-Disc Recording," J. G. Frayne, Westrex Corp.; 2/18/58.

San Fernando Valley

Field Trip of Rocketdyne Test Facility, speakers: P. Vogt, F. Williams, Rocketdyne; 2/12/58.

Santa Barbara

"Gamma Radiation Rate Effects on Semi-Conductor Diodes," J. W. Clark, Hughes Aircraft Co.; 2/11/58.

Come again!
Radio Engineering Show
March 23-26, 1959
New York Coliseum

A Viking Fable

When the terrible green monster suddenly appeared alongside the good ship Viking Queen, all hands save one promptly disappeared over the side into the chill waters of the North Atlantic. Only Lief Smorgasbord, radar operator, remained aboard to face the beast.

If we may take a trembling Lief from history, we will follow the conversation that ensued:

Lief (trembling): Why . . . why didn't you show up on my scope?

Monster (in a high, feminine voice): I'm enchanted, that's why! Oh, Mr. Viking, I'm just a poor princess who has been bewitched and transformed into a teen-age she-sea serpent! If you could answer the Mysterious Riddle, you could break the spell and marry me!

Lief (still trembling): The Mysterious Riddle?

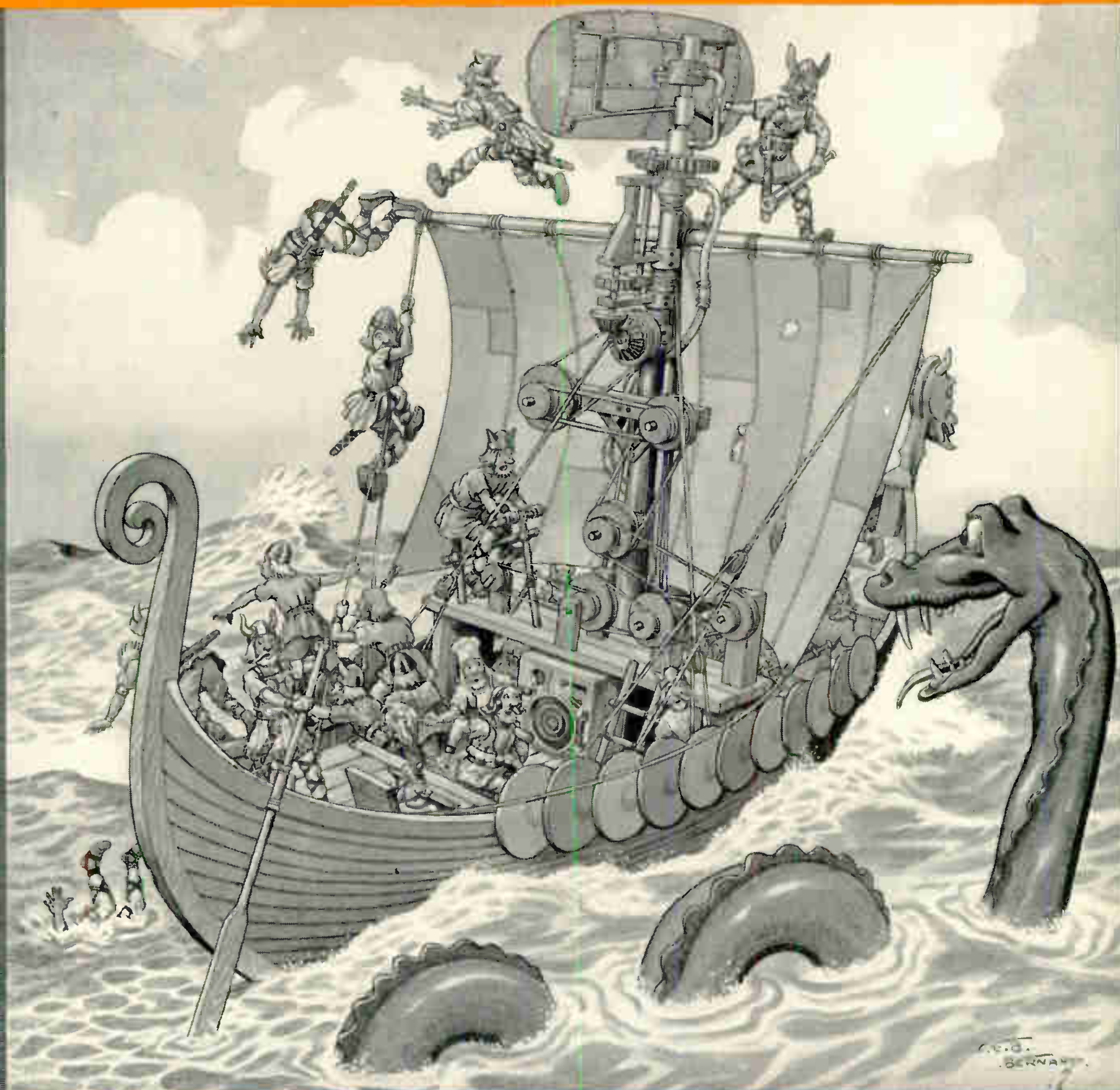
Monster (hopefully): It goes like this.

Heart of that which has no ears, but hears;
No eyes, but sees; no nose, but knows . . .
Tube B or not Tube B, that is the question!

*Lief managed to answer the riddle, breaking the spell and instantly transforming the monster into a lovely princess. And so they were married and lived happily ever after.**

** The single word was "Bomac," of course. Lief knew "Tube B or not Tube B" must refer to Bomac tubes, heart of any radar system ("that which has no ears, but hears, etc.") Smart one, that Smorgasbord.*

No. 5 of a series . . . BOMAC LOOKS AT RADAR THROUGH THE AGES



* Bomac makes the finest microwave tubes and components either side of the Atlantic

WRITE FOR FREE SIX PAGE FOLDER

Bomac LABORATORIES, INC.
Salem Road, Beverly, Massachusetts

Leaders in the design, development and manufacture of TR, ATR, Pre-TR tubes; shutters; reference cavities; hydrogen thyratrons; silicon diodes; magnetrons; klystrons; duplexers; pressurizing windows; noise source tubes; high frequency triode oscillators; surge protectors.

Offices in major cities—Chicago • Kansas City • Los Angeles • Dallas • Dayton • Washington • Seattle • San Francisco • Canada: R-D-R Associates Limited, 1470 Don Mills Road, Don Mills, Ontario • Export: Maurice I. Parisier, 741-745 Washington St., N.Y.C. 14, N.Y.

Industrial Engineering Notes*

EIA Activities

The electronics industry continued its growth during 1957 to become the fifth largest U. S. manufacturing group, the 1958 EIA "Electronics Industry Fact Book" reveals. The 4th edition of this EIA publication has been prepared by the Marketing Data Department to provide a reference guide for industry use. The publication is being sent to all EIA member companies this week and later will be made available to non-members and the general public at 50 cents a copy, Executive Vice-President J. D. Secrest said.

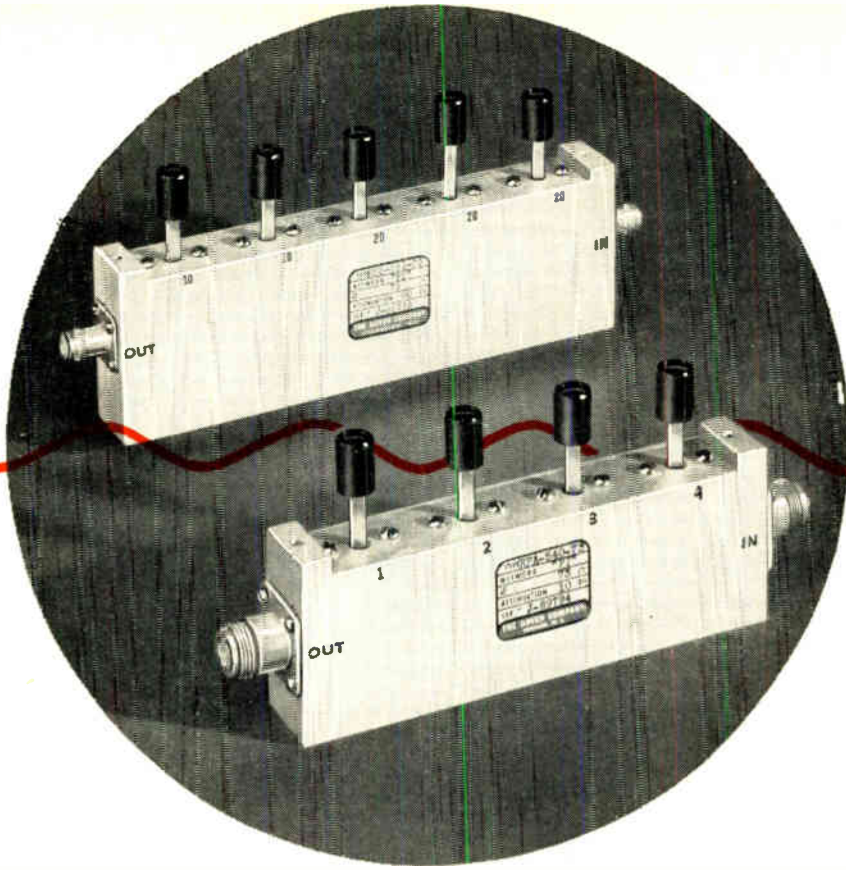
The booklet's preface states that the value of consumer products, tubes, semiconductors, components, and military and industrial equipment reached $7.6 billion in 1957. When distribution and maintenance costs and broadcasting revenues are added, electronics industry sales exceed $13 billion—a twelve-fold expansion during the 12 years since the end of World War II. Continuing, the booklet points out: "An outgrowth of the prewar radio industry, electronics finds its techniques used increasingly by all segments of the economy. Government purchases of military electronic equipment reached an all-time high of $3.9 billion during 1957. Industrial and commercial electronics increased by fifty per cent from $950 million to $1.3 billion. Annual factory sales of consumer products are up from $1.4 billion to $1.5 billion reflecting the resurgence of radio and phonograph popularity. The 15.5 million radios that moved from the production line during 1957 are the greatest number since 1948. Phonographs, particularly "Hi-Fi," plus record player attachment sales passed the five million mark for the first time in the industry's history."

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

Government

The Weather Bureau will use radio-television CONELRAD signal to flash emergency flood and storm warnings, but success of the warning plans will depend largely on the ability of radio manufacturers to include a low-cost automatic alarm feature in future radio receivers, according to F. W. Reichelderfer, Chief of the Bureau. The Weather Bureau chief estimated there are more than 200,000 CONELRAD alert receivers in operation in the U. S. . . . **The FCC's Technical Research Division in the Chief Engineer's office has released a report titled "A Preliminary Analysis of Multicasting" which evaluates the advantages of a possible new approach to overcome difficulties encountered in providing uhf television service in rough terrain from a single transmitter.** The technical information contained in the report may be of special interest to TV engineers. A copy may be obtained from the Technical Research Division, Room 7506, New Post Office Building, Washington 25, D. C., upon individual request. . . . **The Federal Communications Commission was issuing an order to delay the processing of all subscription TV applications until 30 days after Congress adjourns while some Congressional members were offering arguments in favor of pay-TV trials.** Warned by both the House and Senate Interstate and Foreign Commerce Committees not to go through with the plan to accept applications for pay-TV after March 1 until Congress had a chance to act on pending bills affecting pay-TV, the Commission announced it will not process any applications for it until 30 days after Congress adjourns, variously estimated to be sometime in late fall. Meanwhile, Congress could pass legislation which would flatly prohibit pay-TV.

(Continued on page 78A)



FOR ACCURATE ATTENUATION OVER A WIDE FREQUENCY RANGE...

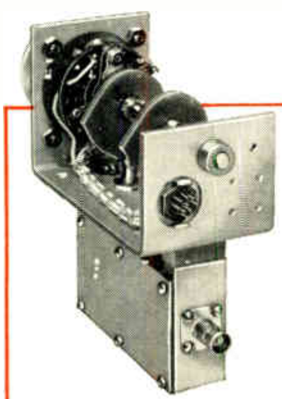
RF Attenuators by DAVEN

These units are used in signal generators, wide-band amplifiers, pulse generators, field intensity meters, micro-wave relay systems, and repeater stations. They find application as laboratory standards, test equipment, and for checking out all types of instruments.

Daven RF Attenuators are available, in combination, with losses up to 120 Db in two Db steps; or 100 Db in one Db steps. Due to their internal circuitry and construction, they have a **zero insertion loss** over the **frequency range from DC to 225 megacycles.**

Standard impedances are 50 and 73 ohms, with special impedances available on request. Resistor accuracy is within ±2% at DC. An unbalanced circuit is used which provides constant input and output impedance. The units are supplied with either UG-58/U or UG-185/U receptacles or Coaxial lead terminations. Individual units with single-section cavities can be obtained.

Many of these types are available for delivery from stock.



Solenoid actuated RF Attenuators are also available in various decibel combinations and any number of steps up to 5.

TYPE	LOSS	TOTAL Db	STANDARD IMPEDANCES
RFA & RFB 540	1, 2, 3, 4 Db	10	50/50Ω and 73/73Ω
RFA & RFB 541	10, 20, 20, 20 Db	70	50/50Ω and 73/73Ω
RFA & RFB 542	2, 4, 6, 8 Db	20	50/50Ω and 73/73Ω
RFA & RFB 543	20, 20, 20, 20 Db	80	50/50Ω and 73/73Ω
RFA & RFB 550	1, 2, 3, 4, 10 Db	20	50/50Ω and 73/73Ω
RFA & RFB 551	10, 10, 20, 20, 20 Db	80	50/50Ω and 73/73Ω
RFA & RFB 552	2, 4, 6, 8, 20 Db	40	50/50Ω and 73/73Ω

Other Db loss combinations are available.

Write for complete information

THE DAVEN CO.

Livingston, N. J.

WORLD'S LARGEST MANUFACTURER OF ATTENUATORS

Industrial Engineering Notes

(*Continued from page 76A*)

Military Electronics

The Office of Technical Services, Commerce Department has made available to industry the following reports and catalogs of technical reports (CTRs): "Research, Problem-Solving, and the Use of Technical Information in Small and Medium Sized Manufacturing Firms," number PB 131578, priced at 75 cents each. The report is the result of a survey of over 500 firms ranging in size from 50 to 1,500 employes. The CTRs recently published of interest to electronic firms, and priced at 10 cents each, are: "CTR-249—Printed Electronic Circuits, 1947–58"; "CTR-250—Electronic Miniaturization, 1944–58"; "CTR-99—Paper Capacitors, 1937–57," and "CTR-336—Transformers, 1927–58.". . . **Though spending by the Military Departments for electronics in the first quarter of the current fiscal year was at a near-record level, second quarter spending showed a substantial increase, according to the latest EIA compilation.** Defense electronics procurement in the quarter ending Dec. 31 totaled $967.5 million, up from the $926 million spent in the first quarter and over the $876 million spent by the military for electronics in the second quarter of fiscal year 1957. Cumulative spending for electronics by the Defense Department during the first six months of the current fiscal year amounted to $1,893.5 million compared with the $1,513 million spent during the like six-month period in fiscal year 1957. . . . **The Signal Corps has announced the development of an extremely bright and high-speed character display tube and its use as the essential display device in the Army's MISSILE MASTER system.** The tube was developed by the Allen B. DuMont Laboratories, Inc., and displays distinctively shaped characters to indicate the various types of information gathered by the complete system relative to enemy aircraft, friendly aircraft, and NIKE batteries in the area, the Signal Corps said. The important features of the tube are that the characters are generated by electronic equipment outside the tube, which means savings in replacement cost. It also produces a far brighter display without the inconvenience of flickers. . . . **Douglas Aircraft Co. has received a $43.5 million contract from Western Electric Co. for fabrication and production of additional units of the Army NIKE-HERCULES missile. . . . Lt. Gen. D. L. Putt, Air Force Deputy Chief of Staff for Research and Development, revealed a new "dyna-soar" research vehicle which he said "will far exceed the performance of the X-15."** The term, he said, is a contraction of the words *dynamic soaring* and means that utilization of forces employed for flight are both centrifugal and aerodynamic. The facts were told to House Armed Services Committee members recently. The proposed research vehicle, he said, "is both small and

(*Continued on page 80A*)

Now, Tung-Sol offers designers a complete line of high reliability Germanium PNP Transistors!

EIA TO-9
Outline Package

EIA TO-3
Outline Package

Tung-Sol types and ratings are listed below with the types they replace. From these, spot your needs! Then, for long-life operation, specify Tung-Sol!

TYPE	APPLICATION	MAXIMUM RATINGS (25° C)					TYPICAL VALUES (25° C)				SIMILAR TYPE REFERENCE
MEDIUM POWER AUDIO TYPES (To-9 Outline)		Pc mw	Vce volts	Vcb volts	Ic ma	Tj °C	MAX. Icbo μa	Hfe	fab mc	Ge db	
2N381	Output Amplifier	200	25	25	200	85	20	36	1.2	31	2N61, 2N186/A, 2N187/A, 2N266
2N382	Output Amplifier	200	25	25	200	85	20	54	1.5	33	2N60, 2N180, 2N181, 2N185, 2N188/A, 2N226, 2N311, 2N403, 2N408
2N383	Output Amplifier	200	25	25	200	85	20	72	1.8	35	2N59, 2N224, 2N241/A, 2N265, 2N270
2N460	General Purpose Industrial	200	—	45	400	100	15	25	1.5	39	2N44
2N461	General Purpose Industrial	200	—	45	400	100	15	50	1.5	41	2N43
HIGH FREQUENCY TYPES (To-9 Outline)											
2N404	Computer	100	24	30	400	85	5	30	12	—	2N581
2N425	Computer	100	20	30	400	85	5	30	4	—	2N394, 2N578
2N426	Computer	100	18	25	400	85	5	40	6	—	2N269, 2N395, 2N579
2N427	Computer	100	15	20	400	85	5	55	11	—	2N123, 2N315, 2N396, 2N580
2N428	Computer	100	12	15	400	85	5	80	17	—	2N316, 2N397, 2N582
2N413	RF Amplifier	100	15	—	200	85	5	—	3	—	2N111, 2N135, 2N410
2N414	RF Amplifier	100	15	—	200	85	5	—	5	—	2N139, 2N112, 2N136, 2N218, 2N412
2N416	RF Amplifier	100	15	—	200	85	5	—	10	—	2N113
2N417	RF Amplifier	100	15	—	200	85	5	—	20	—	2N114
HIGH POWER TYPES (To-3 Outline)		Pc w	Vce volts	Vcb volts	Ic Amps.	Tj °C	MAX. Icbo ma	Hfe	fab mc	Ge db	
2N242	Audio Amplifier	15	45	—	2	85	1.0	50	0.4	34	2N155, 2N176, 2N250, 2N257, 2N301/A, 2N350, 2N351, 2N554, 2N555
2N378	Power Switch	15	20	40	3	85	0.5	35	0.3	24	2N255
2N379	Power Switch	15	40	80	3	85	0.5	30	0.3	23	2N158/A, 2N251, 2N296, 2N297
2N380	Power Switch	15	30	60	3	85	0.5	60	0.4	29	2N156, 2N256, 2N387
2N459	Power Switch	15	60	105	3	85	0.5	40	0.3	24	2N375

NOTE: *Similar type references are listed at time of printing and should be interpreted as approximate equivalents. This reference does not necessarily imply exact electrical or mechanical interchangeability.*

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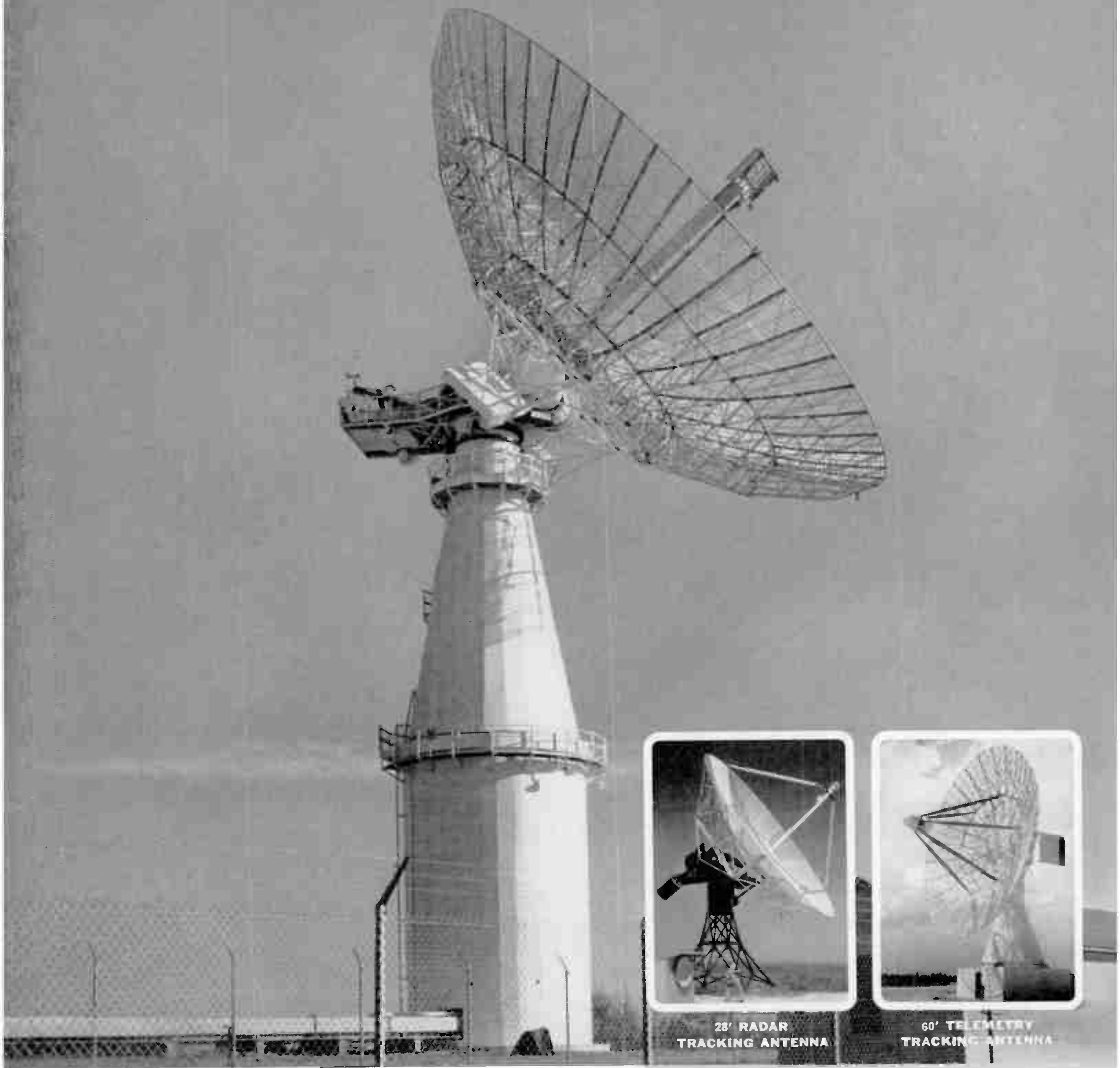
(Continued on page 86A)

Big dish with a big job . . .

NEW 84′ KENNEDY TRACKING ANTENNA



TOWERING 90 feet above the New England countryside at Westford, Mass., this giant 84′ tracking antenna is part of a new, long-range radar installation now studying problems in ballistics missile defense.

Equipped with an elevation-azimuth type mount designed and fabricated in cooperation with M. I. T.'s Lincoln Laboratory, the big dish can make a full 360° horizontal sweep and has a vertical rotating capability of 90°. Like all Kennedy steerable antennas, it features a light weight, aluminum dish supported by a steel pedestal mounted on a concrete base.

This kind of achievement in antenna design and construction is solid proof that Kennedy is the name to remember when you are faced with antenna problems.

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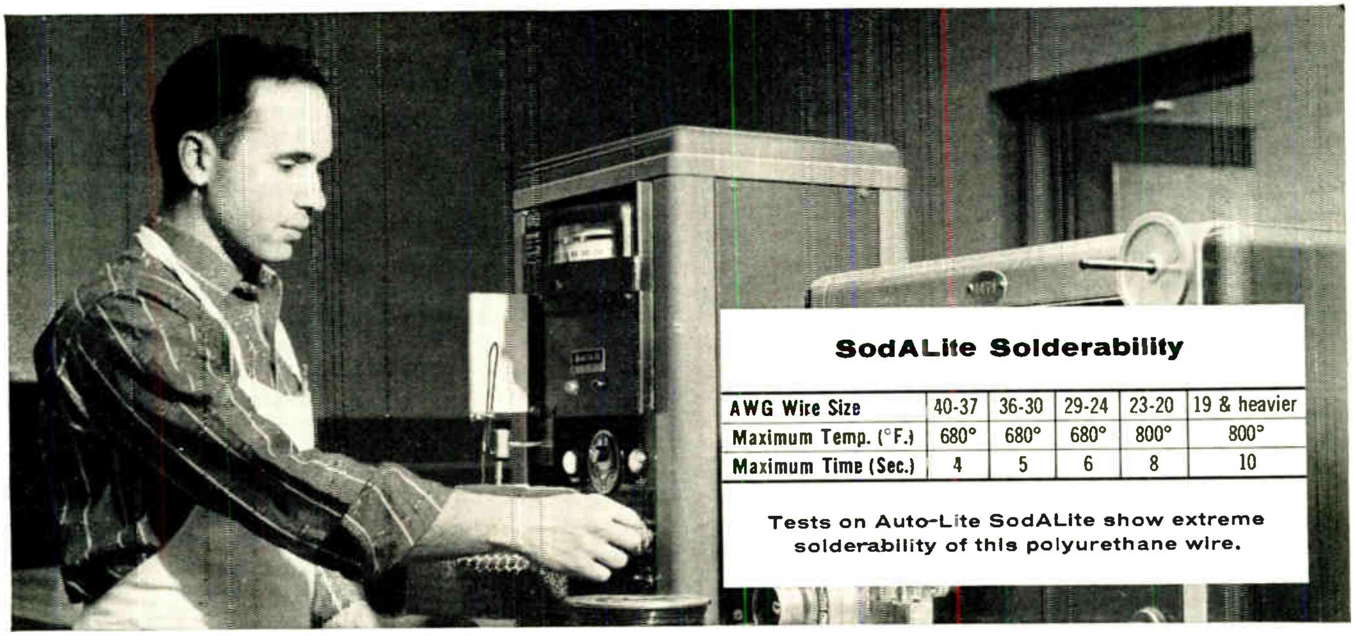
SodALite means Solderability!



SodALite Solderability

AWG Wire Size	40-37	36-30	29-24	23-20	19 & heavier
Maximum Temp. (°F.)	680°	680°	680°	800°	800°
Maximum Time (Sec.)	4	5	6	8	10

Tests on Auto-Lite SodALite show extreme solderability of this polyurethane wire.

Here's a polyurethane base magnet wire insulation that's self-fluxing, outstandingly easy to solder!

Fine sizes of SodALite magnet wire can be soldered at 680° F. in approximately 5 seconds. Heavy wire sizes can be soldered at 800° F. in approximately 10 seconds.

Conventional dip, iron, torch or gun methods will produce excellent connections. There is no need for brushing or chemical stripping because SodALite vaporizes to produce a clean, solderable surface. Although SodALite is self-fluxing, some operations may require a non-corrosive flux for best results. Excessively high temperatures will delay soldering and may cause poor connections.

SodALite has higher dielectric strength values when compared with other standard films. Tests indicate only a small drop in dielectric strength after immersion in water at room temperature. High frequency characteristics and corona resistance, even in humid conditions, exceed nylon insulations. SodALite is compatible with a variety of phenolic alkyd, silicone and polyester impregnating varnishes. Field reports show it equal to other popular wires in abrasion resistance and handling characteristics.

SodALite has excellent physical characteristics and electrical properties in addition to good resistance to solvents, moisture, acids, and bases. SodALite has unusual thermal properties and, when tested to method of AIEE #57, has 10-15° C. higher thermal rating than other widely used Class A insulations.

SodALite is offered as a 105° C. magnet wire, or better. Higher temperature usage should be considered only after testing to the specific applications, because polyurethanes such as SodALite cannot withstand excessive overload conditions. For moderate overload conditions SodALite may be considered for use up to 120° C.

Availability: Single, heavy, triple and quadruple films in round AWG #8 through #40.

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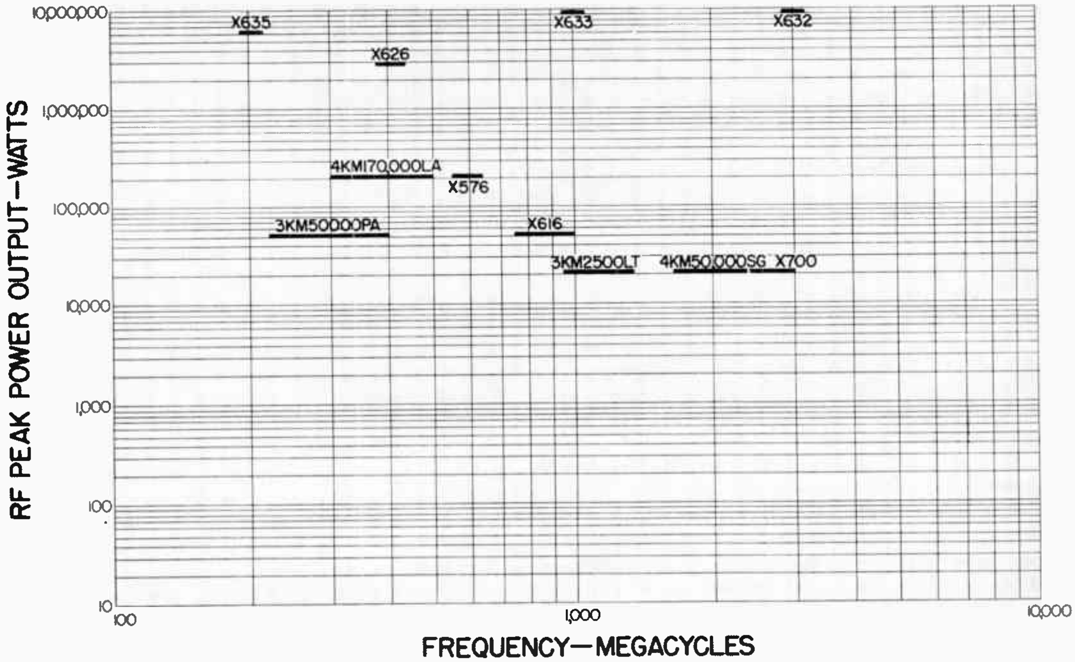
(Continued on page 89A)

Eimac First for Power Amplifier Klystrons

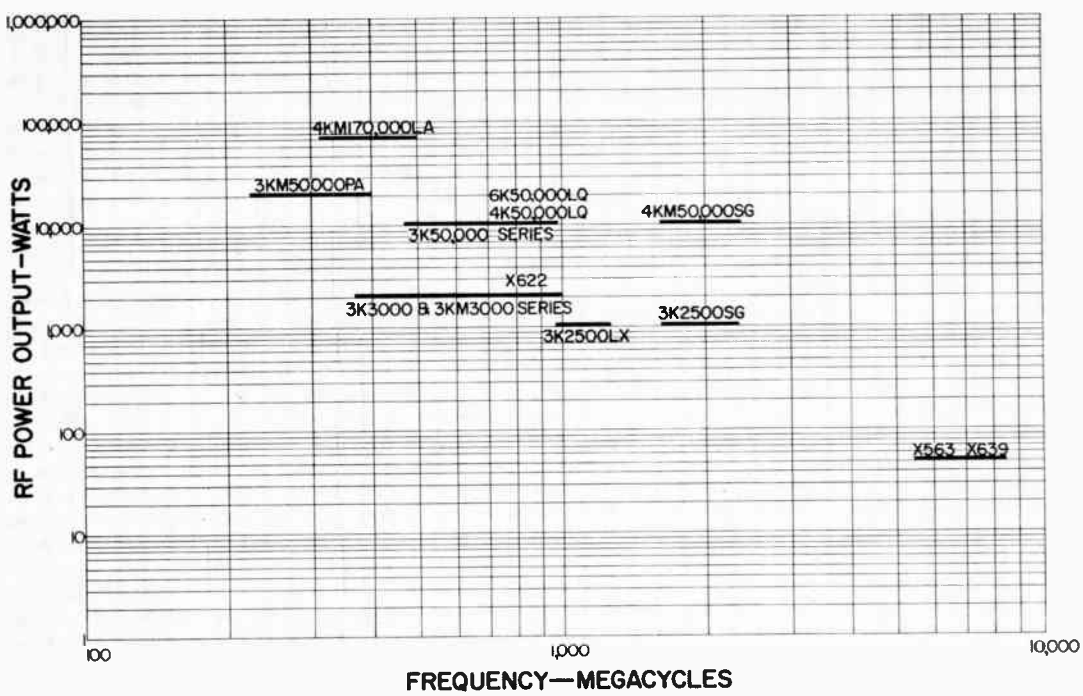
PULSE AMPLIFIER KLYSTRONS



FIGURE

. . . from VHF into SHF

CW AMPLIFIER KLYSTRONS



FIGURE

The broad frequency coverage and wide range of power levels now offered by Eimac amplifier klystrons is shown in the accompanying charts. Frequency coverage extends into the SHF range and multi-megawatt pulse output powers are available.

The exceptional ability of Eimac amplifier klystrons to conveniently and reliably generate high RF power at ultra-high and super-high frequencies has led to their widespread use in tropo-scatter communications, high power radar, UHF television broadcasting, linear accelerators and many other applications.

For more detailed information on Eimac's reliable, simplified approach to high power at high frequencies, write for a copy of Klystron Facts Case Five. The Eimac Application Engineering Department will gladly assist you in planning new systems using Eimac power klystrons.

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New trends and developments in designing electrical products . . .

P-6 . . . A special General Electric magnetic alloy with high hysteresis loss and torque characteristics

P-6—a cobalt-nickel-vanadium-iron alloy developed by G-E research engineers — possesses a unique combination of high residual induction and low coercive force.

These properties make it ideal for applications where high residual induction is required without the excessive magnetizing forces necessary with other magnetic materials such as the high cobalt steels, the cobalt-nickel-iron alloys, the cobalt-vanadium-iron alloys, and the alnicos.

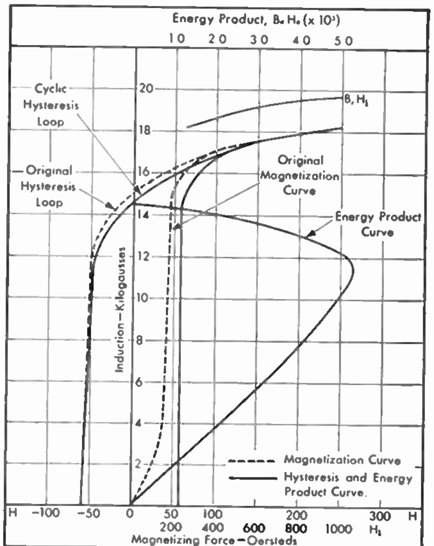

FIG. 1—Hysteresis loop, energy product, and magnetization curves for P-6 alloy.

The residual induction of P-6 is 14,000 gausses . . . over 30% greater than either 17% cobalt-steel alloy or 6% tungsten-steel alloy. The maximum energy product of P-6 is 410,000 gauss-oersteds. Figure 1 shows the magnetization, hysteresis loop, and energy product curves for P-6 alloy.

The P-6 alloy is also useful where ease of workability of the magnetic material is desired. Because it is aged at a high temperature, it can withstand greater operating temperatures than tungsten-steels and cobalt-steels without undergoing subsequent aging.

One of the major uses for P-6 is in rotors* for hysteresis motors (fig. 2). These motors exhibit high starting torques and can synchronize high inertia loads without auxiliary starting equipment. Most important, the

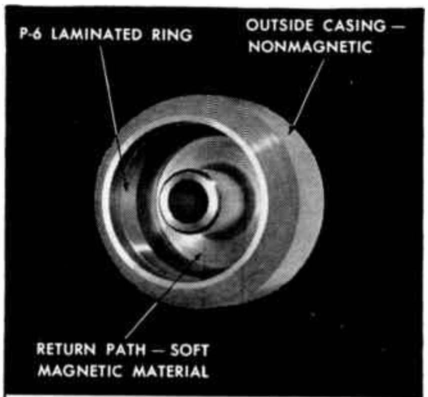

FIG. 2—Rotor for gyroscope motor uses laminated P-6 hysteresis ring.

torque produced is constant and doesn't fluctuate with rotational speed.

In this type of motor, there is a hysteresis loss in the rotor material caused by the revolving magnetic field. This loss produces torque between the permanent magnet rotor and the spatially revolving magnetic field.

Because the theoretical torque produced is directly proportional to the hysteresis loss in the rotor, it is advantageous to obtain the highest possible value of hysteresis loss in the rotor material.

To do this, a material must possess a high hysteresis loss for a given applied magnetizing force. Figure 3 shows a comparison between P-6 and other magnetic materials. Of the materials tested, P-6 best fulfills the desired characteristics.

Another use for P-6 is in hysteresis clutches. This type of clutch is formed when the wound field in the hysteresis motor is replaced by a rotating permanent magnet member to provide field excitation.

Although hysteresis clutches are larger than friction or magnetic particle clutches of the same torque rating, they have many advantages such as high degree of reliability, repeatability, linearity, and freedom from excessive drag torque.

General Electric P-6 alloy is available in strips from .010″ to .100″ thick, and in widths up to four inches. In wire form, it is available in .0201″ to .102″ diameters. P-6 should be capable of being swaged, welded, extruded and drawn. General Electric Engineers currently are experimenting with these forming operations.

The development of P-6 alloy by General Electric is one of the many examples of how G.E.'s research in magnetic materials is paying off. The same experience, skill and facilities that made this development possible can be put to work solving your magnet problems.

To get the expert design assistance of G-E Magnet Engineers, or your copy of the new G-E Magnet Design Manual, simply write: *Magnetic Materials Section, General Electric Company, 7808 N. Neff Blvd., Edmore, Michigan.*

Material	Peak H Oersteds	Peak B Gausses	Coercive Force H_c—Oersteds	Residual Induction B_r—Gausses	Loop Area Gauss-Oersteds
P-6 Alloy	70.0	15,500	51.7	13,400	2,860,000
	65.0	15,000	51.0	12,900	2,720,000
	60.0	14,500	50.8	12,200	2,540,000
	55.0	10,900	45.2	8,500	1,610,000
5.75% Chrome Steel	50.0	4,100	25.0	2,100	290,000
	70.0	6,900	38.0	4,600	1,210,000
	90.0	8,700	44.0	6,100	1,700,000
17% Cobalt Steel	118.0	12,500	68.7	9,250	2,872,500
	91.4	10,000	61.9	6,950	1,885,000
	79.2	7,500	52.1	4,540	1,005,000
	69.6	5,000	35.0	2,315	446,000
6% Tungsten Magnet Steel	100.0	12,450	28.0	9,150	1,297,000
	61.5	10,000	23.2	7,200	774,000
	50.0	8,500	20.5	6,050	624,000
	45.5	8,000	18.0	5,650	499,000

FIG. 3—Comparison of P-6 and three typical magnetic alloys.

*Gyro spin motor rotor courtesy of Minneapolis-Honeywell Regulator Company

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Chazen, J., West End, N. J.
Clark, W. H., Jr., Dayton, Ohio
Clarke, W. E., Belleville, Ont., Canada
Coker, Q. W., San Antonio, Tex.

(Continued on page 90A)

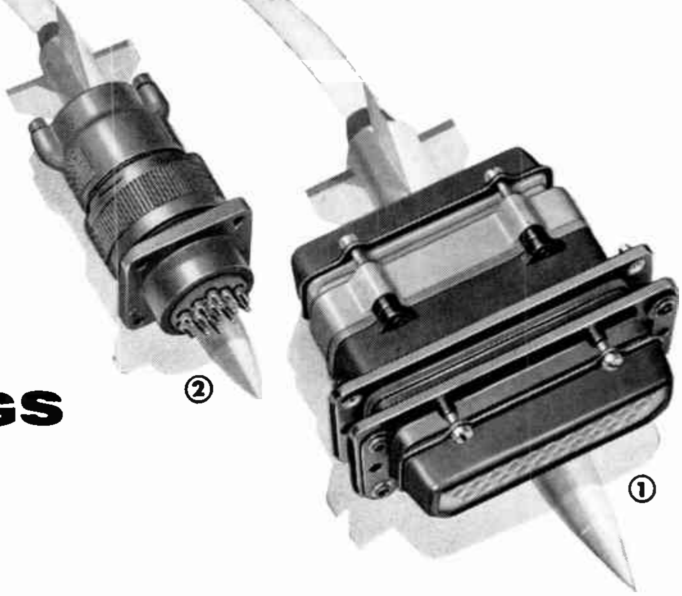
Membership

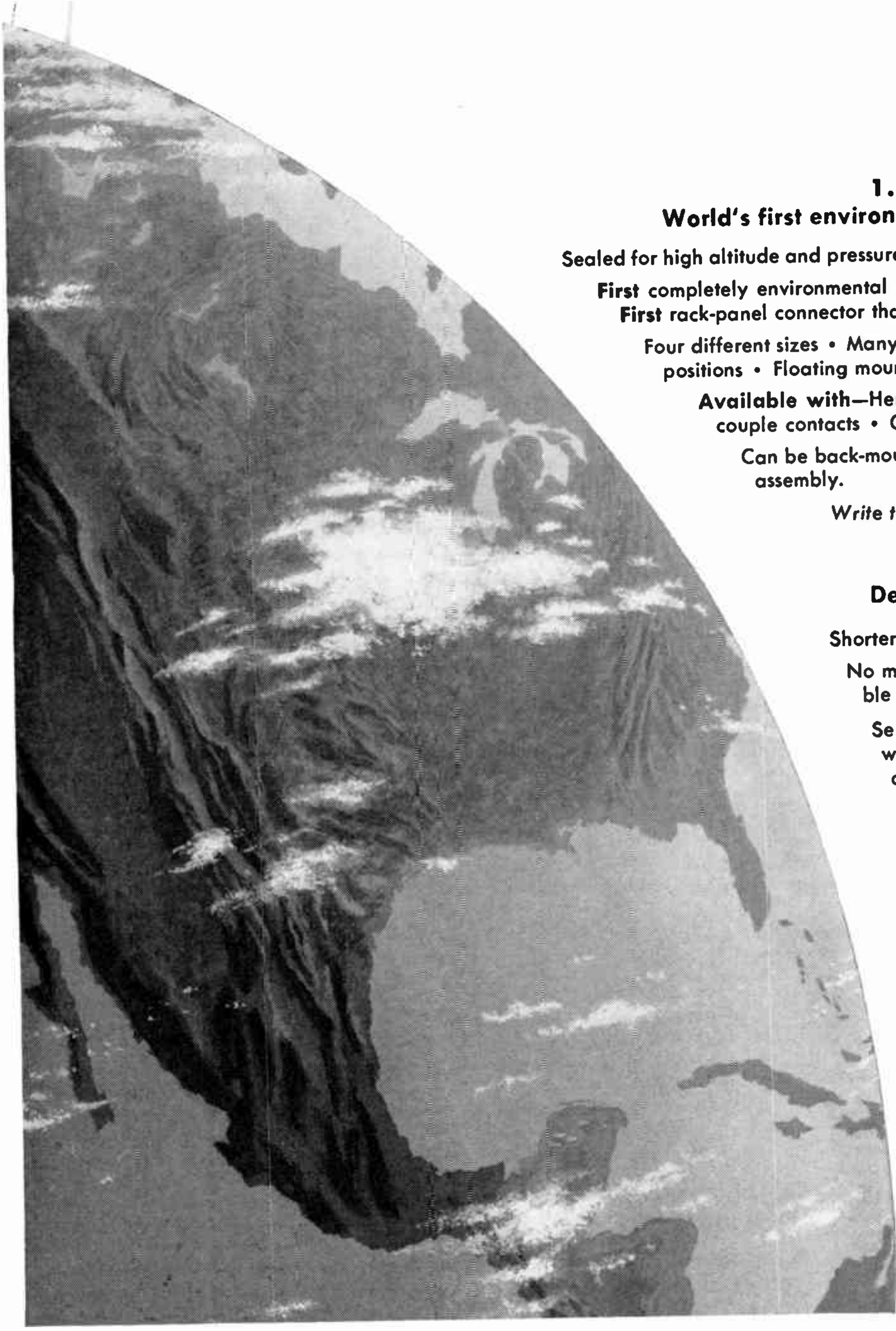
(Continued from page 89A)

Coleman, D. T., Jr., Baltimore, Md.
Colin, R. I., Nutley, N. J.
Collins, R. F., Torrance, Calif.
Collmer, R. C., Tempe, Ariz.
Conroy, T. R., Long Branch, N. J.
Conti, E. A., Redwood City, Calif.
Corbett, H. L., Jr., Portland, Ore.
Crosby, J. C., Arcadia, Calif.
Crowley, T. J., New York, N. Y.
Cubbage, J. F., Phoenix, Ariz.
Daugherty, L. C., Oklahoma City, Okla.
Davidson, K. W., Richardson, Tex.
Davis, A. J., Hinsdale, Mass.
Davis, C. F., Jr., Brighton, Mass.
Davis, E. S., Arlington, Mass.
Davis, H. H., Ontario, Calif.
Davis, W. B., Toronto, Ont., Canada
Day, B. E., Scotia, N. Y.
de Haas, T., Montreal, Que., Canada
Dell'Acqua, L., Nutley, N. J.
Demo, R. D., Liverpool, N. Y.
DeMyer, D. L., Flint, Mich.
den Brinker, C. S., Taplow nr. Maindenhead (Berks.), Great Britain
Denny, J. H., Syracuse, N. Y.
Desautels, G. L., Haddonfield, N. J.
Dewire, G. M., Syracuse, N. Y.
Divilbiss, J. L., Champaign, Ill.
Doelger, R. J., Jr., Shrewsbury, N. J.
Dowden, W. M., Montoursville, Pa.
Drumeller, C. C., Oklahoma City, Okla.
DuBose, T. I., Milwaukee, Wis.
Dumert, V., London, England
Eisenberg, M., Philadelphia, Pa.
Ekstrom, J. L., Baltimore, Md.
Ellis, A., Manhattan Beach, Calif.
Engler, C. D., Rochester, N. Y.
Enz, R. E., Torrance, Calif.
Fagan, B. B., Arlington, Va.
Feaster, J. A., South Bend, Ind.
Fiol, L. E., Mayaguez, Puerto Rico
Fitta, E. R., Belleville, Ill.
Flaherty, R. J., Poughkeepsie, N. Y.
Flattau, W., Elmhurst, L. I., N. Y.
Forsberg, A. W., Amityville, L. I., N. Y.
Fossler, K. L., Houston, Tex.
Fowler, R. H., Westhampton Beach, L. I., N. Y.
Fowler, T. C., Ridgecrest, Calif.
Fox, R. M., Webster, N. Y.
Freeman, P. S., Rosedale, L. I., N. Y.
Friddell, H. G., Zenith, Wash.
Fuchs, W. C., Libertyville, Ill.
Ganter, C. F., Syracuse, N. Y.
Gass, C. A., Grandview, Mo.
Gay, R. E., Mt. Pleasant, S. C.
Gillespie, C. M., Jr., Tucson, Ariz.
Gilliam, T. J., Puente, Calif.
Gilmore, R. P., Bellaire, Tex.
Golec, E. J., White Plains, N. Y.
Goodman, R. L., Inglewood, Calif.
Greaney, E. J., Long Beach, Calif.
Greear, D. W., Alexandria, Va.
Green, F. D., Saskatoon, Sask., Canada
Greenbaum, M., New York, N. Y.
Greenwood, S. W., Bradford, Yorks, England
Grieco, L., Philadelphia, Pa.
Grigsby, L. L., Lubbock, Tex.
Goldberg, J. J., Van Nuys, Calif.
Grim, E. D., Merchantville, N. J.
Grimes, W. H., Dayton, Ohio
Guditz, E. A., Lexington, Mass.
Gunby, G. C., Seattle, Wash.
Gunderson, D. C., Whippany, N. J.
Haas, C. J., Columbus, Ohio
Hagerett, R. E., Whippany, N. J.
Haines, G. L., Cedar Rapids, Iowa
Halpin, A. T., Jr., Massapequa Park, L. I., N. Y.
Hand, J. E., Indianapolis, Ind.
Hankey, I. L., Frederick, Md.
Hansen, R. E., Fayetteville, N. Y.

(Continued on page 102A)

2
great new environmental
CANNON PLUGS





1. The DPSM
World's first environmental magnesium connector

Sealed for high altitude and pressure applications... exceptionally light!

First completely environmental rack-panel connector

First rack-panel connector that can be used as cable connector

Four different sizes • Many insert arrangements • Polarized—4 alternate positions • Floating mounts • Moisture proof

Available with—Hermetic sealing • Co-axial contacts • Thermocouple contacts • Crimped contacts • Air lines

Can be back-mounted—permitting bench fabrication of harness assembly.

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Designed for commercial jet airliners

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No moisture trap • Moisture proof • Easy to assemble • Vibration proof

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SERVO MOTOR TACH GENERATORS

to your precise specification

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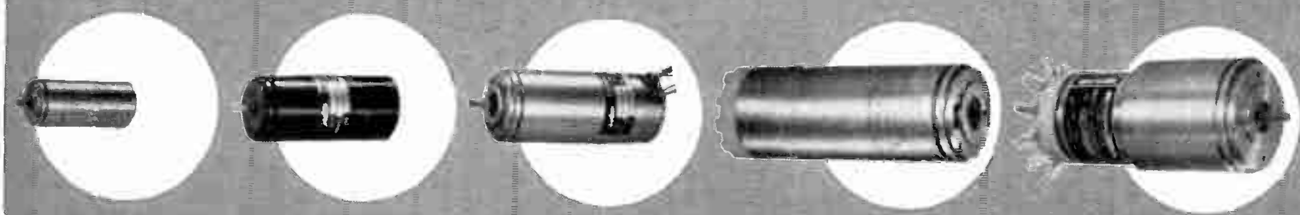
- Both Damping and Integrating types available with parameters to your requirement.
- Complete size range: 8, 10, 11, 15, 18. Can be designed with gear train.
- —54°C to +125°C ambient temperature range.
- Designed to MIL-E-5272.
- Assembled under closely controlled environmental conditions.

Burton Browne Advertising



400 CPS SERVO MOTOR — TACHOMETER GENERATORS

Oster Type	Size	Length Inches	Wt. Oz.	Rotor Inertia gm cm²	MOTOR: Rated Voltage Ø 1	MOTOR: Rated Voltage Ø 2	MOTOR: No load speed RPM	MOTOR: Watts per phase @ Stall	MOTOR: Stall Torque OZ. IN.	GENERATOR: Excit. Volt.	GENERATOR: Output Volt. per 1000 RPM	GENERATOR: Lin.% to 3600 RPM	GENERATOR: Null MV	GENERATOR: Phase Shift @ 25°C
8MTG-6201-01	8	1.850	2.3	0.77	26	40/20	6,500	2.2	0.16	26	0.25	0.5	15	± 5°
*10MTG-6228-02	10	2.157	4.2	0.72	115	115/57.5	9,500	2.8	0.26	115	0 45	1.5	19	± 10°
10MTG-6229-12	10	2.100	2.9	1.09	33/16.5	52/26	9,500	3.0	0.28	26	0 45	1.5	13	± 10°
*10MTG-6229-03	10	2.100	2.9	1.09	26	26	10,500	3.0	0.26	18	0.3	1.5	12	± 10°
10MTG-6229-15	10	2.100	2.9	1.09	26	26	10,500	3.0	0.26	26	0.3	1.5	12	± 10°
*10MTG-6232-05	10	2.104	4.2	1.1	115	36/18	6,500	3.5	0.26	115	0 30	1.5	15	± 10°
11MTG-6251-13	11	2.531	7.0	1.3	115	115/57.5	6,500	3.5	0.63	115	0.55	0.5	19	± 10°
11MTG-6251-00	11	2.531	7.0	1.1	115	40/20	6,500	3.5	0.63	115	0 55	1 5	19	± 10°
11MTG-6254-01	11	2.200	6.0	1.1	115	115/57.5	6,500	3.5	0.63	115	0.55	1.5	19	± 10°
15MTG-6280-01	15	3.281	14.0	5.3	115	115/57.5	5,000	6.2	1.5	115	3.0	0.2	13	± 5°
†*15MTG-6276-03	15	3.875	15.0	4.4	115	57.5	8,500	5.8	0.70	115	2.75	0.2	13	± 0.5°
18MTG-6302-02	18	3.680	20.0	5.7	115	115/57.5	9,000	16.0	2.7	115	3.0	0.2	13	± 5°
18MTG-6302-04	18	3.680	20.0	5.7	115	115/57.5	4,800	9.2	2.4	115	3.0	0.2	13	± 5°

*These units designed for 85°C ambient but same characteristics can be designed for 125°C. †Additional 21.4 watts for heater, the values given are independent of ambient temperature.

Other products include servos, synchros, resolvers, motor-gear-trains, AC drive motors, DC motors, servo mechanism assemblies, reference and tachometer generators, servo torque units, actuators and motor driven blower and fan assemblies.

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Engineers For Advanced Projects: Interesting, varied work on designing transistor circuits and servo mechanisms. Contact Mr. Robert Burns, Personnel Manager, in confidence.

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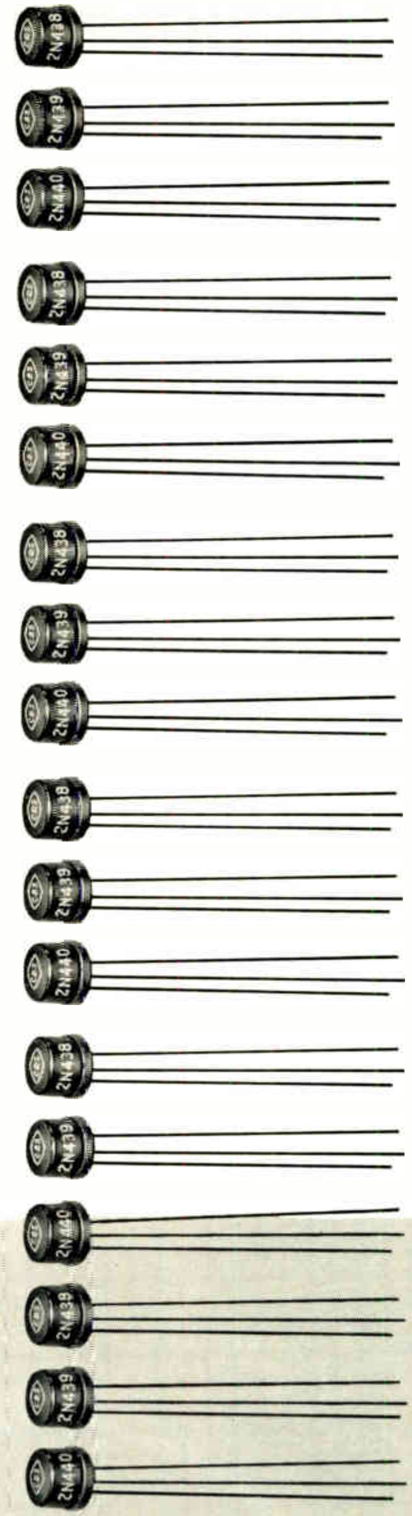
For high-speed switching
For high-frequency amplification

Production of CBS high-frequency transistors has been stepped up for a growing variety of applications. The 2N438, 2N439 and 2N440 are now available in quantity for multivibrator and blocking oscillators . . . gate and flip-flop circuits . . . and r-f, i-f, and high-fidelity a-f amplifiers.

All three types are alloy-junction for greater uniformity, higher voltage and current, flatter gain and lower saturation resistance. They employ the standard JETEC TO-9 package welded to achieve reliability never before approached with NPN transistors.

Note the many desirable features. You may be able to use these CBS transistors. Write for Bulletin E-268 giving complete data and helpful application notes.



CHECK THESE FEATURES

1. High frequency response:
   2N438 .... 2.5 to 5 mc.
   2N439 .... 5 to 10 mc.
   2N440 .... 10 mc and up.
2. High operating voltage .... up to 30 volts.
3. High switching speed .... below 0.2 μsec.
4. High current amp. factor .... up to 100.
5. High dissipation rating .... up to 100 mw.
6. Low leakage current .... 2 μamps av.
7. Low collector capacitance .... 10 μμf.

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CBS-HYTRON
Semiconductor Operations, Lowell, Mass.
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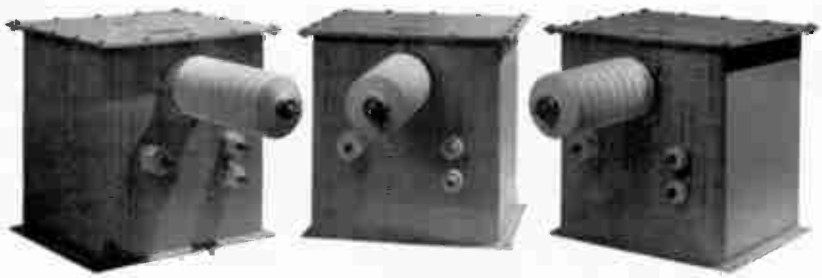
HOW LEVINTHAL CAN HELP



yours for a favorable reactance

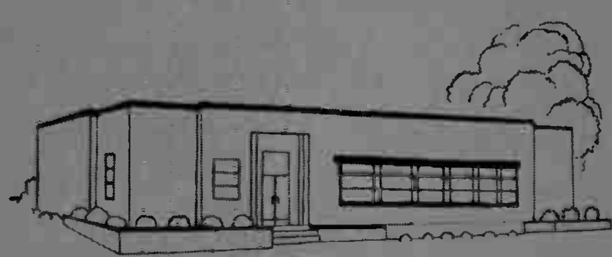
This communication notes the entrance of Levinthal into a new branch of the equipment-engineering field – pulse transformers and other reactance units.

The examples shown are in no way typical. They just happen to have been among the first group produced. Pulse transformers for spark-gap crowbar application to 100 kv, they are rated for 135-kv secondary peak voltage, 0.5-μsec max pulse width, and 0.002 max duty.

If you need pulse transformers, including those to MIL-T27A, we'd like to hear from you, and if you indicate an interest in crowbars, we'll send you a specification sheet. Please note, however, that the bottleneck in our production is the baking oven. We can't make anything larger than 3½ x 5½ x 8 ft.

Specialists in the unusual, Levinthal engineers comprise a highly qualified staff for providing solutions to the most difficult r-f power-handling problems.

Among the transmitters, modulators, power supplies, and accessories for radar, communications, and tube development which have already been produced, there may be units suitable for the solution of your problems. Let us send you a descriptive folder on these or discuss with you the design, development, and production of specialized equipment for your particular needs.



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STANFORD INDUSTRIAL PARK PALO ALTO 4, CALIFORNIA

EQUIPMENT DIVISION

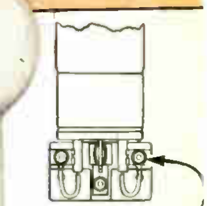
THREE *steps forward in rectifier design by* Amperex®

designers and manufacturers of the world's most complete line of rectifiers

1

THE NEW 7136 single-anode, high-voltage mercury-vapor rectifier. A plug-in replacement for the 575A, surpassing it in capacity and dependability, with higher peak inverse voltage. Offers trouble-free operation in induction and dielectric heaters.

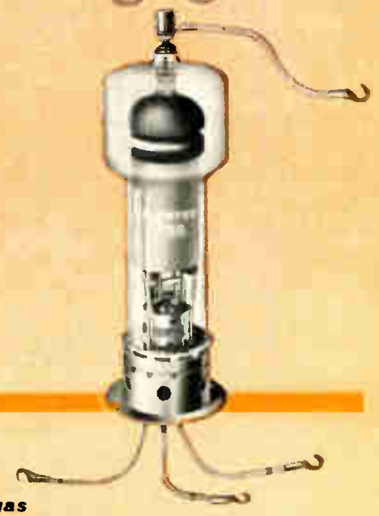
Peak Inverse Voltage: 15 kv
Average Anode Current: 3 amps



2

THE NEW 869 BL heavy-duty mercury-vapor rectifier. Uses short, flexible filament leads to eliminate possibility of high contact resistance. This prevents under-emission and tube damage, and insures proper voltage drop across the filament at all times. Preferred by equipment manufacturers and users for broadcasting, induction and dielectric heating equipment.

Peak Inverse Voltage: 15 kv
Average Anode Current: 5 amps



3

THE 6786 super-power, grid-controlled, mercury-vapor rectifier. Guarantees precise electronic control for industrial oscillators up to and in excess of 100 kw output. Proven long life in actual field tests.

Peak Inverse Voltage: 15 kv
Average Anode Current: 10 amps continuous; 15 amps intermittent

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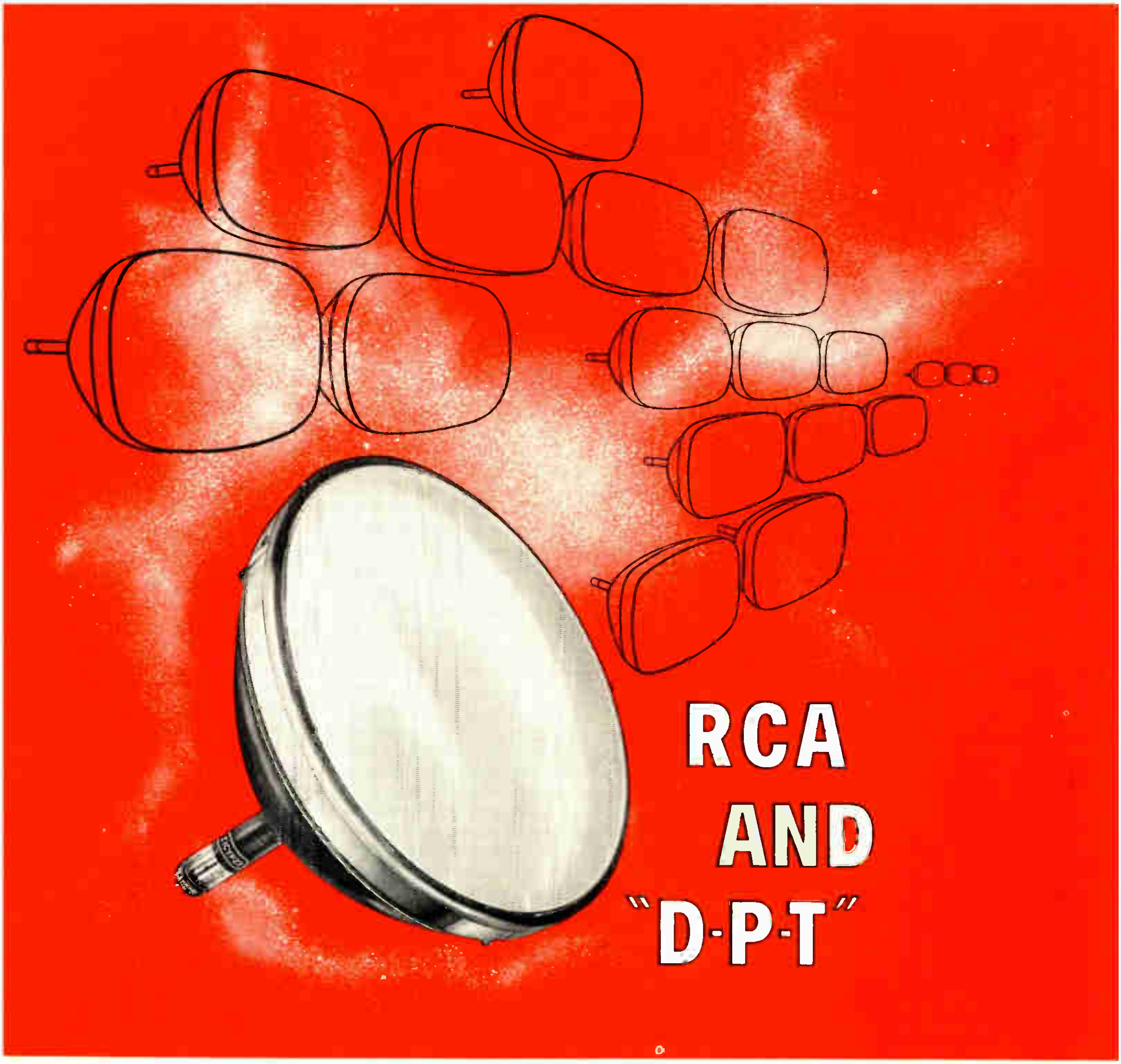
In Canada: Rogers Electronic Tubes & Components, 11-19 Brentcliffe Road, Leaside, Toronto 17, Ont.

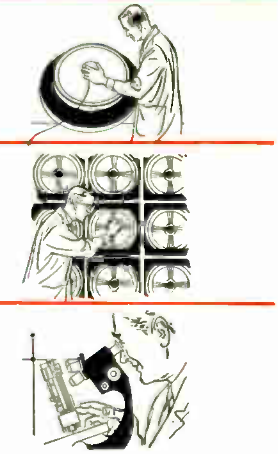


. . . means Dependable-Picture-Tubes for your color and black-and-white TV set designs!

More than a decade ago...RCA took the lead in the design and development of superior quality, dependable picture tubes. And...through intensive concentration on superior quality...has maintained industry leadership.

For example, RCA's 21CYP22 is the only *commercially available* picture tube for color-TV. Featuring an all-glass envelope, the RCA-21CYP22 offers improved color-purity and registration, life-like flesh-tones, dynamic black-and-white performance, and simplicity of adjustment on the TV-set production line. RCA's complete line of picture tubes provides any and all types required by equipment designers, in any price bracket.

RCA's picture-tube production is carefully monitored by continuously operating 1000 hour life-test racks, under "accelerated life" conditions. 2000 hour life-tests under simulated "home conditions", inspections for uniformity of aluminum and phosphor coatings, screen brilliance, resolution, spot size, and contrast. Such rigid, self-imposed, RCA quality controls gives you maximum assurance of "D-P-T"—*Dependable Picture Tubes for your TV-set designs!* For the latest sales information on RCA picture tubes for color or black-and-white, contact your RCA Field Representative.



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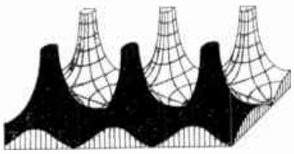
RADIO CORPORATION OF AMERICA

Electron Tube Division — **Harrison, N. J.**

May, 1958
Vol. 46 No. 5

Proceedings of the IRE

Poles and Zeros



The Editorial Survey. The results of last fall's poll of Section and member opinion on the PROCEEDINGS and other IRE publications, summarized on page 822, have been a subject of recent discussion by the Editorial Board. The first reaction has already appeared in the changed location of the PROCEEDINGS Contents page, and other changes are being considered.

The Editorial Board has interpreted the Survey results as a directive to increase its efforts in the publication of tutorial and review papers. Such papers are not easy to obtain and usually must be the result of concerted solicitation by the Editorial Department; they do not often come painlessly through the mail slot. Of this work, a difficult point is to decide on just what topics you members feel the educative urge. The survey did yield a lengthy list of titles for such papers and a program is being undertaken to locate suitable authors and impress upon them the importance of the paper—as well as its shipping date to us.

We feel that the Survey was largely interpreted by our members as applying to the current publications, and probably does not reflect in any major manner the research use of past issues in the libraries. With the PROCEEDINGS directed to publish articles of major significance as well as readable articles, it becomes difficult to judge a paper in terms of the reading ability of our audience ten years hence. Who among us understood the first solid-state papers of ten years ago—yet today, holes, donors, minority carriers, intrinsic conduction, are commonplace. Do we make our point?

Space given to the various departments was generally confirmed by member opinion, although some changes in manner of presentation and treatment are under consideration. In this connection it has long been the Editor's opinion that further emphasis might be given to the meetings of the Board of Directors, so you members might learn that they are just people, after all, rather than some abstract group meeting on Cloud Nine. The fact that the Board, as a matter of policy, is now holding more of its meetings in connection with regional conferences, seems to confirm our view.

Present reports of Board actions filter down through Section officers, or by press releases covering election or award results. After the next meeting of the Board an attempt will be made to write a brief story covering the meeting highlights, without reading like a set of minutes. In particular we do not intend to report on who moved to buy which vendor's red ink, first because it is not important (the motion, that is), and second because this is a detail of Headquarter's operation with which we do not clutter the Board agenda up—if our meaning is clear. It is hoped, however, to let our members know when major changes in policy are discussed and adopted.

Down-Under Land. Continuing our travelogue of last month, it probably should be recorded that in February we had an opportunity to visit our counterpart organization, IRE Australia. This brother organization, which celebrated its 25th anniversary last October, is composed of another bunch of good guys just like we are. Having sat at lunch in Sydney with their Council, it was our feeling that should their group of twenty-four sit down at lunch with our Board of Directors, you would be unable to separate the twain, except possibly by phonic analysis. While the pronunciation may differ, and we had some difficulty understanding that when "the mile came at ten," it was merely the arrival of a letter, yet the basic problems and operations of our two organizations are identical and so are our governing boards.

A society of 2000 members and 8 Sections, they appear about to begin a period of expansion, and thus are probably following U. S. pattern with a time delay. Such a delay is evident in many aspects of Australian life. Economic factors such as small markets and high transportation costs occasioned by the great distances have limited industrial development until recent years. However, with increased immigration of displaced persons now being permitted, and with natural population growth, the markets are becoming reasonable. Local production can reduce costs and raise quality, and the latter improvement is badly needed if we assume that the Australian housewife can be made as hungry for gadgets as her U. S. sister. We saw no evidence of a Madison Avenue influence, but complaints of several ladies who had lived in England or the U. S. convinced us of their susceptibility to special pleading. Certainly the present high costs and lowered quality were evidenced by a small washing machine, not automatic, at \$450 U. S., a well-known 17" table model TV set at \$450, or one of the "lowest-priced three" automobiles at \$5000.

Close relations already exist between our two societies, and we can hope that the investment of IRE Australia, in a lobster luncheon for the IRE USA, Editor, will pay off in even closer ties.

Space. In beginning this travelogue last month, the Editor reported the not-profound conclusion that travel is broadening. He now wishes to report further not-profound observations resulting from 26,000 miles of travel in the Far East and Australia, to wit: Five-sixths of the world's surface is indeed covered by water. The remainder of the earth's surface is always covered by clouds. Maps utilizing Mercator's projection are inaccurate. There are 300,000 bicycles on the streets of Saigon at 5:30 each day. The girls in Saigon look nice on bicycles. All air travel in that part of the world will be rough. Cities in the Far East should be moved closer together.—J.D.R.

Kenneth V. Newton

Director, 1957–1958

Kenneth V. Newton was born in Odessa, Mo. on May 26, 1922. He received his Bachelor's degree in education, physics, and mathematics from Central Missouri State College. In 1942 he entered the Graduate School at the University of Missouri as a Graduate Assistant in the Physics Department. After serving as Instructor, he was commissioned into the United States Naval Reserve in 1944 and served as Radar Officer. As Design Engineer with the Systems Research Laboratory of Johns Hopkins University, Capehart-Farnsworth Corporation, and Emerson Electric Manufacturing Company from 1946 to 1950, he was chiefly concerned with military electronic systems, particularly bombing computers. During this time, Mr. Newton taught part-time at Indianà Technical College and did graduate work in electrical engineering at Washington University. He is now a registered Professional Engineer in the State of Missouri. In August, 1950, he joined the Kansas City Division of the Bendix Aviation Corporation and served as Liaison Engineer and Senior Project Engineer. He has been Supervisor of the Components Engineering Department since December, 1955.

Mr. Newton joined the IRE as an Associate in 1946, and in 1953 he became a Senior Member. He has actively participated in the Kansas City Technical Conference, now called the Mid-America Electronics Conference, for the past several years, and in 1954 he served as Chairman of the Kansas City IRE Section.

He also served on the Papers Committee of the Electronic Components Symposium last year, as a member of the Nominations and Appointments Committee and IRE Region Six Director this year.

Scanning the Issue

JTAC—Ten Years of Service (Fink, p. 823)—This month marks the tenth anniversary of the founding of one of the most unique advisory bodies ever to serve a profession or industry. The Joint Technical Advisory Committee, composed of eight distinguished engineers, was established by the IRE and the Electronic Industries Association to provide the FCC with objective, expert technical assistance on matters pertaining to the allocation and utilization of the radio spectrum. In ten years, the JTAC has amassed an extraordinary record of public service and an equally extraordinary reputation throughout the world for impartial and sound technical judgment on some of the knottiest problems ever to confront an industry. The inspiring record of this group of "consulting engineers to the world" is now recounted by the President of the IRE and former Chairman of JTAC.

A Statistical Description of Coincidences Among Random Pulse Trains (Stein and Johansen, p. 827)—The author considers the following problem: Suppose there are several trains of pulses, all proceeding at the same time. The pulses occur at random. All we know is that the pulses in each train occur at some average rate and that the pulse lengths are distributed according to some probability law. What are the chances that pulses will occur in all trains at the same instant, and for how long an instant? The answer to this question automatically provides an answer to a similar question of much more immediate practical importance: In a communication system using diversity reception techniques, what are the chances that serious fading will occur on all receiving channels at the same instant, and for how long an instant? The reviewers felt that the concise and clever answer provided by this paper would interest specialists in many different fields. Between them they named seven: communication systems, information theory, antennas and propagation, broadcast transmission systems, electronic computers, instrumentation, and radar.

Transient Response of Drift Transistors (Johnston, p. 830)—The drift or graded-base transistor was proposed about two years ago as a way of obtaining a high cutoff frequency without the difficulty of making a very thin base. The resistivity of the base increases from emitter to collector in such a way as to hurry the minority carriers toward the collector by means of a "built-in" field. This paper analyzes the improvement in transient response caused by this built-in field, thus filling out our understanding of a type of transistor which is now coming into important use.

Terminal Properties of Magnetic Cores (Chen and Papoulis, p. 839)—Magnetic cores are being widely used as memory and switching elements in computers. Their operation in applications involving events much longer than one microsecond is well understood and readily determined from the B-H hysteresis loop for the material. For shorter events, cores less than one thousandth of an inch thick are required and the B-H loop no longer suffices to predict the circuit performance of the cores. A new approach has been developed which now permits the designer to determine the circuit characteristics of thin cores.

Some General Properties of Nonlinear Elements—Part II. Small Signal Theory (Rowe, p. 850)—Two years ago the PROCEEDINGS published a study of nonlinear reactance devices which related the signal powers at the input, output and local oscillator frequencies, giving very useful information about the gain and stability of such devices as magnetic and dielectric amplifiers. Since then, the great interest that has arisen in parametric amplifiers in general, and the new ferromagnetic amplifiers in particular, has made this study one of the most basic and widely referenced papers to appear in the PROCEEDINGS in the recent past. The results have even found application in quantum mechanical systems such as masers. The sequel to this study is now presented. It goes beyond general energy considerations and delves into specific device characteristics such as gain, bandwidth, impedance and sensitivity, showing the way these quantities depend on the amount of nonlinearity. This, like its predecessor, will undoubtedly become somewhat of a classic.

Very Low-Noise Traveling-Wave Amplifier (Kinaman and Magid, p. 861)—Two months ago a backward-wave amplifier was described that had a record low noise figure of 3.7 db. Two letters in this issue report an even lower 3.5 db. This paper reports on another low noise tube capable of at least 6.3 db, and in a number of instances showing 4.8 to 5 db. The 3.7 db tube, it must be remembered, was a research model whereas the tube described here is now commercially available. It represents the culmination of six long years of effort to reduce the noise figure of traveling-wave amplifiers below the old 9 db level.

New Applications of Impedance Networks as Analog Computers for Electronic Space Charge and for Semiconductor Diffusion Problems (Čremošnik, *et al*, p. 868)—Many mathematical problems involve partial differential equations for which no exact solution is known. In this paper the authors set up networks of resistances and resistance-capacitance chains whose voltages, currents and impedances correspond to significant parameters in equations of this sort. They are thus able to produce approximate solutions to important equations which arise in the design of triodes and the study of diode junctions, offering both a short-cut in designing tubes and new information on the behavior of semiconductors.

IRE Standards on Solid-State Devices: Methods of Testing Point Contact Transistors for Large-Signal Applications (p. 878)—This Standard stipulates the methods of measuring the important characteristics of point-contact transistors in power amplifiers, pulse amplifiers, oscillators, multivibrator-type switches and other large-signal applications.

Carrier-to Noise Statistics for Various Carrier and Interference Characteristics (Clarke and Cohn, p. 889)—Procedures are presented for determining what per cent of the time the carrier-to-noise ratio will fall below any given minimum value under a variety of signal and noise conditions. The results are presented graphically in a form that can be converted to any practical situation and are applicable to the design of radio communication systems of many types.

Simultaneous Asynchronous Oscillations in Class-C Oscillators (Disman and Edson, p. 895)—This paper covers the general problem of oscillator starting and the specific problem of simultaneous oscillations at two unrelated frequencies. Although few systems require such a phenomena, asynchronous oscillations do sometimes occur as an undesired effect, such as spurious responses in a crystal, and this study sheds light on how to eliminate them. Perhaps more important, the concept of negative discrimination by which the authors explain the existence of asynchronous oscillations provides a new tool which might well find useful applications in predicting the behavior of nonlinear devices.

Theoretical Diversity Improvement in Frequency-Shift Keying (Pierce, p. 903)—The advantages of using more than one receiver to help overcome the effects of randomly fading signals have been recognized for over a decade. The growing importance of binary data transmission has led more recently to a study of diversity reception as applied to frequency-shift keying systems in particular. The analysis presented here determines the best methods for combining the several signals received by the several receivers under various circumstances so as to insure the most reliable transmission of the message.

***Scanning the Transactions* appears on page 924.**

Section Survey of IRE Editorial Policies

LAST year the Editorial Board asked the help of IRE Sections in obtaining a grass-roots expression of opinion from the membership concerning the present policies and procedures governing the PROCEEDINGS and other IRE publications. Toward that end, the Editorial Board submitted to each Section Chairman a series of 12 questions on which information was desired, as set forth on page 884 of the June, 1957 PROCEEDINGS.

Forty-three Sections, one Subsection and one Student Branch responded by conducting surveys among their members and summarizing the results for the Editorial Board. It was recognized that the fashion in which the polls were taken would not lead to statistically accurate results. It was hoped, however, that the results would provide a crude measure of any major trends of opinion or preference that might be helpful to the Editorial Board.

The major findings of the survey can be summarized as follows:

1) PROCEEDINGS *Papers*

Most members agree that the PROCEEDINGS should continue to carry technical papers on important work, even though advanced and highly technical in nature. They do not want to lower its standards or value as a high quality technical journal. Some papers, it is felt, are too specialized and should be published in the TRANSACTIONS. Others are considered unnecessarily obscure to the nonspecialist and should be clarified, but not simplified. The primary and most widely expressed request was that the number of review and tutorial papers be increased. As a result, the Editorial Board is stepping up its program of review papers.

2) PROCEEDINGS *Departments*

The PROCEEDINGS departments are all put to good use except "Technical Committee Notes." "Scanning the Issue" is the most widely read and "Abstracts and References" the most valued for reference. There were a number of suggestions that TRANSACTIONS articles be listed or reviewed in the PROCEEDINGS. Apparently, quite a few members are unaware of the "Abstracts of IRE TRANSACTIONS" section of PROCEEDINGS. The new "Scanning the TRANSACTIONS" department, which began in February, 1958, takes care of those who requested reviews.

3) *Mechanical Aspects of* PROCEEDINGS

Some members have difficulty finding the contents page and binding the editorial matter. All other mechanical aspects, including mailing, are good. The Table of Contents has since been moved to the front page of the issue.

4) *Other Publications*

The TRANSACTIONS, DIRECTORY, CONVENTION RECORD and STUDENT QUARTERLY are, in general, well received and approved in their present form.

5) *Miscellaneous Comments*

(a) The special issues of the PROCEEDINGS are very popular with all members.

(b) Most members save and make reference to back issues of the PROCEEDINGS.

(c) A majority find the advertisements of definite value and spend at least half of their reading time on advertisements.

(d) The IRE Standards which appear in the PROCEEDINGS are received with considerable interest.

(e) It was suggested that subject and author entries in the annual indexes list the issue and page number rather than a cumulative index code number. This suggestion is being adopted for future indexes.

The Editorial Board feels that the survey was a worthwhile endeavor in that it produced a number of helpful guidelines and useful suggestions. As noted on page 819 of this issue, several changes have already been put into effect. Others are under consideration and will be announced as they materialize.

We are most grateful to the Sections and to the many members who participated in the survey for the considerable time and thought they gave. The results have proved well worth the effort.—*The Editor*

JTAC—Ten Years of Service*

DONALD G. FINK†, PRESIDENT, IRE

This month marks the tenth anniversary of the founding of the Joint Technical Advisory Committee (JTAC) under the sponsorship of the IRE and the Electronic Industries Association. This distinguished group of eight engineers, in the course of advising the FCC on many knotty problems in the administration of the radio spectrum, has achieved an outstanding reputation for disinterested and expert judgment. To mark the occasion, we have asked the IRE President, former member and past chairman of the Committee, to review the activities of JTAC during its first decade.

—*The Editor*

ON March 23, 1948 the late Wayne Coy, Chairman of the Federal Communications Commission, addressed the President's Luncheon at the IRE National Convention. The Guest of Honor at the luncheon was IRE President B. E. Shackelford and the Toastmaster was W. R. G. Baker, Director of the Engineering Department of the Electronic Industries Association. Ben Shackelford and Doc Baker listened well to what Mr. Coy had to say: the FCC was in need of assistance in arriving at an adequate national allocation of television facilities, there were many other conflicting demands for spectrum space (notably those of the land mobile service), and the FCC was therefore in need of advice of a particularly disinterested kind.

At the time, the established instrument for conveying professional and industrial opinion on allocations to the FCC was the Radio Technical Planning Board. This was a massive aggregation of 13 panels which had been formed during the war to advise the Commission on post-war allocation problems. The RTPB was organized according to the services demanding spectrum space; one panel for vhf (FM) broadcasting, one for tv, one for portable mobile communications, and so on. The RTPB panel chairmen reported individually at the FCC hearings and the Planning Board itself did not attempt to moderate the often conflicting opinions and recommendations of the panels. It was Chairman Coy's point that the administration of the spectrum should not be guided by groups devoted to particular services. On July 1, 1948, the RTPB was disbanded and its functions were taken over by the existing committees of IRE and EIA (then known as RMA).

Less than two weeks before the RTPB expired, on June 17 and June 20, respectively, Drs. Shackelford and Baker put their signatures on the charter of a new organization, the Joint Technical Advisory Committee. JTAC was specifically conceived to meet the need expressed by Chairman Coy. It was to be a committee of eight members. Its jobs in the words of its charter were to "obtain and evaluate information of a technical or engineering nature relating to the radio art, for the purpose of advising government bodies and other industrial and professional groups . . . to determine what technical information is required to insure the wise use and regulation of radio facilities . . . to sift and evaluate information thus obtained so as to resolve conflicts of fact, to separate matters of fact from matters of opinion, and to relate the detailed findings to the broad problems presented to it . . . to present its findings in a clear and understandable manner . . . available to the profession and the public . . . (and) to appear as necessary before government bodies or other parties to interpret the findings of the Committee. . . . ."

The genesis of JTAC goes back to Dr. Shackelford. When he assumed the IRE Presidency in 1948, he proposed the formation of an IRE Technical Committee on Spectrum Utilization to gather evidence on the characteristics of different portions of the spectrum and to correlate them with the needs of particular classes of service. Encouraged by Chairman Coy's speech, he met with Dr. Baker and together they roughed out a plan for a joint IRE-EIA Committee to consider the techniques of spectrum utilization and to assist the FCC as requested. The idea was broached to the respective Boards of Directors and received their blessing. Two men were selected to serve as the first JTAC Chairman (P. F. Siling) and Vice-Chairman (D. G. Fink). These men met in IRE Headquarters on May 12, with a number of other interested engineers, to draw up the charter which was later ratified by the sponsoring bodies. To fill out the roster, Ralph Bown, Melville Eastham, John V. L. Hogan, E. K. Jett, Haraden Pratt, and David B. Smith were appointed to membership. The members and their terms of office during the past ten years are listed in Table I. L. G. Cumming, Technical Secretary of the IRE, has served as JTAC's Secretary since 1948.

The past and present members of JTAC include five IRE presidents and eleven members of the IRE Board

* Original manuscript received by the IRE, March 11, 1958.
† Director of Research, Philco Corp., Philadelphia, Pa.

TABLE I

PRESENT AND FORMER MEMBERS OF JTAC

Lloyd V. Berkner	President, Associated Universities, Inc.	1952–1955
Ralph Brown	Retired; formerly Vice-President and Director of Research, Bell Telephone Laboratories	1948 to date
A. B. Chamberlain	Chief Engineer, Columbia Broadcasting System—Television	1954–1955
Melville Eastham	Retired; formerly President, General Radio Company	1948–1950
Donald G. Fink	Director of Research, Philco Corporation; formerly Editor, *Electronics*	1948–1954; 1956–1957
T. T. Goldsmith, Jr.	Vice-President, Allen B. DuMont Laboratories, Inc.	1950–1952
Ralph N. Harmon	Director and Vice-President for Engineering, Westinghouse Broadcasting Corporation	1955–1956
John V. L. Hogan	President, Hogan Laboratories, Inc.; Founder, IRE	1948 to date
Dorman D. Israel	Executive Vice-President, Emerson Radio and Phonograph Corporation	1952 to date
Ewell K. Jett	Vice-President, Director of Television Baltimore Sunpapers; formerly Federal Communications Commissioner	1948–1950
I. J. Kaar	Vice-President and Engineering Director, Hoffman Electronics Corporation	1950 to date
Arthur V. Loughren	Vice-President, Airborne Instruments Laboratory, Inc.	1951–1955; 1957 to date
Haraden Pratt	Vice-President, Dualex Corporation; formerly Telecommunications Advisor to the President of the United States	1948–1951
William H. Radford	Professor of Electrical Communications and Associate Director, Lincoln Laboratory, M.I.T.	1956 to date
Philip F. Siling	Director, RCA Frequency Bureau	1948 to date
David B. Smith	Vice-President—Research, Philco Corporation	1948–1952
Ernst Weber	President, Polytechnic Institute of Brooklyn	1955 to date

of Directors. All are Fellows of the Institute. Concerning membership, the JTAC charter has this to say: "The members shall be chosen on the basis of professional standing, integrity, and competence to deal with the problems to be considered by the Committee. The members shall be chosen from among all qualified engineers irrespective of the organizations to which they belong or the companies by whom they are employed, and shall operate without instruction. Half of the members shall be nominated by IRE and half by EIA, and the appointment of all members shall be confirmed by both bodies. None of the members shall receive any regular compensation for services from the National or any State Government."

THE EARLY YEARS—TELEVISION ALLOCATIONS AND COLOR

Early in May, 1948, Commissioner Coy asked that IRE and EIA assist in collecting information on allocation of uhf channels to television. At the JTAC organization meeting of May 12, it was agreed that the first formal activity would be in answer to this request. The Vice-Chairman met with the FCC staff on June 18 and obtained a list of technical questions which were transmitted to IRE and EIA committees and to the technical press. By September 16, the JTAC had been fully organized and a 150-page report, "Utilization of Ultra-High Frequencies for Television," had been produced and approved by the Committee. This report, Volume 1 of the *Proceedings of the JTAC*, was presented that month before the FCC Hearing on UHF Allocations, Docket 8976.

In the first report, JTAC established three degrees of reliability of the information it presented. Class A data were defined as "Established fact—data on which there is a general agreement among informed specialists, based on adequate measurements or theory." Class B was assigned to "Engineering estimates—data based on limited experience, or statements based on theory not fully confirmed." Class C was flatly labeled "Speculation—conjectures based on more or less arbitrary extrapolation from limited experience."

The report laid out the principles of predicting coverage and interference on uhf channels and stated typical values of noise, noise figure, transmission line loss, effective heights of antennas, transmitter power, and radiator gain then available at the center and the extremes of the uhf band. Pointing out that measurements of field strength generally ran well below theoretical predictions, the report made the then astounding estimate that a uhf station would need from 5 to 20 megw of effective radiated power to lay down a signal strength of 5000 μv per meter at 40 miles from a 500-foot radiator. Ten years later such power levels were actually authorized for uhf stations (5 megw have been permitted by the FCC since 1956, and several stations are now operating at levels above a megawatt).

JTAC made the observation in this report that "the place in the spectrum in which it is technically possible for 6-megacycle black-and-white television immediately to expand its number of channels is in the immediate vicinity of the present commercial channels." This advice, if it had been taken in the light of the remarkably accurate concurrent prediction of the Senate Advisory Committee on Color Television respecting the competition to be expected between uhf and vhf stations, might well have altered the whole face of television allocations today. JTAC advised against shared operation and "interim" allocation of tv channels. It concluded that "from available data, coverage comparable with that of the vhf service, using available or potentially available transmitter power, is not possible on the ultra-high frequencies with ground-based transmitters."

The second volume of the JTAC *Proceedings*, "Allocation Standards for VHF Television and FM Broadcasting," came along only three months later, and was presented to the FCC Engineering Conference on that subject (Docket 9175) in December, 1948. Four EIA

Committees and two IRE Committees acted for JTAC in assembling data. The upshot of this study as a set of specific recommendations for tv and FM allocations, giving signal strengths, co-channel and adjacent-channel protection ratios for urban and rural service. With minor modifications, these numbers are now parts of the FCC Standards of Good Engineering Practice for these services. Noted for this first time in this report was the improvement in co-channel interference resulting from "phase-synchronization" of television carriers, later embodied in the offset-carrier principle of television frequency assignments.

Next came Volume 4.[1] It bore the innocuous title, "Comments on the Proposed Allocation of Television Broadcast Services." It dealt with the issues before the FCC in the Hearing that began September 26, 1949 (Dockets 8736, 8975, 8976, and 9175), one of which was color television. Part 1 of the report dealt with the then proposed allocations of vhf and uhf channels. JTAC strongly recommended that channels be grouped within given service areas so as to minimize the damage to the service caused by local oscillator radiation and image responses. Such a plan assumed standard intermediate frequencies in receivers, and JTAC pointed to the EIA proposal for 41.25 mc (sound IF) as a suitable choice. These recommendations are now thoroughly embedded in the official FCC uhf television allocation plan. JTAC gave credit to one of its consultants, John D. Reid, for suggesting the principle of alternate grouping of channels and to Raymond Simonds for conducting detailed studies of this and other proposals. Also reported were extensive tests of co-channel interference with offset carrier assignments, compared with results under the older nonsynchronous and phase-synchronized systems.

The color television issue, considered at the same hearing, was whether 6-mc color television service was feasible. JTAC undertook in Part 2 of Volume 4 to describe and analyze no fewer than 12 different color proposals, using 6 and 12-mc bandwidth, of the simultaneous, field-sequential, line-sequential, and dot-sequential types. The systems were compared on the basis of eight essential characteristics, ranging from resolution and color fidelity to flicker and color breakup. JTAC recommended no one of the color systems but suggested a six-step procedure to be followed by the FCC before choosing any color system. The FCC did not follow the latter advice and shortly thereafter authorized the field-sequential color system. A reluctant public and the Korean War then took over. In retrospect, the subsequent chapters in the history of color tv seem to follow closely the JTAC six-step program.

These were parlous times. The Table of Color Systems prepared at JTAC's request by an EIA Committee was approved for presentation to the FCC on August 18, 1949. On August 24, a letter describing the work of RCA on a 6-mc compatible color system was sent to the JTAC Chairman. On September 22, JTAC authorized the addition of the new information to the table, just in time to present it to the Commission on September 26. The writer's copy of Volume 4 contains the inked-in entry of the RCA data, since there was no time to reprint the report. A supplement containing the revised table was issued later.

One remaining study of television allocation data was then pending. In March, 1950, Volume 5 of the JTAC *Proceedings*, "Adjacent-Channel Interference in Monochrome Television," was issued. This authoritative statement of propagation data and the properties of receivers was prepared by a new JTAC creature, a Subcommittee of Consultants specially chosen to deal with this particular issue. A very high proportion of the data in this report were asserted by JTAC to have Class A reliability. Through EIA, the selectivity characteristics of more than 100 different models of tv receivers, representative of nearly 3 million receivers (the great majority of those then in the hands of the public) were studied and applied to interference predictions. Subjective tests of adjacent-channel interference, in which 109 observers participated, were also reported.

Spectrum Conservation and Single Sideband

By 1951, when the new television allocation was ready to go into effect and color was safely in the hands of NTSC, the JTAC members were at last free to start something on their own. At the meeting of January, 1951, the discussion reverted to the purposes first proposed by Dr. Shackelford early in 1948, that is, the influence of technical factors on spectrum allocations in the broad sense. A subcommittee of three JTAC members was formed and directed to prepare, with the assistance of qualified experts, a report outlining a program of spectrum conservation. A stable of 22 authorities was thereafter assembled, the report was outlined, and the work was parceled out.

A year later the final draft of the report was approved by JTAC. Things then moved fast. It was decided that the report merited wide distribution and that it should in fact be published in book form. IRE and EIA agreed to provide a fund of $7000 to underwrite the printing of 3000 copies, a publisher was found, and by September, 1952, the book "Radio Spectrum Conservation" was off the press and in the mails.

This volume (Volume 8 of the JTAC *Proceedings*), more than any other project, has brought international recognition to JTAC. Copies were sent to radio regulatory bodies throughout the world. The book was widely and favorably reviewed and it remains today the only definitive treatment of the problems of spectrum administration. Reading it, Martin Codel was inspired to label the JTAC members "consulting engineers to the world." Contained in the book are a history of allo-

[1] Volume 3 and eight other volumes, issued annually to date, contain the correspondence and minutes of the Committee.

cations, a 90-page summary of propagation characteristics prepared by six internationally recognized authorities under the supervision of Dr. J. H. Dellinger, an ideal allocations table, and a critique of the present allocation. The book concluded with the proposed program of dynamic conservation and contained a bibliography of 137 items.

Another major publication project of JTAC culminated in the December, 1956, special issue of the PROCEEDINGS OF THE IRE on single-sideband techniques. In October, 1955, the FCC issued a Public Notice stating that it proposed to consider single-sideband techniques for future use in the mobile radiotelephone service. At its January, 1956, meeting JTAC decided, on its own motion, to conduct a study in support of the FCC inquiry and to collect information from organizations known to be active in this field. Another subcommittee was set up, consultants were assembled, and invitations to contribute were sent out. The response was phenomenal. No fewer than 34 papers were submitted and 30 of these were selected to appear in the JTAC report. Meanwhile, IRE had been approached concerning its interest in publishing the papers and the SSB special issue was the result. Chairman McConnaughey of the FCC wrote the preface to the issue, stating that "the work of the IRE and the JTAC will be of material assistance to the Commission in making provision for single-sideband transmission."

OTHER INVESTIGATIONS

Space does not permit a detailed account of the many other thorny issues on to which the JTAC members and their consultants have torn their hands. But mention must be made of the important contribution made to the allocation of channels in the land-mobile service. An expert subcommittee of consultants under the devoted leadership of the late Fred T. Budelman investigated, at length and under great pressure, the possibility of splitting channels in this service. JTAC confirmed the opinion of this group that channel splitting was feasible and so informed the Commission. The Commission then amended the rules governing the 152–162-mc band, and required the use of narrower channels in new equipment built after October, 1958, and in all equipment by 1963.

Several JTAC subcommittees have dug into the troublesome problems of radio interference and spurious emissions. One of these worked closely with the manufacturers of rf stabilized arc welders to find means of mitigating interference. In January, 1957, JTAC reported to the Commission that "the technique of reduced duty cycle operation of welders was a technical solution for significantly reducing unwanted radiation."

Following the appearance of the special issue of the PROCEEDINGS OF THE IRE in October, 1955, on forward scatter propagation, JTAC addressed itself to the questions of allocating scatter communications service, in view of the interference potentials inherent in the high transmitter power and receiver sensitivity of this new mode of operation. Three subcommittees are still at work on this subject. JTAC has also helped in the setting up Cooperative Interference Committees, to assist the FCC in running down the causes and cures of radio interference on a local community basis.

It is too much to expect that every project undertaken by JTAC would be crowned with success. For example, it became increasingly apparent to JTAC and industry groups that there was insufficient coordination on problems of frequency utilization and conservation among the United States Government, nongovernment users, and manufacturers. In 1954 the then chairman of JTAC, Lloyd V. Berkner, acting for the Committee, attempted to get various government agencies interested in implementing a program of spectrum conservation. Nothing ever came of it. Then as now, the ponderous machinery of government administration proved invincible to prodding from without. Another well-intentioned activity was an effort conducted from 1952 to 1954 to recommend simplifications in the technical requirements embodied in the FCC Rules and Regulations and the Standards of Good Engineering Practice. In March, 1954, the Committee advised the Commission that it was unable to make any specific recommendations, beyond a few minor suggestions dealing with the land mobile and broadcasting services. JTAC wisely concluded that changes in this area were so deeply involved in administrative procedures and policy that they lay beyond the purview of a committee bounded by technical considerations.

The record of JTAC is a brilliant one of public service in a difficult context. On behalf of IRE, the writer can find no better words to underscore the ten-year record than those which appeared in the special issue of the IRE PROCEEDINGS sponsored by the JTAC on single-sideband techniques: "This Committee of eight senior statesmen has, since its creation in 1948, conducted its affairs with such singular wisdom that it has achieved, preeminently among engineering advisory bodies, the unquestioned confidence of the profession, the industry and the government. . . . This standing has been achieved not by avoiding controversial issues but by meeting them with pertinent technical data, carefully identified as to reliability, and by truly objective evaluation of their significance. The JTAC charter requires its members to serve wholly without instruction from their employers or any other organization, and the record shows wholehearted compliance with this necessary prerequisite to public and professional trust."

Congratulations, JTAC members, and best wishes for the next decade!

A Statistical Description of Coincidences Among Random Pulse Trains*

S. STEIN†, SENIOR MEMBER, IRE AND D. JOHANSEN†, MEMBER, IRE

Summary—A simple formulation is developed for describing statistically the time coincidences among a set of random pulse trains. Specifically, each pulse train is assumed to consist of pulses occurring randomly and having pulse lengths with some known distribution. It is then possible to determine the statistical distribution for the resulting durations of coincidences among several individually specified random pulse trains. From this can be derived such corollary results as the average duration of the resulting coincidences.

I. Introduction

In scatter communications, it is often important to know what effect diversity operation has upon the effective duration of fades of the diversity combined signal. Conceptually, this problem is essentially identical to that of determining the statistics of the time coincidences among a set of random pulse trains. A very simple formulation for this problem has been developed, which does not appear to have been previously described in the literature. The important result is the derivation of a statistical distribution for the durations of the coincidences, when a similar distribution is known for the pulse lengths in the individual pulse trains. This result is obtained by setting up the problem in such a manner that deriving the distribution of the durations of the coincidences is reduced to the almost trivial problem of determining the probability of a coincidence at any instant.

It is expected that the results may have application in other fields, and hence the problem is presented in the form of the random pulse train model.

II. Statistical Description of a Random Pulse Train

We consider first the statistical description of a random pulse train, such as shown on any line of Fig. 1. The pulse train consists of a series of pulses occurring in a random manner, specified by some average rate of occurrence and with pulse lengths distributed according to some probability law. [The average rate of occurrence and the distribution of pulse lengths may be given separately. If so, they can be combined (see Appendix I) to give the statistical quantity, $q_i(x)$, used below.] The function $q_i(x)$, which we regard as the basic statistical description of the ith random pulse train, is defined by the property:

$q_i(x)dx$ = expected number of pulses per unit time, whose length is in the range $(x, x+dx)$.

* Original manuscript received by the IRE, November 26, 1957.
† Hycon Eastern, Inc., Cambridge, Mass.

We also define a cumulative function,

$Q_i(x)=\int_x^\infty q_i(x)dx$ = expected number of pulses per unit time, whose length exceeds x.

We shall consider that our final statistics are representative of sufficiently long periods of time so that these "expected" values are in fact the probability distributions which would be measured for the quantities described above.

We next note that the total fraction of time during which pulses are occurring is

$$R_i = \int_0^\infty xq_i(x)dx \text{ (time per unit time).}$$

III. Coincidences Among Several Random Pulse Trains

We now consider several independent random pulse trains, such as the four designated as "inputs" in Fig. 1. The pulses are shown by the heavy outlines. Let us investigate the statistics for the intervals of time during which pulses are occurring *simultaneously* in *all* the "input" trains. (We may visualize a circuit which produces output pulses whenever, but only when, pulses are occurring simultaneously in all input lines.) The resulting coincidence pulses are designated in Fig. 1 as the output. The manner in which they occur is clearly indicated. Now let us designate by

$$q_{on},\ Q_{on}, \text{ and } R_{on}$$

quantities which describe the output in a manner analogous to the quantities q_i, Q_i and R_i already defined for the inputs. Here n is the total number of random pulse train inputs. Our problem clearly is:

Given

$$q_i(x) \text{ or } Q_i(x); \qquad \text{for } i = 1, \cdots, n,$$

to determine

$$q_{on}(x) \text{ or } Q_{on}(x).$$

Let us first consider the simplest aspect, namely that of determining R_{on}. By assumption, the occurrence of pulses in any input train is a completely random process, independent of the occurrence of pulses in any other input train. Hence R_{on} is obtained from the various R_i by the law governing the probability of simultaneous occurrence of independent events. In other words, since R_i is the fraction of time for which pulses occur in the ith train, the probability, or fraction of time, for which

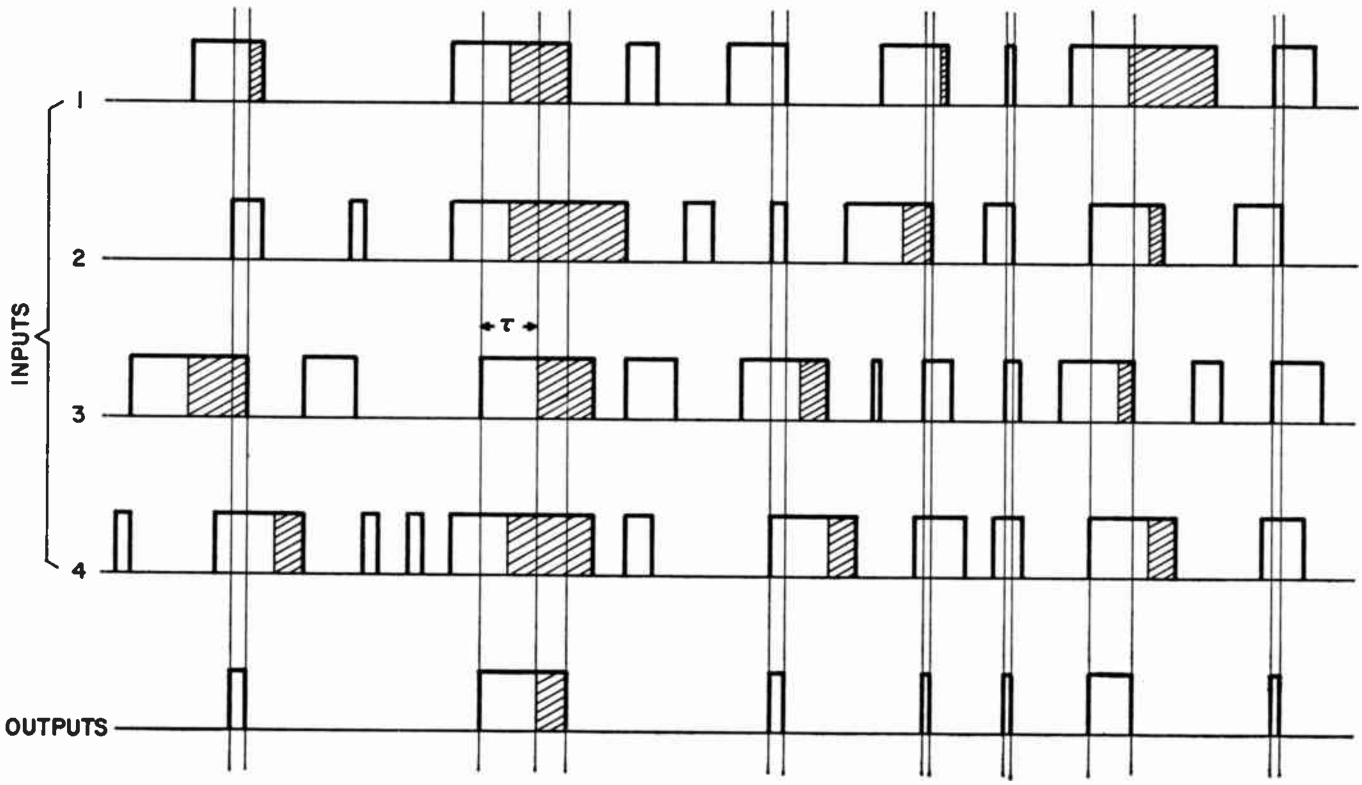


Fig. 1—Coincidences among random pulse trains.

pulses occur simultaneously in all trains, is simply

$$R_{on} = \prod_{i=1}^{n} R_i. \tag{1}$$

Returning now to our stated problem, we need to determine the statistics with respect to the length of coincidence in the output. Clearly, we require a criterion enabling us to sort the conditions producing coincidence pulses, according to the length of the latter. This is most simply accomplished as follows:

Let us choose a specific length of coincidence pulse, τ, and imagine deleted from Fig. 1 all pulses of length less than τ (input pulses shorter than τ can certainly never contribute to coincidences longer than τ). Let us also substract a length τ off at the leading edge of all remaining pulses; we thus obtain a reduced picture indicated by the hatched pulses in Fig. 1. Carefully considering this reduced picture, we note that *any* simultaneous coincidence of hatched elements for *all* input trains always corresponds to a hatched element in the output; that is to say, since *every* input train actually has a pulse in progress for an interval τ preceding the hatched element, the output hatched element will also correspond to a coincidence pulse in progress for a preceding interval of time τ. Furthermore, it is readily visualized that this characterization must include every event leading to an output coincidence pulse of length τ, or greater.

But we have already indicated the occurrence-probability relation which will hold between the hatched elements in the input and output pulse trains. Since the hatched elements occur independently among the inputs, we can directly apply the reasoning which led to our earlier statements concerning R_{on} and R_i.

Specifically, the fraction of time occupied by hatched elements (pulses, with lengths diminished by τ) in, say, the ith input train, is

$$\int_{\tau}^{\infty} (x - \tau) q_i(x) dx.$$

Similarly, for the output, the comparable quantity is given in terms of the (unknown) function $q_{on}(x)$ as

$$\int_{\tau}^{\infty} (x - \tau) q_{on}(x) dx.$$

The probability law for independent events now gives immediately

$$\int_{\tau}^{\infty} (x - \tau) q_{on}(x) dx = \prod_{i=1}^{n} \int_{\tau}^{\infty} (x - \tau) q_i(x) dx. \tag{2}$$

Integration by parts gives directly

$$\int_{\tau}^{\infty} (x - \tau) q(x) dx = \int_{\tau}^{\infty} Q(x) dx. \tag{3}$$

so that our result has the alternative form

$$\int_{\tau}^{\infty} Q_{on}(x)dx = \prod_{i=1}^{n} \int_{\tau}^{\infty} Q_i(x)dx. \tag{4}$$

This is valid for any value of τ, and if it is now differentiated with respect to τ, we have the desired result

$$Q_{on}(\tau) = -\frac{d}{d\tau} \prod_{i=1}^{n} \int_{\tau}^{\infty} Q_i(x)dx. \tag{5}$$

This is also readily seen to be identical to the form

$$Q_{on}(\tau) = \left\{ \prod_{i=1}^{n} \int_{\tau}^{\infty} Q_i(x)dx \right\} \sum_{i=1}^{n} \frac{Q_i(\tau)}{\int_{\tau}^{\infty} Q_i(x)dx}. \tag{6}$$

This is the solution to our stated problem. It is not difficult to use for computational purposes, even if the Q_i can only be integrated numerically. For the relatively common case where the input random pulse trains have identical statistics, $Q_i(x) \equiv Q(x)$,

$$Q_{on}(\tau) = nQ(\tau) \left[\int_{\tau}^{\infty} Q(x)dx \right]^{n-1}. \tag{7}$$

A useful corollary of our computations is the result (see Appendix II) that the *average* duration of coincidence pulses, $\bar{x}_{on}$, is given by

$$\frac{1}{\bar{x}_{on}} = \sum_{i=1}^{n} \frac{1}{\bar{x}_i}, \tag{8}$$

where $\bar{x}_i$ is the average pulse length in the ith train. For n statistically identical input trains, this has the simple form

$$\bar{x}_{on} = \frac{\bar{x}_i}{n}. \tag{9}$$

For the interesting communications application to fading carriers, where diversity is based on selection of the strongest signal among n available, fades below a specific level occur in the output only when all n channels fade below that level simultaneously. It is interesting to note that the average resulting fade durations are then reduced in length by a factor of n.

Finally, it may be remarked that interest may lie not in the time rate of occurrence $[q_i(t)]$ of coincidences of various length, but rather in the distribution of durations among such coincidences as do occur. If so, the various relations introduced in the appendices can readily be used to recast the various results so that they involve only such distributions.

Appendix I

Alternative Description of the Random Pulse Trains

Often, measurement data on random pulse trains may be obtained not in a form involving $q_i(x)$ or $Q_i(x)$ directly, as defined in the text, but rather through the quantities

R_i = probability, or fraction of total time, that a pulse is in progress in the ith random pulse train.

$W_i(x)$ = probability that any particular pulse in the ith train has a length exceeding x.

Also, related to $W_i(x)$ is the probability density $w_i(x)$, such that

$w_i(x)dx$ = probability that any particular pulse in the ith train has a length in the range $(x, x+dx)$

and, of course,

$$w_i(x) = -\frac{d}{dx} W_i(x). \tag{10}$$

The average pulse length in the ith train is then (integrating by parts)

$$\bar{x}_i = \int_0^{\infty} x w_i(x)dx = \int_0^{\infty} W_i(x)dx. \tag{11}$$

Then, clearly, the average rate at which pulses occur in the ith train is

$$\frac{R_i}{\bar{x}_i} = \text{(pulses/second)}.$$

It immediately follows that the quantities used in the text are given by

$$q_i(x) = \frac{R_i}{\bar{x}_i} w_i(x), \tag{12}$$

$$Q_i(x) = \frac{R_i}{\bar{x}_i} W_i(x). \tag{13}$$

Appendix II

Average Length of the Coincidences

If we define $w_{on}(x)$ and $W_{on}(x)$ for the output coincidence pulses, analogous to the probabilities defined above, we may write

$$Q_{on}(x) = \frac{R_{on}}{\bar{x}_{on}} W_{on}(x). \tag{14}$$

This may be solved conveniently for $\bar{x}_{on}$, by setting $x=0$. By definition $W_{on}(0)=1$, and thus

$$\bar{x}_{on} = \frac{R_{on}}{Q_{on}(0)} \cdot \tag{15}$$

But, from our basic result, (6),

$$Q_{on}(0) = \left\{ \prod_{i=1}^{n} \int_0^\infty Q_i(x)dx \right\} \sum_{i=1}^{n} \frac{Q_i(0)}{\int_0^\infty Q_i(x)dx} \cdot$$

Also,

$$\int_0^\infty Q_i(x)dx = \frac{R_i}{\bar{x}_i} \int_0^\infty W_i(x)dx = R_i. \tag{16}$$

Hence,

$$Q_{on}(0) = \left\{ \prod_{i=1}^{n} R_i \right\} \sum_{i=1}^{n} \frac{Q_i(0)}{R_i}$$

$$= R_{on} \sum_{i=1}^{n} \frac{Q_i(0)}{R_i} \cdot \tag{17}$$

Finally

$$Q_i(0) = \frac{R_i}{\bar{x}_i} W_i(0) = \frac{R_i}{\bar{x}_i}, \tag{18}$$

and so

$$\frac{1}{\bar{x}_{on}} = \frac{Q_{on}(0)}{R_{on}} = \sum_{i=1}^{n} \frac{1}{\bar{x}_i} \cdot \tag{19}$$

Transient Response of Drift Transistors*

R. C. JOHNSTON†, MEMBER, IRE

Summary—**The short-circuit transient response of a drift transistor is found for a step of input current. The one-dimensional partial differential equation for minority carriers in the base is solved by a Laplace transform technique yielding the rise, storage, and fall times in the common-emitter and common-base configurations. The improvement in rise time due to the built-in field in a drift transistor is found to be less in the common-emitter than in the common-base configuration, which in turn is found to be less than the improvement previously predicted. The built-in field is shown to lengthen storage time, but other effects of the field such as low α_I tend to cancel this out in a practical transistor.**

Introduction

The drift or graded-base transistor has been proposed as a way of obtaining a high cutoff frequency without the difficulty of making a very thin base. The resistivity of the base increases from emitter to collector in such a way as to cause the minority carriers to be swept toward the collector by a "built-in" field. The effect of this field on the common-base cutoff frequency has been analyzed,[1,2] but the effect of the field on the transient response in the common-emitter and common-base configurations has not been treated. The Laplace transform method used by Macdonald[3] to find the common-base transient response for a homogeneous-base transistor is extended to the drift transistor and to common-emitter and saturation conditions.

Assumptions

The following assumptions are made:

1) The transistor is considered one dimensional.
2) The minority carrier density is much less than that of the majority carriers.
3) The mobility is constant. Actually, it decreases for high conductivities and injection levels. This

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[1] J. L. Moll and I. M. Ross, "The dependence of transistor parameters on the distribution of base layer resistivity," Proc. IRE, vol. 44, pp. 72–78; January, 1956.

[2] H. Krömer, "Zur theorie des diffusions- und des drifttransistors," *Arch. Elec. Übertragung*, vol. 8, pp. 223–228, 363–369, 499–502; 1954.

[3] J. R. Macdonald, "Solution of a transistor transient response problem," IRE Trans. on Circuit Theory, vol. CT-3, pp. 54–57; March, 1956.

effect has recently been taken into account in a computer solution.[4]

4) The field due to the grading of impurities is constant throughout the base region. This implies that the collector voltage is large enough to insure that the part of the base region having a nonconstant field is within the collector space charge region.
5) The base lifetime is infinite.
6) The collector load is small enough to allow the effect of collector capacity and space charge widening to be neglected.
7) An n-type base is assumed for convenience in notation.
8) The emitter transition capacity, C_{TE}, can be neglected. For some graded-base types this requires that a bias emitter current be used. Fig. 1 illustrates the effect of C_{TE}. In a graded-base transistor C_{TE} tends to be larger because the resistivity on the edge of the base region is lower than in a homogeneous-base transistor, while C_{diff} tends to be smaller because of the higher cutoff frequency. Therefore, a bias current may be required so that the condition $C_{diff} \gg C_{TE}$ is fulfilled.

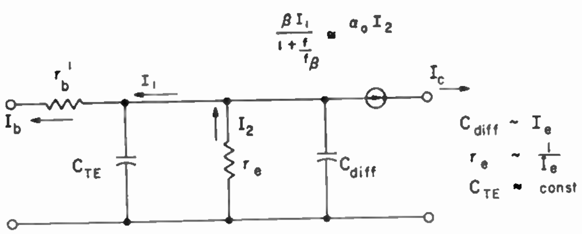

Fig. 1—Effect of emitter transition capacity.

The field in the base due to grading of impurity densities can be obtained from the following equations.

$$n = n_i e^{(q/kT)(\psi-\phi)} \tag{1}$$

$$p = n_i e^{-(q/kT)(\psi-\phi)} \tag{2}$$

$$n - p = N_D - N_A \triangleq N \tag{3}$$

$$E = -\frac{d\psi}{dx}. \tag{4}$$

A solution for the field yields

$$E = -\frac{kT}{q}\frac{1}{n+p}\frac{dN}{dx}. \tag{5}$$

If n and p differ greatly in magnitude

$$E = -\frac{kT}{q}\frac{d\ln N}{dx} \tag{6}$$

with a solution for the constant field case

$$N = N_0 e^{-(qE_0/kT)x}. \tag{7}$$

Substitution of (7) into (5) gives

$$E = E_0\frac{N}{n+p} = E_0\frac{N}{\sqrt{N^2+4n_i^2}} \tag{8}$$

which shows that the field due to an exponential gradation is constant for $N \gg n_i$ but decreases as N approaches n_i. In practice this impurity gradation is obtained by gaseous diffusion of n-type impurities into near intrinsic material of either n or p type. The substitution of the resulting distribution, an exponential plus or minus a constant, into (5) gives the field. The exponential minus a constant distribution gives a field which may actually rise as x approaches the collector junction.[5]

The error function distribution obtained for some types of diffusion is similar to the exponential distribution, but substitution into (5) gives a field less constant than (8). The nonuniformity of mobility should also be taken into account in any exact solution. In view of the difficulties of taking the exact field into account, a constant field has been assumed.

Basic Equations

The two fundamental equations[6] governing minority carrier flow in a semiconductor are the continuity equation

$$-\frac{1}{q}\frac{\partial I_p(x,t)}{\partial x} - \frac{p(x,t)-p_0(x)}{\tau_p} = \frac{\partial p(x,t)}{\partial t} \tag{9}$$

and the transport equation

$$\frac{I_{p(x,t)}}{q} = -D_p\frac{\partial p(x,t)}{\partial x} + \mu_p E(x)p(x,t). \tag{10}$$

The derivative of (10) is substituted in (9) to give the differential equation governing hole flow in the base of a p-n-p transistor:

$$D_p\frac{\partial^2 p(x,t)}{\partial x^2} - \mu_p\left[E(x)\frac{\partial p(x,t)}{\partial x} + p(x,t)\frac{dE(x)}{dx}\right] - \frac{p(x,t)-p_0(x)}{p} = \frac{\partial p(x,t)}{\partial t}. \tag{11}$$

Assumptions 4) and 5) lead to the simpler form

$$\frac{\partial^2 p(x,t)}{\partial x^2} - \left(\frac{\mu_p}{D_p}E\right)\frac{\partial p(x,t)}{\partial x} = \frac{1}{D_p}\frac{\partial p(x,t)}{\partial t}. \tag{12}$$

With zero initial conditions the Laplace transform with respect to time gives

$$\frac{\partial^2 P(x,s)}{\partial x^2} - \left(\frac{2}{f}\right)\frac{\partial P(x,s)}{\partial x} = \frac{s}{D_p}P(x,s) \tag{13}$$

where $2/f$ is the notation of Krömer[2] for $(\mu_p/D_p)E$.

[4] H. B. von Horn and W. Y. Stevens, "Determination of transient response of drift transistors using the diffusion equation," *IBM J.*, vol. 1, pp. 189–191; April, 1957.

[5] R. C. Johnston, "Transient response of drift transistors," M.I.T., Electrical Engineering thesis, App. 1; January 22, 1958.

[6] J. L. Moll, "Junction transistor electronics," Proc. IRE, vol. 13, pp. 1807–1819; December, 1955.

Note that f has the dimensions of length. Eq. (13) is easily solved by means of the substitution $P=e^{mx}$.

$$m = \frac{1}{f} \pm \sqrt{\frac{1}{f^2} + \frac{s}{D_p}} \stackrel{\Delta}{=} \frac{1}{f} \pm \Gamma \tag{14}$$

$$P(x, s) = e^{x/f}(A \cosh \Gamma x + B \sinh \Gamma x). \tag{15}$$

The constants A and B of (15) may be found for the common-base, common-emitter, and saturation boundary conditions, and the inverse transforms obtained. Before this is done it is appropriate to consider the expression w, the base width, over f, which will appear as a parameter in the solutions.

$$\frac{w}{f} = \frac{\mu_p}{2D_p} Ew = \frac{q}{2kT} Ew = \frac{\Delta V}{2kT}. \tag{16}$$

ΔV is the potential energy drop across the base. A second useful expression for w/f is obtained from (6).

$$\frac{w}{f} = \frac{q}{2kT}\int_0^w Edx = -\frac{1}{2}\int_0^w \frac{d \ln N}{dx} dx$$

$$= \frac{1}{2} \ln \frac{N(0)}{N(w)}. \tag{17}$$

Common-Base Active Solution

Transient Case

Due to the assumption of infinite base lifetime, α, the current gain, is one. However, calculations for the homogeneous-base case[7,8] show that for reasonable value of α, the major effect of recombination is just to change the amplitude of the output. Thus in the equations of this section, α can be introduced as a multiplier of the emitter current with little error. A small step of emitter current, ΔIe, is considered to be superimposed on the initial current. The transient solution for the collector current is obtained by applying the boundary conditions to (15) and taking the inverse transform.

$$\left\{\begin{aligned} P(s, w) &= 0 \\ I(s, 0) &= \frac{\Delta Ie}{s} \end{aligned}\right. \tag{18}$$

$$P(s, x) = \frac{\Delta I_e e^{x/f}}{qD_p\Gamma s} \frac{\sinh \Gamma(w - x)}{\cosh \Gamma w + \frac{w/f}{\Gamma w}\sinh \Gamma w}. \tag{19}$$

The collector current is obtained by applying the transport equation (10) to (19) and setting $x=w$.

$$I(s, w) = \frac{\Delta I_e e^{w/f}}{s} \frac{1}{\cosh \Gamma w + \frac{w/f}{\Gamma w}\sinh \Gamma w}. \tag{20}$$

[7] J. S. Schaffner and J. J. Suran, "Transient response of the grounded base transistor amplifier with small load impedance," *J. Appl. Phys.*, vol. 29, pp. 1355–1357; November, 1953.
[8] Macdonald, *op. cit.*

It is now convenient to normalize the time scale. A basic theorem[9] states that if s is multiplied by a constant, say w^2/D_p and the function multiplied by the same constant, then the time scale is multiplied by the reciprocal or D_p/w^2.

$$\bar{I}(\lambda, w) = \frac{\Delta I_e e^{w/f}}{\lambda} \cdot \frac{1}{\cosh \sqrt{\lambda + \frac{w^2}{f^2}} + \frac{w/f}{\sqrt{\lambda + \frac{w^2}{f^2}}} \sinh \sqrt{\lambda + \frac{w^2}{f^2}}} \tag{21}$$

where

$$\lambda = \frac{w^2}{D_p} s \quad \text{and} \quad T = \frac{D_p}{w^2} t.$$

There are two approaches to finding the inverse transform. For a homogeneous-base transistor, Schaffner and Suran[10] have expanded the hyperbolic function into a series of exponentials which is the transform of a series of error functions. However, the series converges slowly for large values of time and it is not possible to solve for switching times. Another objection for the present purpose is that when the method is used on (21), complex error functions result. It is more fruitful to perform a partial fraction expansion which involves finding the residues at the poles.

The above function has poles at $\lambda=0$ and at

$$\cosh \sqrt{\lambda + \frac{w^2}{f^2}} = \cos \sqrt{-\lambda - \frac{w^2}{f^2}}$$

$$= \frac{-w/f \sin \sqrt{-\lambda - \frac{w^2}{f^2}}}{\sqrt{-\lambda - \frac{w^2}{f^2}}}$$

or

$$\frac{-y}{w/f} = \tan y$$

where

$$y = \sqrt{-\lambda - \frac{w^2}{f^2}}. \tag{22}$$

Table I column (a) gives λ_{1n}, the values of λ which satisfy this equation, vs w/f.

The inversion of (21) is accomplished as follows:

[9] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y., p. 226; 1942.
[10] Schaffner and Suran, *op. cit.*

TABLE I

NUMERICAL RESULTS

Common Base

w/f	n	(a) λ_{1n}	(b) K_{1n}	(c) $\frac{T_0(0)}{T_0(w/f)}$
0	0	$-(\pi/2)^2$	$-4/\pi$	1
	1	$-(3\pi/2)^2$	$+4/3\pi$	
	2	$-(5\pi/2)^2$	$-4/5\pi$	
2	0	-9.2395	-2.2665	3.05
	1	-29.8745	$+2.1948$	
	2	-69.544	-1.6235	
4	0	-22.607	-5.7031	5.77
	1	-44.666	$+9.6242$	
	2	-84.938	-9.1722	
	3	-144.478	$+7.8609$	
	4	-223.590	-6.6010	

Common Emitter

w/f	Sinusoidal common base		$\alpha_n=0.95$		$\alpha_n=0.90$	
	(d)	(e)	(f)	(g)	(h)	(i)
	κ	$\frac{\omega_\alpha(w/f)}{\omega_\alpha(0)}$	λ_1	$\frac{\lambda_1(w/f)}{\lambda_1(0)}$	λ_1	$\frac{\lambda_1(w/f)}{\lambda_1(0)}$
0	2.4324	1	-0.1	1	-0.2	1
2			-0.2697	2.697	-0.5478	2.739
4	17.67	7.26	-0.4662	4.662	-0.953	4.76

w/f	Sinusoidal common emitter		Saturation
	(j)	(k)	(l)
	κ_e	$\frac{\omega_b(w/f)}{\omega_b(0)}$	$f(w/f)$
0	2.0169	1	1.0
2			0.299
4	9.17	4.55	0.0218

$$\frac{\bar{I}(\lambda, w)}{\Delta I_e} = \frac{e^{w/f}}{\lambda\left(\cos y + \frac{w/f}{y}\sin y\right)}$$

$$= \frac{K_0}{\lambda} + \sum_{n=0}^{\infty} \frac{K_{1n}}{\lambda - \lambda_{1n}} \tag{23}$$

$$K_0 = \frac{e^{w/f}}{\cos j\frac{w}{f} + \frac{1}{j}\sin j\frac{w}{f}} = 1 \tag{24}$$

$$K_{1n} = \left.\frac{e^{w/f}(\lambda - \lambda_{1n})}{\lambda\left(\cos y + \frac{w/f}{y}\sin \lambda\right)}\right|_{\substack{\lambda=\lambda_{1n}\\ y=y_{1n}}}$$

$$= \frac{2e^{w/f}y_{1n}}{\lambda_{1n}\left[\left(1+\frac{w/f}{y_{1n}^2}\right)\sin y_{1n} - \frac{w/f}{y_{1n}}\cos y_{1n}\right]}. \tag{25}$$

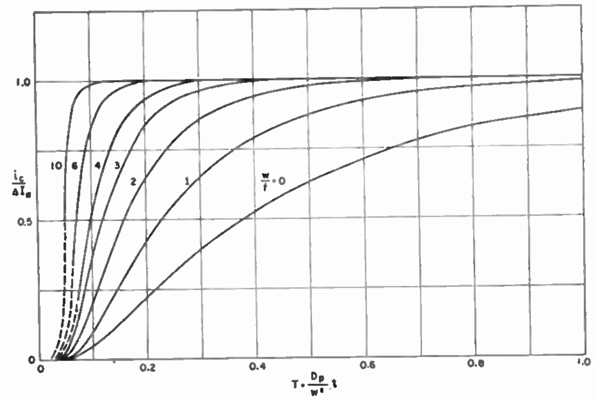

Fig. 2—Collector current vs normalized time for step of emitter current with built-in field as a parameter.

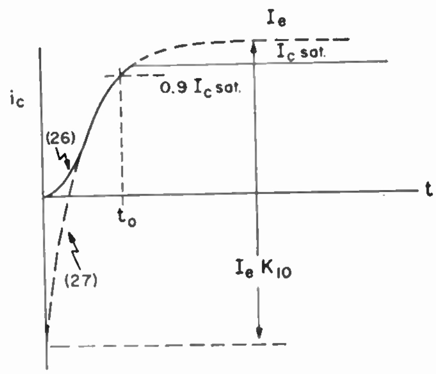

Fig. 3—Method of finding rise time.

Values of the residues, K_{1n}, are given in Table I column (b). The inversion of (32) gives

$$\frac{i(T, w)}{\Delta I_e} = 1 + \sum_{n=0}^{\infty} K_{1n}e^{\lambda_{1n}T}. \tag{26}$$

Fig. 2 is a plot of (26) with parameter w/f. The infinite series converges very slowly near the origin, becoming worse for large values of w/f. The dotted lines represent portions of the curve where the series converged so slowly that it was necessary to estimate. For large values of w/f the delay represents the major portion of the 0–90 per cent rise time. It can be shown that if diffusion is negligible compared to drift, the response is a delayed step with delay

$$T = \frac{1}{2w/f}.$$

For large values of w/f the actual response approaches this delayed step. To find rise times, the series in (26) is terminated after one term.

$$\frac{i(T, w)}{\Delta I_e} = 1 + K_{10}e^{\lambda_{10}T}. \tag{27}$$

This result is seen in Fig. 3. The collector current is considered to rise from zero to the final value, $I_{c\,sat}$. The 0–90 per cent rise time is found using Fig. 3 and (27).

$$I_e(1 + K_{10}e^{\lambda_{10}T}) = 0.9\, I_{c\ \mathrm{sat}}$$

$$T_0 = \frac{D_p}{w^2}\, t_0 = \frac{1}{-\lambda_{10}} \ln \frac{(-K_{10})I_e}{I_e - 0.9\, I_{c\ \mathrm{sat}}}. \qquad 28)$$

For the homogeneous-base case, $w/f=0$, (28) reduces to the equation of Ebers and Moll,[11] with the differences that α has been assumed to be one, the opposite sign convention is used on I_e, and the constants are slightly different so as to take account of the delay present in the common-base mode.

Eq. (28) must be used with caution for large values of w/f and for I_e larger than about twice $I_{c\ \mathrm{sat}}$ because the terminated series, (27), becomes inaccurate. However, for $I_e = I_{c\ \mathrm{sat}}$ the equation is accurate. Thus, a common-base transient improvement factor can be defined as the ratio of the 0–90 per cent rise time of a homogeneous-base transistor to that of a graded-base transistor.

$$\frac{T_0(0)}{T_0\left(\dfrac{w}{f}\right)} = \frac{-\lambda_{10}}{\pi^2/4}\, \frac{\ln (10)(4/\pi)}{\ln (10)(-K_{10})}. \qquad (29)$$

Values are given in Table I column (c) and are plotted in Fig. 4.

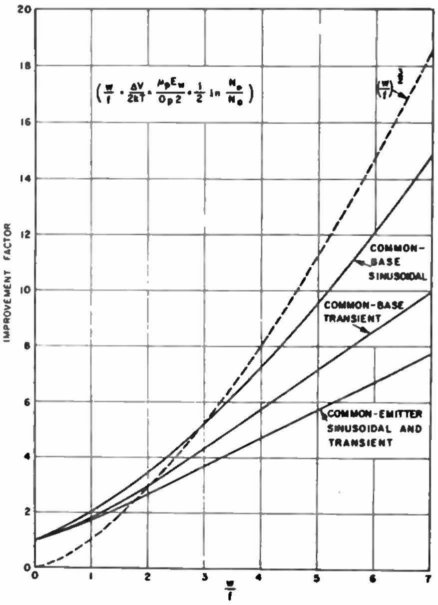

Fig. 4—Improvement factors for sinusoidal response (3-db frequency) and transient response (0–90 per cent for step of input current) of drift transistors vs w/f (for same base width).

Sinusoidal Case

The sinusoidal transfer function previously obtained by Krömer[2] results when s is replaced by $j\omega$ in (20).

$$\frac{I(j\omega, w)}{I_e(j\omega)} = \frac{e^{w/f}}{\cosh \sqrt{\dfrac{w^2}{D_p} j\omega + \dfrac{w^2}{f^2}} + \dfrac{w/f}{\sqrt{\dfrac{w^2}{D_p} j\omega + \dfrac{w^2}{f^2}}} \sinh \sqrt{\dfrac{w^2}{D_p} j\omega + \dfrac{w^2}{f^2}}}. \qquad (30)$$

The frequency at which $|\alpha|$ is 3 db down, ω_α, may be expressed in terms of the dimensionless constant, κ.

$$\omega_\alpha = \kappa \frac{D_p}{w^2}. \qquad (31)$$

It may be found by solving (30) for the value of

$$\frac{w^2}{D_p} j\omega_\alpha = j\kappa$$

for which $|\alpha|$ is 3 db down. Values of κ are given in Table I column (d) for $w/f=0$ and 4. Column (e) gives the improvement factor

$$\frac{\kappa(w/f)}{\kappa(0)} = \frac{\omega_\alpha(w/f)}{\omega_\alpha(0)}$$

which is plotted in Fig. 4. For comparison the improvement factor $(w/f)^{3/2}$ obtained by Krömer[12] is shown as a dotted line.

The magnitude and phase of α vs ω/ω_α is shown in Fig. 5 for $w/f=0$ and 4. Lee[13] has plotted the magnitude of α vs ω for w/f between plus and minus 3 using (30) as derived by Krömer.

The first-order approximation[14] for α with $w/f=0$ can be obtained from (30). The first two terms of the power series for cosh x are retained.

$$\alpha(j\omega) = \frac{1}{1 + j\dfrac{w^2}{2D_p}} = \frac{1}{1 + j\dfrac{\omega}{\omega_\alpha}} \qquad (32)$$

where $\kappa = 2$.

The magnitude and phase of this approximation are also shown in Fig. 5. Middlebrook[15] has also plotted this approximation and the exact function for $w/f=0$. The additional phase shift introduced by the field is related to the relatively large delay found in the step response.

[11] J. J. Ebers and J. L. Moll, "Large-signal behavior of junction transistors," Proc. IRE, vol. 42, pp. 1761–1772; December, 1954.

[12] H. Krömer, "The Drift Transistor" in "Transistors I," RCA Labs., Princeton, N. J.; 1956.

[13] C. A. Lee, "A high frequency diffused base germanium transistor," *Bell Sys. Tech. J.*, vol. 35, pp. 23–34; January, 1956.

[14] E. L. Steele, "Theory of alpha for p-n-p diffused junction transistors," Proc. IRE, vol. 40, pp. 1424–1428; November, 1952.

[15] R. D. Middlebrook, "An Introduction to Junction Transistor Theory," John Wiley and Sons, Inc., New York, N. Y., p. 200; 1957.

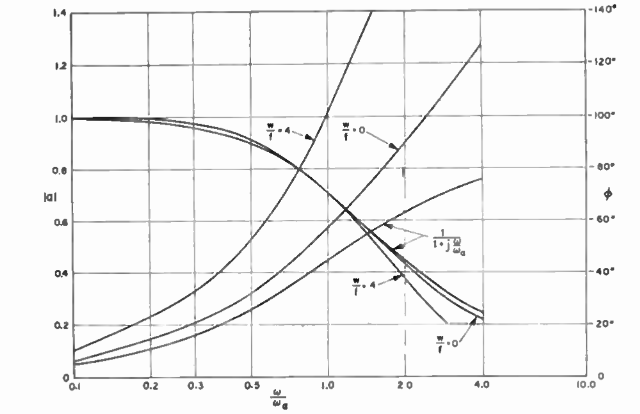

Fig. 5—Absolute value and phase of α vs ω/ω_α with parameter w/f.

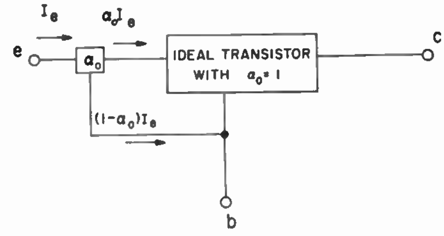

Fig. 6—Assumption regarding α in common-emitter configuration.

and

$$I(s, w) = \frac{\Delta I_b e^{w/f}}{s} \frac{1}{\cosh \Gamma w + \dfrac{w/f}{\Gamma w} \sinh \Gamma w - \alpha e^{w/f}}. \tag{35}$$

Next, the time scale is normalized as before.

$$\bar{I}(\lambda, w) = \frac{\Delta I_b e^{w/f}}{\lambda} \frac{1}{\cosh \sqrt{\lambda + \dfrac{w^2}{f^2}} + \dfrac{w/f}{\sqrt{\lambda + \dfrac{w^2}{f^2}}} \sinh \sqrt{\lambda + \dfrac{w^2}{f^2}} - \alpha e^{w/f}}. \tag{36}$$

Common-Emitter Active Solution

Transient Case

The assumption $\alpha = 1$ will obviously not do for the common-emitter case. However, as mentioned in the common-base solution, the effect of α may be separated out. The assumption is much better for the common-emitter solution because the error is small compared to the slower time constant obtained in common-emitter operation. The three components of α, emitter efficiency, surface recombination, and volume recombination, are lumped into a frequency independent α_0 effectively in cascade with the emitter as shown in Fig. 6.

The boundary conditions applied to (15) for a step of base current are

$$\begin{cases} P(s, w) = 0 \\ I(s, 0) = \alpha \left[\dfrac{\Delta I_b}{s} + I(s, w) \right] \end{cases} \tag{33}$$

This function has a pole at $\lambda = 0$, one for $(\lambda + w^2/f^2)$ real, and an infinite number for $(\lambda + w^2/f^2)$ imaginary or complex. The two major poles are so close together compared to the outer poles that the latter may be neglected. The location of λ_1, the pole nearest the origin, is given in Table I columns (f) and (h) for $\alpha = 0.9$ and 0.95. These values were found by a trial and error solution of the denominator of (36). The inverse transform of (36) is found by letting

$$\bar{I}(\lambda, w) = \frac{K_0}{\lambda} + \frac{K_1}{\lambda - \lambda_1} \tag{37}$$

and solving for the residues in the usual manner.

$$K_0 = \frac{\Delta I_b e^{w/f}}{\cosh \dfrac{w}{f} + \sinh \dfrac{w}{f} - \alpha e^{w/f}} = \Delta I_b \frac{\alpha}{1 - \alpha} \tag{38}$$

$$K_1 = \frac{\Delta I_b 2 \alpha e^{w/f} \sqrt{\lambda_1 + \dfrac{w^2}{f^2}}}{\lambda_1 \left[\left(1 - \dfrac{w/f}{\left(\lambda_1 + \dfrac{w^2}{f^2} \right)} \right) \sinh \sqrt{\lambda_1 + \dfrac{w^2}{f^2}} + \dfrac{w/f}{\sqrt{\lambda_1 + \dfrac{w^2}{f^2}}} \cosh \sqrt{\lambda_1 + \dfrac{w^2}{f^2}} \right]}. \tag{39}$$

with solution

$$P(s, x) = \frac{\Delta I_b e^{x/f}}{q D_p \Gamma s} \frac{\sinh \Gamma (w - x)}{\cosh \Gamma w + \dfrac{w/f}{\Gamma w} \sinh \Gamma w - \alpha e^{w/f}} \tag{34}$$

Numerical evaluation of these residues shows that K_1 is within a few per cent of $-K_0$. Neglecting the outer poles is therefore justified. The inversion of (37) gives

$$\frac{\bar{i}(T, w)}{\dfrac{\alpha}{1 - \alpha} I_b} = 1 - e^{\lambda_1 T}. \tag{40}$$

A solution for the switching time gives

$$T_0 = \frac{D_p}{w^2} t_0 = \frac{1}{(-\lambda_1)} \ln \frac{I_b}{I_b - 0.9 \dfrac{1-\alpha}{\alpha} I_{c\,\text{sat}}} \tag{41}$$

which is the same as that obtained by Ebers and Moll except for the modified time constant. The improvement in transient response,

$$\frac{\lambda_1(w/f)}{\lambda_1(0)},$$

is given in Table I columns (g) and (i) and is plotted in Fig. 4.

Sinusoidal Case

The sinusoidal transfer junction may be obtained from (35) by replacing s with $j\omega$.

$$\frac{I(j\omega, w)}{I_b(j\omega)} = \frac{\alpha e^{w/f}}{\cosh \sqrt{\dfrac{w^2}{D_p} j\omega + \dfrac{w^2}{f^2}} + \dfrac{w/f}{\sqrt{\dfrac{w^2}{D_p} j\omega + \dfrac{w^2}{f^2}}} \sinh \sqrt{\dfrac{w^2}{D_p} j\omega + \dfrac{w^2}{f^2}} - \alpha e^{w/f}}. \tag{42}$$

The frequency at which $|\beta|$, the common-emitter current gain, is 3 db down is defined as ω_β.

$$\omega_\beta = \kappa_e(1-\alpha) \frac{D_p}{w^2}. \tag{43}$$

The factor κ_e may be found as before and is given in Table I column (j) for $\alpha = 0.95$. Column (k) gives the improvement factor

$$\frac{\kappa_e\left(\dfrac{w}{f}\right)}{\kappa_e(0)} = \frac{\omega_\beta\left(\dfrac{w}{f}\right)}{\omega_\beta(0)}$$

which is plotted in Fig. 4 vs w/f. Notice that it coincides with the common-emitter transient response improvement factor. The first-order approximation for β with $w/f = 0$ can be obtained from (42).

$$\beta(j\omega) = \frac{\alpha}{1 + \dfrac{w^2}{2D_p} j\omega - \alpha} = \frac{\dfrac{\alpha}{1-\alpha}}{1 + \dfrac{j\omega}{\omega_\beta}} \tag{44}$$

which gives the value of κ_e as 2. Fig. 7 shows the magnitude and phase of $\beta(j\omega)/\beta(0)$ for $w/f = 0$, and 4, and of the first-order approximation as given in (44).

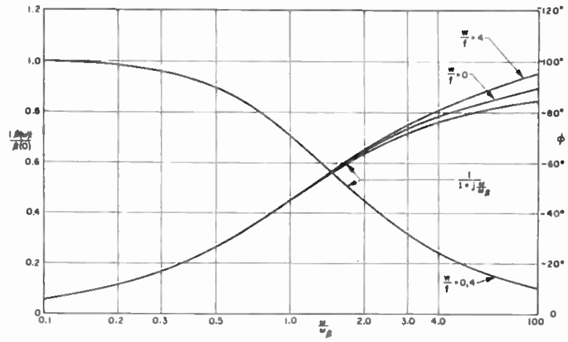

Fig. 7—Absolute value and phase of β vs ω/ω_β with parameter w/f (common-emitter configuration).

The first neglected term of the expansion for cosh x is $(1-\alpha)$ times as large as the corresponding term in the common base (32). This fact explains why the first-order approximation given in (44) is so good. It also accounts for the near exponential common-emitter transient response.

Saturation

In the saturation region the collection junction becomes forward biased and the hole density at $x = w$ is allowed to rise from zero. A straightforward application of the boundary conditions

$$\begin{cases} I(s, 0) = \alpha \dfrac{\Delta I_e}{s} \\ I(s, w) = 0 \end{cases}$$

to (15) could be used. However, the assumption of a one-dimensional transistor is usually poor in saturation operation. The method used by Ebers and Moll[11] has been successful for the homogeneous-base transistor and will be used here. The transient solution is broken up into a normal and an inverse component as shown in Fig. 8, and the effects of the transistor not being one dimensional are included in the characteristics of the inverse component. The result of the separation, shown in Fig. 9, is a fictitious circulating current which is useful in introducing the effect of α_N and α_I, the forward and inverse current gains, respectively. The curvature shown is due to the built-in field in a graded-base transistor.

The hole density plots may be replaced by (20), the common-base transfer function. The sign of f is negative for the inverted solution.

Let the forward transfer function be F_1 and the inverse function be F_2. Then

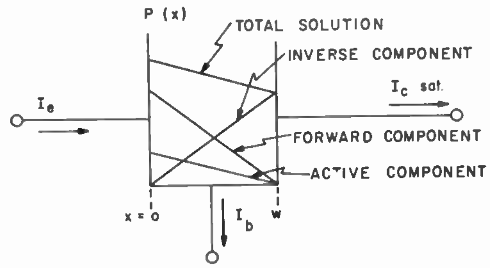

Fig. 8—Base hole density in saturation and manner of decomposition into three components.

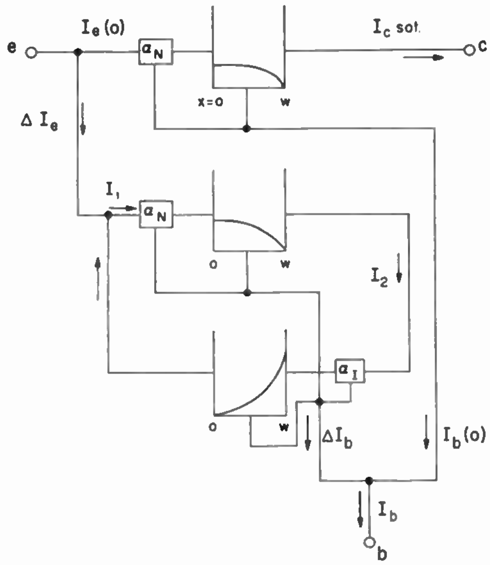

Fig. 9—The three components of hole density in saturation.

$$I_2 = F_1 I_1 = F_1(\Delta I_e + F_2 I_2)$$

$$I_2 = \frac{F_1 \Delta I_e}{1 - F_1 F_2}.$$

The expression for I_2 is then substituted into (19) (with f replaced by $-f$) and the transform of the hole density at the collector obtained.

imaginary or complex. The outer poles will be neglected as in the common-emitter solution. The first pole from the origin, λ_1, may be found for $w/f = 0$ by retaining two terms of the power series for $\cosh \sqrt{\lambda_1}$.

$$\cosh^2 \sqrt{\lambda_1} = \alpha_N \alpha_I \cong \left(1 + \frac{\lambda_1}{\alpha}\right)^2$$

$$s_1 = \frac{D_p}{w^2} \lambda_1 = -\omega(1 - \sqrt{\alpha_N \alpha_I}) \quad (46)$$

which is the same as the corresponding expression of Ebers and Moll[11]

$$\frac{\omega_N \omega_I}{\omega_N + \omega_I}(1 - \alpha_N \alpha_I) \quad (47)$$

if $\omega_N = \omega_I$ and $\alpha_N \alpha_I$ is near one. By considering the forward and inverse components to have different effective base widths it is possible,[16] with a few approximations, to derive (47) from the transform of the collector hole density. For the drift transistor, the bracketed term in the denominator of (45) must be solved numerically. However, by expressing the hyperbolic functions in terms of exponentials and by retaining only the major terms, an expression for s_1 is obtained which agrees with the numerical solution for $w/f > 2$.

$$s_1 = -\frac{D_2}{w^2}[f(w/f)](1 - \alpha_N \alpha_I)$$

$$= -\frac{D_p}{w^2}\left(4 \frac{w^2}{f^2} e^{-2w/f}\right)(1 - \alpha_N \alpha_I). \quad (48)$$

Values of $f(w/f)$ are given in Table I column (1). Notice that the presence of the built-in field tends to reduce this factor and thus to lengthen storage time. The effect of this field on storage time has been the subject of much speculation. At first glance one might expect the presence of the field to sweep out the holes more quickly. A qualitative explanation for the above

$$\bar{P}(\lambda, w) = \frac{\Delta I_e \alpha_N \alpha_I e^{w/f} \sinh \sqrt{\lambda + \frac{w^2}{f^2}}}{q D_p \lambda \sqrt{\lambda + \frac{w^2}{f^2}} \left[\cosh^2 \sqrt{\lambda + \frac{w^2}{f^2}} - \left\{\frac{w/f}{\sqrt{\lambda + \frac{w^2}{f^2}}}\right\}^2 \sinh^2 \sqrt{\lambda + \frac{w^2}{f^2}} - \alpha_N \alpha_I\right]}. \quad (45)$$

The function has a pole at $\lambda = 0$, one for

$$\sqrt{\lambda + \frac{w^2}{f^2}}$$

real, and an infinite number for

$$\sqrt{\lambda + \frac{w^2}{f^2}}$$

results can be obtained by examination of (47), the expression of Ebers and Moll. The effect of the field is to increase ω_N and decrease ω_I. The resultant factor is roughly equivalent to $f(w/f)$ in (48).

For the same base width and $(1 - \alpha_N \alpha_I)$, then, the graded-base transistor would have a longer storage time than the homogeneous base. However, measurements

[16] Johnston, *op. cit.*, App. IV.

show that α_I is quite low. The field tends to force the holes away from the emitter and to the surface. It also causes the hole density to build up at the collector resulting in an increase in volume recombination and a decrease in the efficiency of the collector as an emitter. The three factors combine to reduce α_I to about 0.6 for the Philco micro-alloy drift transistor 2N501. A rough check on the theory may be obtained by comparing this transistor to a homogeneous-base transistor of similar geometry, the Philco micro-alloy 2N393. The ratio of storage time, homogeneous to graded, is given by the ratio of (48) to (47). The values for ω and α given in Table II are median values based on measurements.

TABLE II
MEASURED PARAMETERS REGARDING STORAGE TIME

	Homogeneous Base 2N393		Graded Base 2N501
ω_N	$(2\pi)(41+10^6)$	$2\frac{D_p}{w^2}$	$(2\pi)(41+10^6)(2.6)$
ω_I	$(2\pi)(8+10^6)$	w/f	2.3−3.4
$\alpha_N\alpha_I$	0.916	$\alpha_N\alpha_I$	0.6

The width of the constant-field portion of the base is estimated to be 0.39 times the base width of the homogeneous-base unit. The value of w/f is difficult to measure but lies in the range given. The ratio of storage times, given by (48) over (47), is in the range 1.7 to 7.9 while actual measurement gives about 2.3. It is encouraging that the theory gives numbers in the right range and perhaps more careful measurements would show better agreement.

Conclusion

The short-circuit transient response of a drift transistor has been found for a step of input current. The method used was to apply various boundary conditions to the one-dimensional partial differential equation which was then solved by a Laplace transform technique. The transform, containing hyperbolic functions of the square root of s, was expanded into a series of partial fractions which were then inverted into a series of exponential functions of time. The improvement in rise time due to the built-in field was obtained by neglecting the higher order terms and is presented in numerical form. This improvement is shown to be less in the common-emitter than in the common-base configuration. The improvement in common-base cutoff frequency is shown to be less than the value given by Krömer.

Since the common-emitter performance is not directly related to the common-base performance, the α-cutoff frequency is not a good way to specify the high-frequency performance in common-emitter circuits. A measurement coming into use is the frequency at which the short-circuit common-emitter current gain is unity. For ease of measurement, this frequency is usually K times the frequency at which the gain is K where K is about 5 to 10. This measurement is variously known as f_T, fh_{fe}, and gain-bandwidth product.

The partial differential equation was also solved, subject to the conditions of saturation. In this case, the "improvement factor" turned out to be much less than unity. However, it was found that many of the assumptions, such as constant mobility, constant field, and one dimensionality, break down in saturation. An attempt was made to include some of these effects through the inverse current gain and a much better experimental agreement was obtained.

Acknowledgment

The author is indebted to Prof. R. B. Adler, Dr. D. J. Eckl, and C. T. Kirk for many helpful suggestions and to Mrs. A. Harvey for the numerical evaluation of (30).

Terminal Properties of Magnetic Cores*

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Summary—**The reversal of magnetization of cores with rectangular hysteresis loop is investigated. For materials whose thickness is more than one thousandth of an inch, the switching is determined from the static loop and the solution of the appropriate field problem; for thinner materials the *B-H* concept is not sufficient to determine the instantaneous state of magnetization and to characterize the core as a circuit element a new approach must be used. In this paper it is shown that the terminal properties of these cores can be obtained from their dynamic step response; the analysis is based on the physics of magnetization and is experimentally verified.**

I. Introduction

In recent years, small magnetic cores with rectangular hysteresis loop are being used widely, particularly in digital computers and similar data processing systems, as shift registers, counters, storage devices and other logical elements.[1] In these applications the

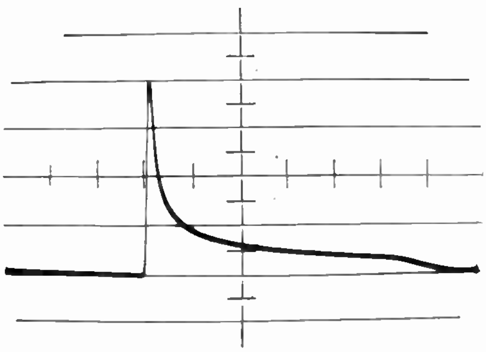

Fig. 1—Output voltage of a 4-mil Orthonik core driven by a step-current driver. Calibration: 60 mv/div, 3 μsec/div.

cores are switched totally or partially from one direction of magnetization to the opposite. In the simplest form of operation the magnetization reverses through the application of a constant magnetizing current; Fig. 1 and Fig. 2 give typical voltage outputs for cores of thickness 4 mil and $\frac{1}{8}$ mil, respectively. For the analysis and design of a system containing such elements one should be able to describe their terminal properties with the smallest possible number of parameters, that is, to "predict" their circuit performance under any driving conditions in terms of these parameters.

The magnetic properties of a core are usually derived from its static hysteresis loop; in interpreting it one assumes that for a given H, due to all causes (external drivers, eddy currents, and free magnetic poles), B takes instantly the value given by the B-H loop. The above interpretation is adequate if the times involved are much longer than 1 microsecond; this is the case for the switching of cores of the order of 2 mils or more. For such cores, which in the following will be called "thick," the state of magnetization can be predicted from the B-H curve and the solution of the appropriate field equations. Since the B-H curve is nonlinear, in general one can give only a numerical solution to this problem; however, for cores with ideally rectangular hysteresis loop (see Fig. 3) an analytic solution is possible.[2,3]

Fig. 2—Output voltage of a thin core driven by a step-current driver. Core: 80 wraps of $\frac{1}{8}$-mil 4-79 Mo-Permalloy. Calibration: 0.75 v/div, 1 μs/div.

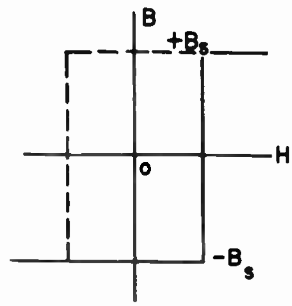

Fig. 3—Idealized rectangular hysteresis loop.

For "thin" magnetic cores (of the order of $\frac{1}{8}$ mil) the eddy-current effect is negligible; the application of a current establishes instantly the same H_i throughout the cross section of the material. With the usual interpretation of the B-H curve one would expect the magnetization to reach instantly the value $B(H_i)$ and the voltage output would be an impulse; but this is not the

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[1] W. N. Papian, "A coincident-current magnetic memory cell for the storage of digital information," Proc. IRE, vol. 40, pp. 475–478; April, 1952.

[2] W. MacLean, "Theory of strong electromagnetic waves in massive iron," *J. Appl. Phys.*, vol. 25, pp. 1267–1270; October, 1954.

[3] A. Papoulis, "Penetration of an electromagnetic wave into a ferromagnetic material," *J. Appl. Phys.*, vol. 25, pp. 169–176; February, 1954.

case as one can see in Fig. 2. Thus, the familiar concept of the B-H curve fails to describe the dynamics of magnetization of thin cores; to describe the time-lag between the application of a field and the resulting magnetization, and to determine the output signals, a new theory must be developed.[4-6]

In Section II the switching of "thick" cores is reviewed; it is assumed that B takes instantly the value given by the static B-H curve, and the resulting field problem is solved for square-loop materials. Various driving and loading conditions are considered and the developed results are experimentally verified.

In Section III the reversal of magnetization of "thin" cores is considered; since the eddy currents are negligible, H is uniform throughout the material but B does not take instantly the value given by the B-H curve. It is shown that the output due to an arbitrary driver can be obtained from the step response of the core; its circuit properties are thus determined from a single terminal measurement. The analysis is based on the domain theory of magnetization, however, the results describe the terminal properties of the core and can be followed without any physical considerations; they are justified by the agreement between theory and experiment.

In Section IV the analysis is applied to cores of intermediate thickness. Due to the eddy currents, H is no longer known and B is a function of H and t as in Section III; a field problem results whose solution gives B, H and the output waveforms.

It is shown that in all three cases there exists a unique relationship between the flux Φ in the core and the applied coulomb-turns

$$Q = \frac{1}{l}\sum_{k=1}^{n} N_k \int_0^t i_k(\tau)d\tau$$

per meter, where i_k is the current through the kth winding having N_k turns, and l the length of the magnetic path. From this relationship the terminal properties of the core are derived without any further consideration of the field problem or the internal mechanism of magnetization; thus, for the dynamic study of switching the B-H curve is replaced by the Φ-Q curve.

II. Eddy Currents

The Voltage-Current Characteristic of the Core

The cores under consideration are formed from a toroidal tape of thickness d, height h, and radius r (Fig. 4); d is of the order of 1 mil, h and r a fraction of an inch. Since $d \ll r$, h the resulting field can be approximated by a uniform plane wave with H in the z direction and E in the y direction;[7] x is the direction of propagation and the origin is taken at the surface of the material. Because of symmetry it suffices to determine E and H in the $(0, d/2)$ interval.

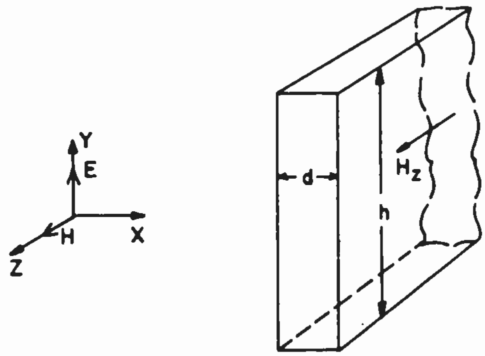


Fig. 4—Cross section of magnetic lamination.

With the above assumptions Maxwell's equations take the form

$$\frac{\partial E}{\partial x} = -\frac{\partial B}{\partial t} \tag{1}$$

$$\frac{\partial H}{\partial x} = -\sigma E \tag{2}$$

where σ is the conductivity of the material and B a known function of H given by the B-H curve of Fig. 3, thus

$$B = -B_s \quad \text{for} \quad H < H_c$$
$$B = +B_s \quad \text{for} \quad H > H_c.$$

For initial and boundary conditions we have

$$H(x, 0) = 0 \tag{3}$$

$$E\left(\frac{d}{2}, t\right) = 0 \tag{4}$$

and

$$H(0, t) = H_i(t) = \frac{1}{l}\sum_{k=1}^{n} N_k i_k(t) \tag{5}$$

where i_k is the current through the kth winding having N_k turns and $l = 2\pi r$ the core circumference; (4) follows from the core symmetry. We assume that the core is in the $-B_s$ state at $t=0$ and that the applied field $H_i(t)$ is greater than H_c thus causing a reversal of magnetization from $-B_s$ to $+B_s$; this reversal will start at the surface of the core and will propagate into the interior reaching the point x_s at time t. Hence

[4] N. Menyuk and J. B. Goodenough, "Magnetic materials for digital computer components, I—A theory of flux reversal in polycrystalline ferromagnetics," *J. Appl. Phys.*, vol. 26, pp. 8–18; January, 1955.

[5] N. Menyuk, "Magnetic materials for digital computer components, II—Magnetic characteristics of ultra-thin molybdenum-permalloy cores," *J. Appl. Phys.*, vol. 26, pp. 692–697; June, 1955.

[6] K. H. Stewart, "Ferromagnetic Domains," Cambridge University Press, Cambridge, Eng.; 1954.

[7] D. F. Hunt, "Saturation Time for Deltamax Transformer," Burroughs Corp., Paoli, Pa., Internal Rep. IR78; July, 1951.

$$B = + B_s \quad \text{for} \quad x < x_s$$
$$B = - B_s \quad \text{for} \quad x > x_s. \tag{6}$$

B is discontinuous at $x = x_s$ but constant in every other point, hence

$$\frac{\partial E}{\partial x} = 0$$

for every

$$x \neq x_s.$$

Therefore, $E(x, t)$ is constant in the $(0, x_s)$ and $(x_s, d/2)$ intervals; the constant value of E in the $(x_s, d/2)$ interval must be zero since $E(d/2, t) = 0$ [see (4)] and the value in the $(0, x_s)$ interval we shall denote by $E_s(t)$. Thus

$$E(x, t) = 0 \qquad \text{for} \quad x > x_s \tag{7}$$

$$E(x, t) = E_s(t) \quad \text{for} \quad x < x_s \tag{8}$$

(see Fig. 5). To determine $E_s(t)$ we integrate (1)

$$E(0, t) = E_s(t) = \int_0^{d/2} \frac{\partial B}{dt} dx. \tag{9}$$

But

$$\int_0^{d/2} B dx = B_s x_s - B_s \left(\frac{d}{2} - x_s \right)$$

hence

$$\frac{d}{dt} \int_0^{d/2} B dx = 2B_s \frac{dx_s}{dt}$$

therefore

$$E_s(t) = 2B_s \frac{dx_s}{dt}. \tag{10}$$

From (7), (8), and (2) one can readily see that H is constant for $x > x_s$ and varies linearly in the $(0, x_s)$ interval; but $H < H_c$ for $x > x_s$ and $H > H_c$ for $x < x_s$ [see (6)] and since H is continuous, we must have

$$H(x, t) = H_c \quad \text{for} \quad x > x_s$$

and

$$H(x, t) = H_c + (H_i - H_c) \left(1 - \frac{x}{x_s} \right) \quad \text{for} \quad x < x_s. \tag{11}$$

From (2), (8), and (10) we readily derive

$$\frac{H_i(t) - H_c}{\sigma x_s} = E_s(t) = 2B_s \frac{dx_s}{dt}. \tag{12}$$

Thus, we obtained a simple differential equation for x_s whose solution is given by

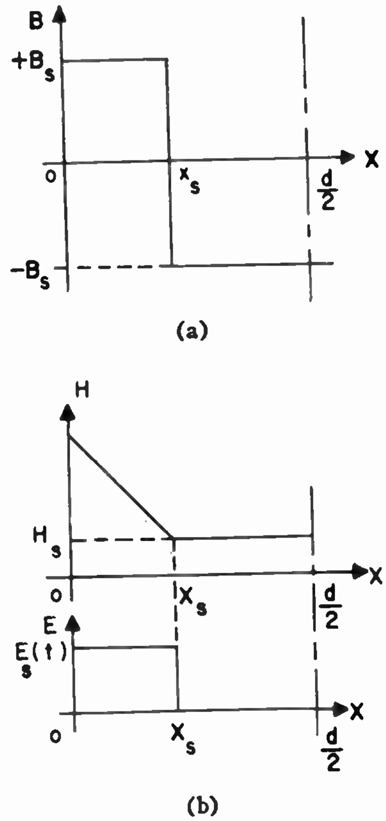


Fig. 5—Variation along x of: (a) magnetic flux density, (b) magnetic and electric field.

$$x_s(t) = \sqrt{\frac{\int_0^t [H_i(\tau) - H_c] d\tau}{\sigma B_s}}. \tag{13}$$

When x_s reaches the center of the core the material is completely switched from $-B_s$ to $+B_s$; the required time t_s we shall call "switching time:"[2]

$$x_s(t_s) = \frac{d}{2}$$

hence [see (13)]

$$\int_0^{t_s} [H_i(\tau) - H_c] d\tau = \frac{\sigma B_s d^2}{4} \tag{14}$$

from which t_s is determined if $H_i(t)$ is known. The voltage output $r(t)$ across a one-turn winding is given by $d\Phi/dt$ where

$$\Phi = 2h \int_0^{d/2} B dx$$

hence [see (9) and (10)]

$$r(t) = 2hE_s(t) = 4hB_s \frac{dx_s}{dt}. \tag{15}$$

From (12), (13), and (15) we have the output

$$r(t) = \frac{2h[H_i(t) - H_c]}{\sqrt{\frac{\sigma}{B_s}\int_0^t [H_i(\tau) - H_c]d\tau}} = \frac{d\Phi}{dt} \tag{16}$$

as a function of the applied field $H_i(t)$.

The Flux Charge per Meter Curve

We define the coulomb-turns per meter Q as the integral of the applied field; thus

$$Q(t) = \int_0^t H_i(\tau)d\tau = \frac{1}{l}\sum_{k=1}^{n} N_k \int_0^t i_k(\tau)d\tau. \tag{17}$$

Assuming $H_i(t) \gg H_c$ we obtain from (16)

$$\Phi + \Phi_s = 4\sqrt{\frac{h\Phi_s Q}{d\sigma}} \tag{18}$$

since

$$\Phi = -\Phi_s = -hdB_s \quad \text{for} \quad Q = 0.$$

The core is switched when $\Phi = \Phi_s$ and the required value of Q is given by

$$Q_s = \frac{d\sigma\Phi_s}{4h}. \tag{19}$$

Thus knowing Q we can determine the flux through the core; (18) characterizes the core completely. For $Q > Q_s$ we clearly have $\Phi = \Phi_s$.

Constant Applied Field

If $H_i(t) = H_a = \text{constant}$, then [see (13)]

$$x_s = \sqrt{\frac{(H_a - H_c)t}{\sigma B_s}}.$$

The switching time t_s and the voltage output $r(t)$ are given by [see (14) and (16)]

$$t_s = \frac{\sigma B_s d^2}{4(H_a - H_c)} \tag{20}$$

and

$$r(t) = \sqrt{\frac{4h^2 B_s(H_a - H_c)}{\sigma t}}. \tag{21}$$

In Fig. 6 the output as given by (21) is plotted; it agrees with the actual response of Fig. 1 except for small t and for $t > t_s$. The discrepancy between the perdicted and the observed response for t close to zero is due to finite rise time of the driver and the finite speed of domain reversal as will be seen later. For a perfectly square loop, $r(t)$ is zero for $t > t_s$. However, in an actual core B is not constant for $H > H_c$ but it increases. Assuming that B increases linearly with H, we can find the output for $t > t_s$ by solving a linear problem with the proper initial

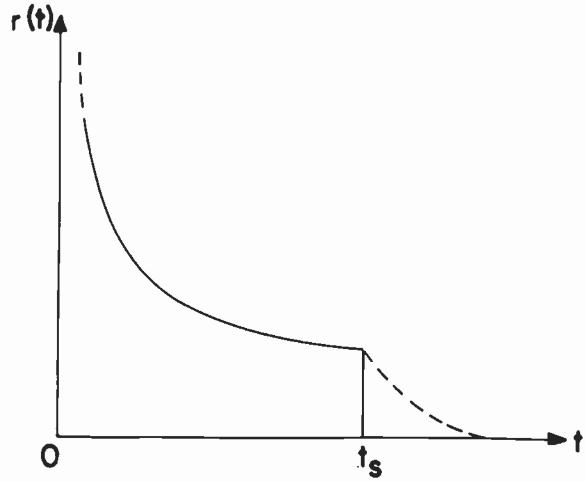


Fig. 6—Predicted voltage output of a thick core driven by a step-current driver.

condition at $t = t_s$. The dotted line in Fig. 6 will result; the knee of the curve in Fig. 1 is reached when x_s reaches the center of the core.

Loaded Core

Next we shall consider a core loaded with a resistance R across an N turn winding and driven by a current driver I. With $r(t)$ the one-turn output, the current through the resistance is given by

$$\frac{Nr(t)}{R}$$

causing a field

$$N_R = \frac{N^2}{R} r(t).$$

The total applied field is given by

$$H_i = H - H_R = H - \frac{2hN^2}{R} E_s(t) \tag{22}$$

[see (5) and (15)] where H is the field due to the driver. With

$$G = \frac{2hN^2}{R} \tag{23}$$

we obtain from (10), (12), and (22)

$$2B_s\sigma x_s \frac{dx_s}{dt} + 2B_s G \frac{dx_s}{dt} = H - H_c. \tag{24}$$

Solving the above equation for x_s and inserting into (15) we obtain the output voltage. The results are plotted in Fig. 7(a) for a constant current driver and various values of R; the dotted lines give the response for $t > t_s$. In Fig. 7(b) the experimentally obtained responses are given.

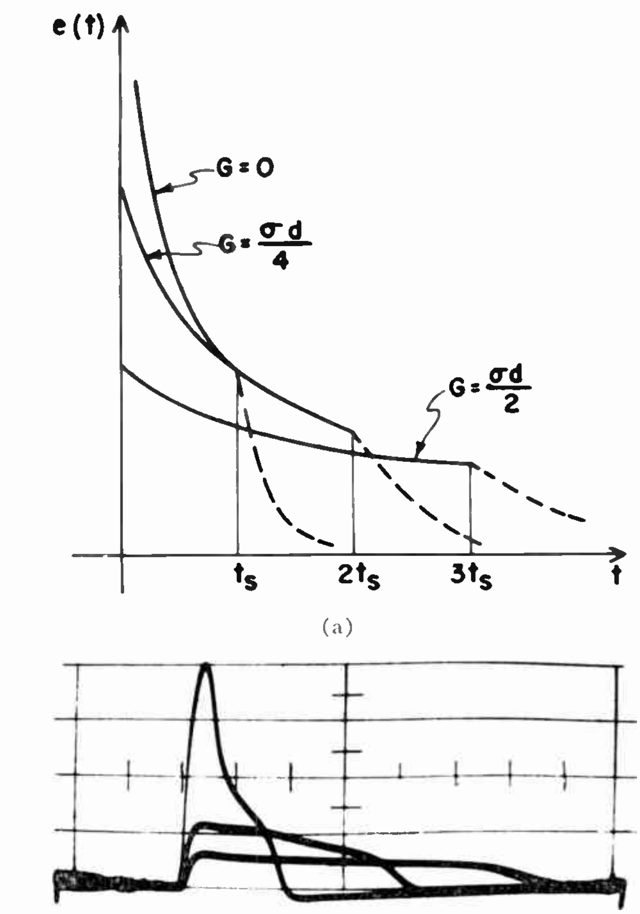


Fig. 7—Voltage output of a loaded thick core driver by a step-current driver: (a) theoretical, (b) experimental. Core: 8 wraps of 1-mil Orthonik. Driving field: 202 AT/m. Read-out winding: 30 turns. Load resistance: $R=\infty$, 42 and 94 ohms. Time scale: 1 μs/div. Voltage scale: 6.9 v/div.

III. Thin Cores

Basic Assumption

It will be assumed that if at a certain time t the material is in the state $B(t)$ and the applied field is $H(t)$, then the rate of increase of B is given by

$$\frac{dB}{dt} = f(B)[H(t) - H_0] \quad \text{for} \quad H(t) > H_0 \tag{25}$$

where H_0 is a constant of the material. In (25) $f(B)$ is a function of the instantaneous state of magnetization and does not depend on the way in which this state is reached; for $H(t) < H_0$ the state of magnetization does not change. A similar relationship was developed empirically by Rising[8] from experimental data obtained mainly through repeated application of short square pulses; it was then observed that the resulting increase in B obeyed (25). In the Appendix it is shown that (25) can be predicted from the domain theory of magnetization, however, for the discussion of the terminal properties of the core it might be accepted as an assumption whose justification will follow from the agreement between the predicted results and the experimentally obtained data. In the following it is seen that, with (25), the voltage-current characteristic of the core driven by an arbitrary driver can be derived from its dynamic step response.

[8] H. K. Rising, "Magnetic-Core Pulse Amplifiers for Digital Computers," Master's thesis, M.I.T., Cambridge, Mass.; 1953.

Voltage-Current Characteristic

From Constant-Current Response: We denote by $a(t)$ the voltage output when a constant field H_a is applied and by $V(t)$ the output resulting from the application of an arbitrary field $H_i(t)$. When $H_i(t)$ is applied, a state B of magnetization is reached at time t, and when H_a is applied the same state is reached at time t; thus we have

$$a(t') = hd\,\frac{dB}{dt'} = f(B)[H_a - H_0] \tag{26}$$

$$V(t) = hd\,\frac{dB}{dt} = f(B)[H_i(t) - H_0]. \tag{27}$$

Considering t and t' as functions of B and eliminating B we obtain t' as a function $t'(t)$ of t. To determine this function we divide (27) by (26);

$$\frac{dt'}{dt} = \frac{H_i(t) - H_0}{H_a - H_0} \tag{28}$$

results, whose solution is given by

$$t' = \int_0^t \frac{H_i(t) - H_0}{H_a - H_0}\,dt. \tag{29}$$

But

$$V(t) = hd\,\frac{dB}{dt} = hd\,\frac{dB}{dt'}\,\frac{dt'}{dt} = a(t')\,\frac{dt'}{dt} \tag{30}$$

therefore

$$V(t) = \frac{H_i(t) - H_0}{H_a - H_0}\,a\left[\int_0^t \frac{H_i(\tau) - H_0}{H_a - H_0}\,d\tau\right]. \tag{31}$$

The above equation establishes the voltage-current characteristic and describes the core as a circuit element. If

$$H_i(t) = H_1 = \text{constant}$$

then

$$V(t) = ka[kt] \quad \text{where} \quad k = \frac{H_1 - H_0}{H_a - H_0}. \tag{32}$$

Thus if the core is driven by constant drivers of various amplitudes, the resulting outputs are similar.

From Constant-Voltage Response: A similar relationship can be obtained in terms of the current response $b(t)$ resulting from the application of a constant voltage V_b across a one-turn winding. With $i_0 = H_0 l$ and $b(t') = H(t')l$ we have from (29)

$$V_b = hd\frac{dB}{dt'} = \frac{hd}{l}f(B)[b(t') - i_0]. \tag{33}$$

If an arbitrary voltage driver $V(t)$ is applied causing a current $i(t)$ then

$$V(t) = hd\frac{dB}{dt} = \frac{hd}{l}f(B)[i(t) - i_0].$$

Reasoning as above we obtain

$$\frac{dt'}{dt} = \frac{V(t)}{V_b}, \qquad V(t) = V_b\frac{i(t) - i_0}{b(t') - i_0}$$

hence

$$i(t) - i_0 = \frac{V(t)}{V_b}\left[b\left(\int_0^t \frac{V(\tau)}{V_b}d\tau\right) - i_0\right]. \tag{34}$$

The Flux Charge per Meter Curve

The flux charge per meter relationship can readily be determined from (25); indeed, by integrating we have

$$\int_{-B_s}^{B} \frac{dB}{f(B)} = \int_0^t [H_i(\tau) - H_0]d\tau.$$

With the charge-turns per meter Q defined as in (17) we have

$$\int_{-\Phi_s/hd}^{\Phi/hd} \frac{dB}{f(B)} = Q - H_0t. \tag{35}$$

If

$$Q \gg H_0t$$

then

$$F(\Phi) = Q \tag{36}$$

where $F(\Phi)$ is the left-hand side in (35). The plot of the Φ-Q curve for a $\frac{1}{8}$-mil 4-79 Mo-Permalloy is given in Fig. 8.

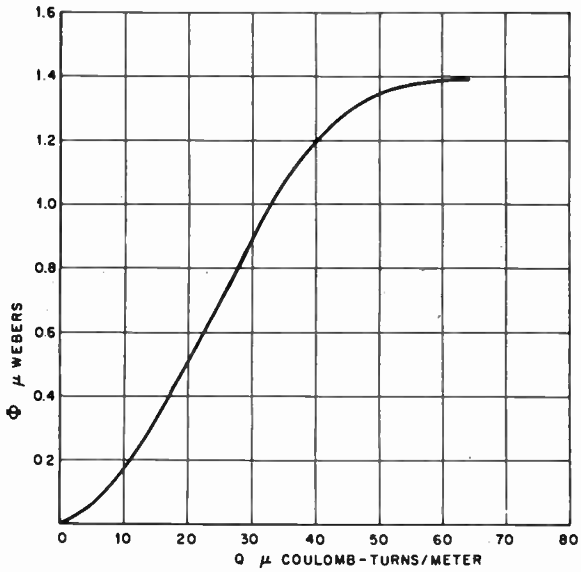


Fig. 8—The flux-coulomb turns per meter curve for thin cores.

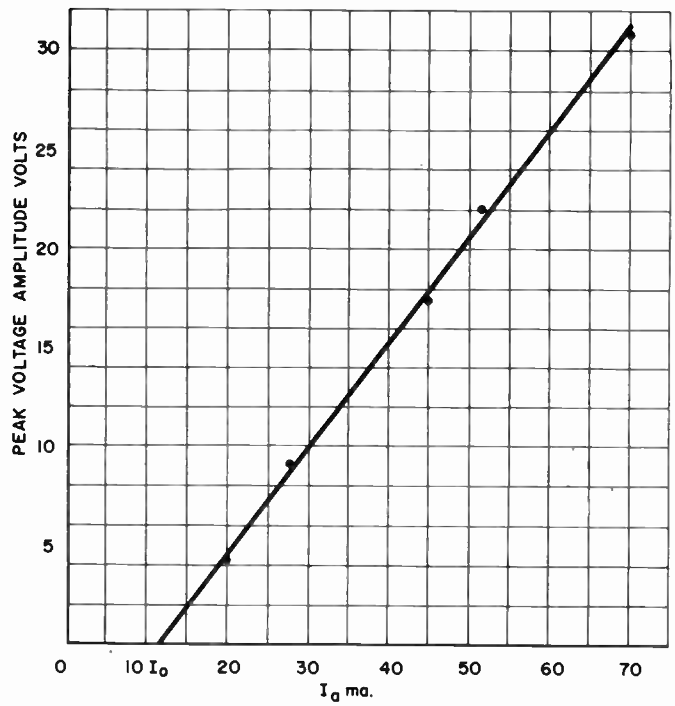


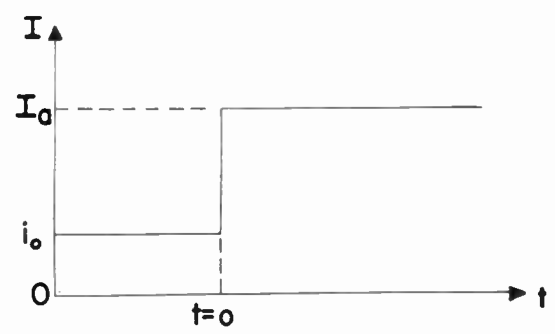
Fig. 9—Plot of peak values of output voltage against driving current.

Experimental Verification of Theoretical Results

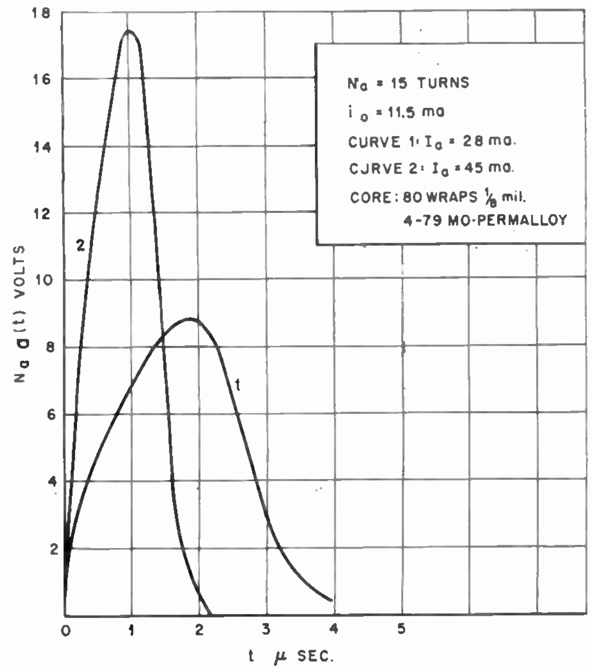
The developed results will be experimentally verified by observing the voltage response $V(t)$ of a thin core driven by a linearly increasing function of time, and comparing with the predicted voltage as given by (31). We have used $a(t)$ rather than $b(t)$ since it is easier to obtain a constant current source. The rise time of the driver was about 0.1 μsec and the magnetic core consisted of 80 wraps of $\frac{1}{8}$-mil 4-79 Mo-Permalloy lamination. The voltage response was observed across the terminals of a 15-turn winding and the core was reset by a separate winding.

For a field $H_i(t) < H_0$ there is a small reversible increase in B which does not satisfy (25); to eliminate its effect, $a(t)$ is obtained by first prepulsing the core with a field equal to H_0 and then applying the switching current. To find H_0 the core is switched with constant current of various magnitudes and the peaks are plotted as functions of the applied currents. From (32) one sees that the points thus plotted should fall on a straight line whose intercept with the current axis will give the current i_0 necessary to establish H_0. From Fig. 9 we see that this is indeed the case; thus we obtain $i_0 = 11.5$ ma. An independent verification of this value is given by applying currents smaller than 11.5; the resulting small flux variation is then reversible. With currents greater than 11.5 ma an irreversible flux variation is observed.

With i_0 established, we apply currents of the form of Fig. 10(a) obtaining $a(t)$ with $I_a = 28$ ma and $I_a = 45$ ma,



(a)



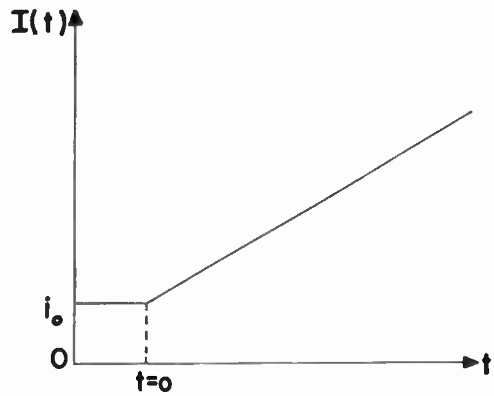
(b)

Fig. 10—Prepulsed step response: (a) driving current, $i_0=11.5$ ma, (b) voltage output with $I_a=28$ ma and $I_a=45$ ma.

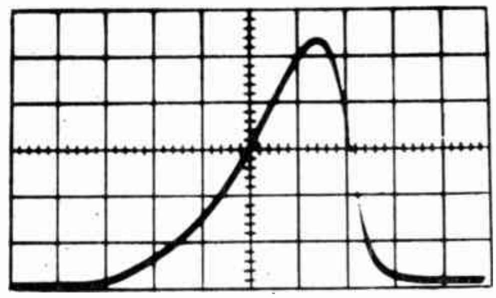
respectively. The results are shown in Fig. 10(b). We next apply a ramp function

$$I(t) = 25t + 11.5 \text{ ma} \quad (t \text{ in } \mu\text{sec})$$

as shown in Fig. 11(a); the output voltage $V(t)$ is given in Fig. 11(b). In Fig. 12 we have traced $V(t)$ from Fig. 11(b) and plotted its value obtained from (33) with $a(t)$ the step response of Fig. 10(b). The difference between the theoretical and experimental results is of the order of measurement accuracy.



(a)



(b)

Fig. 11—Prepulsed ramp response: (a) driving current, (b) observed voltage output.

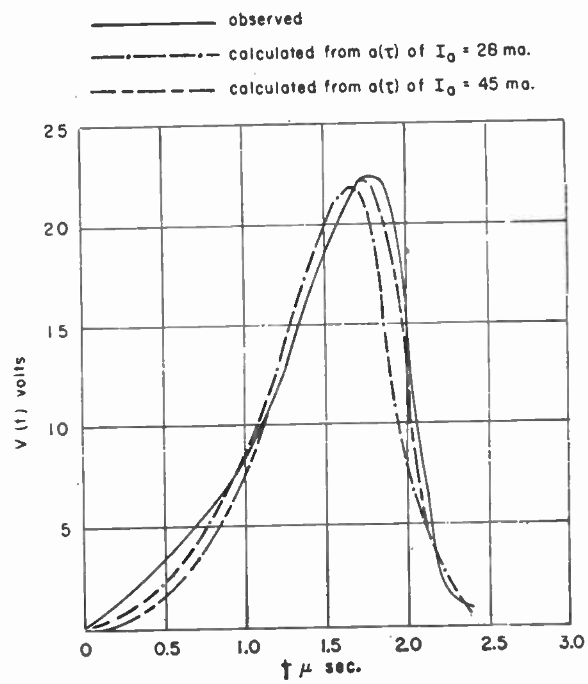


Fig. 12—Calculated and observed voltage output of a thin core driven by a ramp current.

IV. Cores of Intermediate Thickness

The delay in switching cores of intermediate thickness is caused by the induced currents and the finite speed of increase in B [see (25)]. If a constant current is applied, the magnetization will reverse throughout the material but with a speed that will decrease with the distance from the surface, since the field is no longer uniform. The magnetization will be reversed first near the surface, and the reversed zone will widen until it covers the entire cross section of the lamination. Thus we picture the domain growth as in Fig. 13; for t smaller than a certain time t_0, which will be determined, the state of magnetization will be as shown in Fig. 13(a). For $t>t_0$ the magnetization of the material will be reversed in a layer from the surface to the point x,

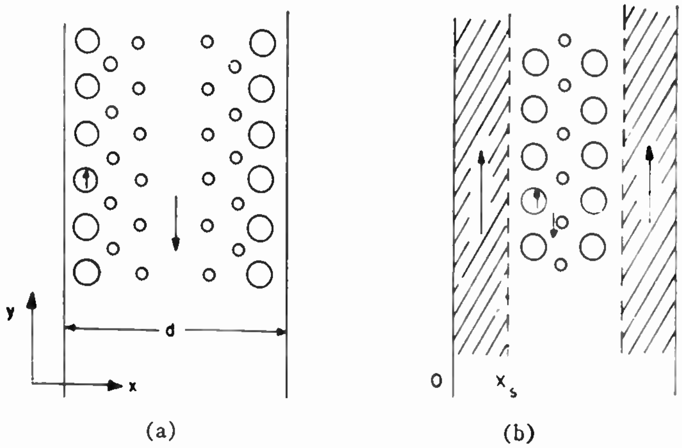


Fig. 13—Reversal of magnetization in a magnetic core of intermediate thickness: (a) for $t<t_0$, (b) for $t>t_0$.

[Fig. 13(b)] but unlike the analysis in Section II, the remaining of the lamination is not in the $-B_s$ state; thus the increase in the flux is not caused by the motion of the plane $x=x_s$, but by a volume growth in the unsaturated portion of the core.

Analysis of Magnetic Reversal

To determine the reversal of magnetization one must solve the field equations (1) and (2) with the boundary conditions (3), (4), and (5). Unlike the discussion in Section II, B is not a known function of H but it satisfies (25) and the boundary condition

$$B(0, x) = -B_s. \tag{37}$$

With an arbitrary $f(B)$ these equations can be solved only numerically; in the following a variational method is used to give an approximate solution. To determine the field distribution inside the material it is assumed that $f(B)$ is constant for $B<B_s$

$$\begin{aligned} f(B) &= C \quad \text{for} \quad -B_s \le B < B_s \\ f(B) &= 0 \quad \text{for} \quad B = B_s \end{aligned} \tag{38}$$

with the value of H thus computed and the actual $f(B)$ the response will be evaluated as in (31).

We define the time t_0 as the maximum time such that for $t<t_0$ the magnetization of the core is nowhere reversed; we then have for $t<t_0$ [see (38)], $f(B)=C$ and from (25)

$$\frac{\partial B}{\partial t} = C(H - H_0). \tag{39}$$

Eqs. (1), (2), and (39) give

$$\frac{\partial^2 H}{\partial x^2} = C(H - H_0) \tag{40}$$

an ordinary differential equation in H, whose solution is given by

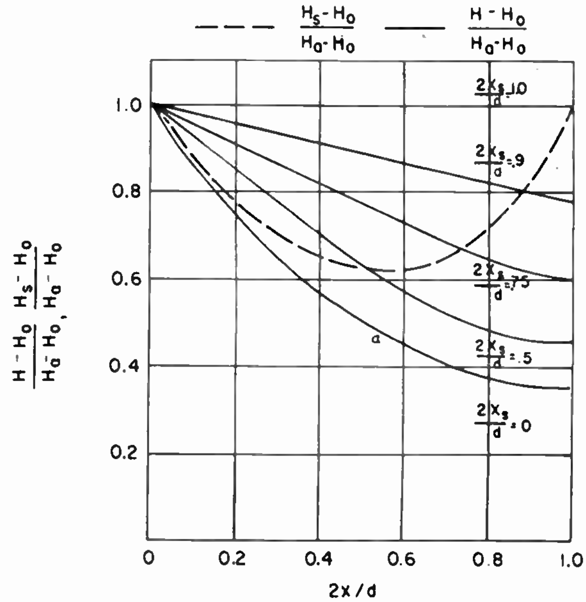


Fig. 14—Magnetic field distribution inside a ½-mil 4-79 Mo-Permalloy lamination.

$$H - H_0 = \frac{H_i(t) - H_0}{\cosh \gamma \frac{d}{2}} \cosh \gamma \left(\frac{d}{2} - x\right) \tag{41}$$

(see curve a in Fig. 14) where

$$\gamma^2 = \sigma C \tag{42}$$

is a constant and $H_i(t)$ is the field at the surface of the material as given by (5). To determine t_0 we integrate (39) [see also (37)] for $x=0$

$$B(0, t) = C \int_0^t [H_i(\tau) - H_0] d\tau - B_s \tag{43}$$

t_0 is such that $B(0, t_0) = B_s$, hence

$$\int_0^{t_0} [H_i(\tau) - H_0] d\tau = \frac{2B_s}{C}, \tag{44}$$

the above gives t_0 if $H_i(t)$ is known.

For $t>t_0$ the magnetization is reversed up to the point $x_s(t)$; with H_s the value of H at the point x_s we have for $x>x_s$

$$H - H_0 = \frac{H_s - H_0}{\cosh \gamma \left(\frac{d}{2} - x_s\right)} \cosh \gamma \left(\frac{d}{2} - x\right) \tag{45}$$

since (40) still holds, and [see (2)]

$$E = \sqrt{\frac{C}{\sigma}} \frac{H_s - H_0}{\cosh \gamma \left(\frac{d}{2} - x_s\right)} \sinh \gamma \left(\frac{d}{2} - x\right). \tag{46}$$

To determine H_s we have as in Section II, (12),

$$E(x_s, t) = E_s(t) = \frac{H_i(t) - H_0}{\sigma x_s} \tag{47}$$

(46) and (47) give

$$H_s - H_0 = \frac{H_i(t) - H_0}{1 + \gamma x_s \tanh \gamma\left(\frac{d}{2} - x_s\right)} \tag{48}$$

inserting (48) into (45) we obtain

$$H - H_0 = \frac{[H_i(t) - H_0] \cosh \gamma\left(\frac{d}{2} - x\right)}{\cosh \gamma\left(\frac{d}{2} - x_s\right) + \gamma x_s \sinh \gamma\left(\frac{d}{2} - x_s\right)}. \tag{49}$$

In Fig. 14, the dotted line, the normalized quantity $(H_s - H_0)/(H_a - H_0)$ is plotted as a function of x_s for a constant applied field $H_i(t) = H_a$; in the same figure the solid lines are the plots of $(H - H_0)/(H_a - H_0)$ for the same driver and various values of x_s. To determine H for a given t we must first find x_s as a function of time. We first determine the initial condition for B at $t = t_0$ from (39) and (41); it is given by

$$B(t_0, x) = C \frac{\cosh \gamma\left(\frac{d}{2} - x\right)}{\cosh \gamma \frac{d}{2}} \int_0^{t_0} [H_i(\tau) - H_0] d\tau - B_s \tag{50}$$

we then integrate (39) and find the value t for which $B(t, x) = B_s$. Thus we obtain, omitting the details of the derivation,

$$C \int_{t_0}^{t} [H_i(\tau) - H_0]\, d\tau = 2B_s\left[\frac{1}{2} \gamma^2 x_s^2 + \gamma x_s \tanh \gamma\left(\frac{d}{2} - x_s\right)\right]. \tag{51}$$

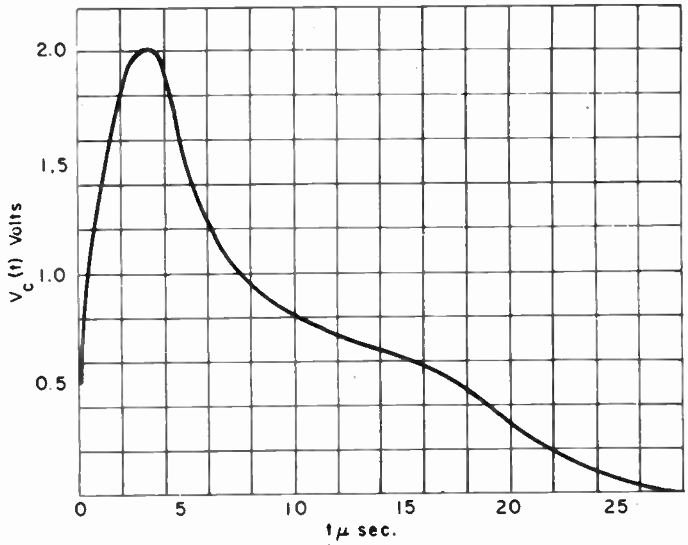
Eq. (51) gives x_s as a function of time. The switching time is the time for which $x_s = d/2$. It can readily be found from (51) if $H_i(t)$ is known. For a constant applied field $H_i = H_a$ (51) gives

$$t_s = t_0 + \frac{\sigma B_s}{4(H_a - H_0)} d^2, \tag{52}$$

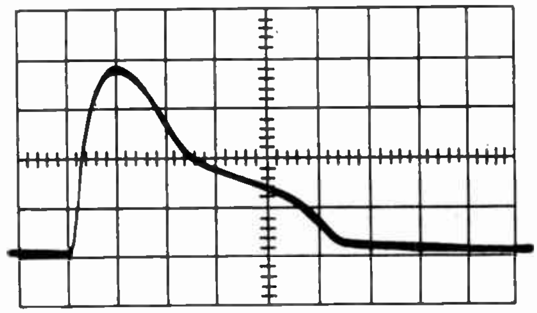
thus the relation between t_0 and d^2 is a straight line with t_0 the time intercept; (52) has been experimentally verified.[4]

Knowing the field distribution as a function of time we can find the voltage output using the results of Section III; the rate of change of flux through a section of the core of thickness dx and height h is given by [see (33)]

$$\frac{dx}{d} \frac{H(x, t) - H_0}{H_a - H_0} a\left[\int_0^t \frac{H(x, \tau) - H_0}{H_a - H_0} d\tau\right],$$



(a)



(b)

Fig. 15—Voltage output of a 1-mil 4-79 Mo-Permalloy core driven by a step current: (a) calculated, (b) observed.

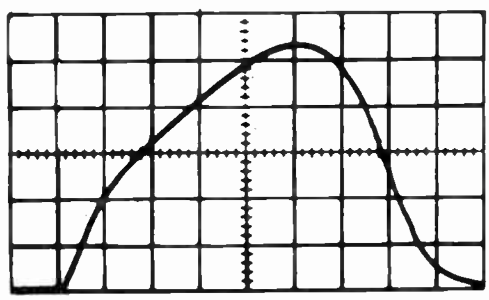


Fig. 16—Voltage output of a prepulsed thin core driven by a step current.

hence the voltage output $V_c(t)$ across a one-turn winding is given by

$$V_c(t) = \frac{2}{d} \int_0^{d/2} \frac{H(x, t) - H_0}{H_a - H_0} a\left[\int_0^t \frac{H(x, \tau) - H_0}{H_a - H_0} d\tau\right] dx. \tag{53}$$

In Fig. 15(a) we have plotted $V_c(t)$ as given by (53) with $H_i(t) = $ constant and $d = 1$ mil; the details of the computation are omitted. In Fig. 15(b) the experimentally obtained output is shown.

The Flux Charge per Meter Curve

The Φ-Q relationship can be established for an arbitrary $f(B)$ without the solution of the resulting field problem. If we assume that the step-response $a(t)$ of the corresponding thin core tends asymptotically to zero, then the field $B(0, t)$ at the surface of the core and the total flux $\Phi(t)$ through the core will increase with t; $B(0, t)$ can, therefore, be considered as a function $w(\Phi)$ of the flux Φ. From (25) we have

$$\frac{dB(0, t)}{dt} = f(B)[H_i(t) - H_0]$$

therefore

$$\int_{-B_r}^{B} \frac{dB}{f(B)} \simeq Q(t) \quad \text{if} \quad Q(t) \gg H_0 t. \qquad (54)$$

The left-hand side in (54) is a function $F(B)$ of B, hence

$$Q(t) = F(B) = F(w(\Phi)) = W(\Phi).$$

The Φ-Q relationship is thus established; it is valid also if $a(t)$ becomes zero for a finite t, as one can readily see by a limit argument.

For the theoretical determination of $W(\Phi)$, the function $w(\Phi)$ should first be determined from the solution of the field problem; however, if the step-response $V_c(t)$ is known then $W(\Phi)$ can readily be obtained; indeed from

$$\Phi(t) = \int_0^t V_c(\tau)d\tau,$$

$$Q(t) = \int_0^t (H_a - H_0)d\tau = (H_a - H_0)t$$

we obtain Q as a function $W(\Phi)$ of Φ by eliminating t.

V. Appendix

Theory of Domain Reversal[9–11]

To establish theoretically the validity of the assumption given by (25), the main features of domain reversal should be considered. The domains are caused by interatomic forces that tend to align with each other the magnetic moments of a ferromagnetic material; with no other forces present the material will thus be uniformly magnetized. However, because of the crystal symmetry, the presence of tensions, and the existence of local fields, certain directions are energetically preferred and the domains are oriented in a direction that minimizes the crystalline, magnetostrictive, and field energy. The field is caused by the applied and induced currents and the free poles resulting from the variation of the magnetic moment M. Since in general there is more than one direction of minimum energy, the domain orientation is not unique. If, for example, the crystalline energy prevails and the crystal is cubic, its six axes will form possible domain directions. When the material is in a nonmagnetized state the domains are oriented at random in each of these directions and the average magnetization is zero. The application of a sufficiently high field H will cause all the domains to turn in its direction and if H is removed they will turn in the direction nearest to the applied field.[12] Thus, the residual magnetization will result since the domains are no longer oriented at random in all possible directions of minimum energy. If in particular one of these directions is parallel to the applied field for all crystals, then there will be no rotation after the field is removed and the material will remain saturated; its "squareness" will equal one.

Suppose that a square-loop core is magnetized in the $-z$ direction and its magnetization is reversed through the application of a field in the $+z$ direction. The reversal will take place either through a direct rotation of domains or through a motion of the wall that separates domains of opposite direction. However, the minimum field necessary to rotate the domains is higher than the field required to switch the core; we must then assume that the switching takes place mainly through wall motion for which smaller fields are required.[13] For the reversal to take place through wall motion there must exist inside the material some regions, the nucleations, that are either oriented in the $+z$ direction before the field is applied, or require a small field for their formation. This is possible because of the local fields caused by the discontinuity of the magnetic moment M at the crystal surfaces and at the surfaces of foreign substances that are present in the material. The resulting free poles change the energy balance so that the external field H_n necessary to rotate the domains is much smaller than the field required without free poles.

The exact determination of the shape of the domain walls and their motion does not seem possible unless cores of special geometrics and treatment are considered.[14] For an analysis of the terminal properties of the core the following simplifying assumptions are made.

The magnetization reverses through wall motion; thus we ignore the effect of the formation of nucleations. For fields that are much greater than the starting field H_0 this approximation is satisfactory; however, to minimize the nucleation effect in the experimental verification of the developed analysis, the core is prepulsed with a

[9] R. Becker and W. Doring, "Ferromagnetism," Springer, Berlin, Germany; 1939.

[10] R. Bozorth, "Ferromagnetism," D. Van Nostrand Co., Inc., New York, N. Y.; 1951.

[11] J. L. Snoek, "New Developments in Ferromagnetic Materials," Elsevier Publishing Co., Inc., New York, N. Y., and Amsterdam, Netherlands; 1947.

[12] A. Papoulis, "The nondestructive read-out of magnetic cores," Proc. IRE, vol. 42, pp. 1283–1288; August, 1954.

[13] J. B. Goodenough, "A theory of domain creation, coercive force in polycrystalline ferromagnetics," *Phys. Rev.*, vol. 95, pp. 917–932; August, 1954.

[14] H. Ekstein, "Theory of remagnetization of thin tapes," *J. Appl. Phys.*, vol. 26, pp. 1342–1343; June, 1955.

field somewhat smaller than H_0 before applying the switching pulse. In Fig. 16 the step response of a pre-pulsed core is shown. One notices that the first maximum is not present and it can be attributed to the formation of the nucleating centers.[4]

The domains grow uniformly throughout the cross section of the material; thus the special conditions near the surface of the core are ignored.

We shall further assume that when the magnetization reaches the state B (the term $\mu_0 H$ will be omitted in $B=\mu_0 H+M$), the available wall area depends only on B, independently of the way in which this state is reached. With S the total wall area per unit volume we thus have

$$S = \phi(B) \tag{55}$$

where $\phi(B)$ is a function of state. With v the velocity of wall motion of the 180° walls the rate of increase of magnetization is given by

$$\frac{dB}{dt} = 2B_s vS = 2B_s v\phi(B) \tag{56}$$

since in the unit of time the magnetization changes from $-B_s$ to $+B_s$ in the volume vS.

Since the number and shape of the growing domains is not known, $\phi(B)$ cannot be directly evaluated; in the following it is seen that it can be determined through terminal measurements. However, certain properties of $\phi(B)$ can be stated *a priori;* indeed for small values of B, $\phi(B)$ increases since the area of the domain walls increases with their size. This increase continues until the walls start colliding with each other; $\phi(B)$ thus reaches a maximum and then decreases to zero when the material is completely switched. $\phi(B)$ can be given explicitly for domains of simple shapes. For needle-like domains $\phi(B)$ is constant; for spherical domains of radius r, B is proportional to r^3 and the wall area to r^2, hence

$$\phi(B) = KB^{2/3}.$$

For cylindrical domains of radius r, B is proportional to r^2 and S proportional to r, hence

$$\phi(B) = K\sqrt{B}. \tag{57}$$

It is seen presently that the experimentally computed $\phi(B)$ agrees in its increasing portion with (57), thus suggesting a cylindrical growth.[15]

To determine the rate of domain growth the velocity v of the wall motion must be known as function of the applied field H_i. It is assumed that[16,17]

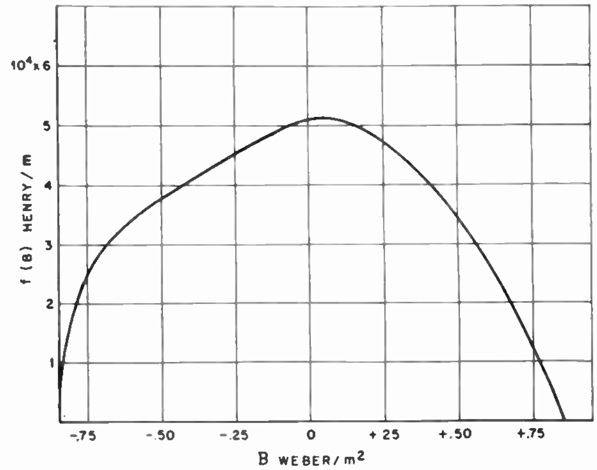


Fig. 17—Plot of $f(B)$ for a $\frac{1}{8}$-mil 4-79 Mo-Permalloy lamination.

$$v = \alpha(H_i - H_0) \tag{58}$$

where α and H_0 are constant of the material. The above is approximate, actually the wall satisfies a second-order differential equation[4] but the acceleration term is negligible for the applications under consideration in which the times involved are greater than 1 μsec.

From (56) and (58) we obtain the desired equation

$$\frac{dB}{dt} = 2B_s\alpha\phi(B)(H_i - H_0) = f(B)(H_i - H_0) \tag{25}$$

where

$$f(B) = 2B_s\alpha\phi(B). \tag{59}$$

The function $f(B)$ can be determined from the constant-current step-response $a(t)$; indeed from

$$a(t) = hd\frac{dB}{dt}$$

and (25) we obtain

$$B(t) + B_s = \frac{1}{hd}\int_0^t a(\tau)d\tau \tag{60}$$

$$f(B(t)) = \frac{a(t)}{hd(H_a - H_0)} \tag{61}$$

eliminating t between (60) and (61) we obtain $f(B)$. With $a(t)$ as in Fig. 16, $f(B)$ is computed as above and the result is plotted in Fig. 17. The initial rise in Fig. 16 is due to the finite rise time of the driver; the remaining part of the increasing portion of $a(t)$ is nearly a straight line. This indicates that the domains grow as cylinders and their area is given by (57) as one can readily see from

$$a(t) = hd\frac{dB}{dt} = ct + c_1$$

and (61).

[15] C. Kittel, "Physical theory of ferromagnetic domains," *Rev. Mod. Phys.*, vol. 21, pp. 541–583; October, 1949.

[16] K. J. Sixtus and L. Tonks, "Propagation of large Barkhausen discontinuities," *Phys. Rev.*, vol. 37, pp. 930–958; September, 1931.

[17] H. J. Williams, H. Shockley, and C. Kittel, "Studies of the propagation velocity of a ferromagnetic domain boundary," *Phys. Rev.*, vol. 80, pp. 1090–1094; December, 1950.

Some General Properties of Nonlinear Elements. II. Small Signal Theory*

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Summary—The simplest types of nonlinear capacitor modulators, demodulators, and negative conductance amplifiers, in which components at only two signal frequencies are present, are studied by means of the well-known small-signal analysis. The results of this analysis of course agree with the general energy relations of Part I,[1] but in addition give the gain, bandwidth, terminal admittances, and sensitivity (to changes in the terminal admittances or in the local oscillator drive) of these devices, and show the way in which these quantities depend on the amount of nonlinearity. In general, the bandwidth of all of these devices approaches zero as the nonlinearity approaches zero.

Three cases are considered; f_1 is the local oscillator frequency.

1) Noninverting modulator and demodulator—signal frequencies f_l and $f_+ = f_1 + f_l$. This device is stable and yields maximum gain with matched source and load. Under matched conditions the gain is equal to the ratio of output to input frequency; for widely separated signal frequencies the modulator has substantial gain, the demodulator an equal loss. Only a relatively small amount of nonlinearity is required to attain a bandwidth equal to the low-signal frequency. Since source and load are matched the sensitivity is zero.

2) Inverting modulator and demodulator—signal frequencies f_l and $f_- = f_1 - f_l$. This device is potentially unstable; its input conductance is negative and match is impossible. The modulator gain again exceeds the demodulator gain, the ratio of the two being the same as in the noninverting case; but now both gains may be made as large as desired at the expense of narrow bandwidth and high sensitivity. For gains equal to the ratio of output to input frequency (as in the matched noninverting device) the bandwidth is slightly smaller than in the noninverting case; again a relatively small amount of nonlinearity yields a bandwidth equal to the low-signal frequency. For a larger amount of nonlinearity, substantially greater gains can be obtained with moderate bandwidth and sensitivity. However, a demodulator with high gain must have a very narrow bandwidth and a high sensitivity.

3) Inverting negative conductance amplifier. The negative input conductance of the inverting device discussed above may be used to provide amplification at a single frequency. Substantial gain at moderate bandwidth and sensitivity may be obtained at either signal frequency; as with the inverting modulator and demodulator, high gain implies narrow bandwidth and high sensitivity. A large nonlinearity is desirable to achieve the maximum bandwidth.

Finally, the general energy relations of Part I for nonlinear reactors are shown to apply to lossless linear variable reactors for arbitrary signal levels if the powers at the different frequencies in these equations include mechanical as well as electrical power.

Introduction

In Part I[1] two independent equations relating the powers at the different frequencies in nonlinear capacitors and inductors were derived. These general energy relations are independent of the nonlinear characteristic, the external circuit, and the signal levels at the various frequencies. These relations have been derived in a simpler way for quantum mechanical systems such as Masers.[2] Alternate classical derivations for nonlinear capacitors and inductors have been given.[3,4]

Although these results give useful information about the general behavior of various nonlinear reactor devices, they are, of course, not a substitute for a detailed analysis of a particular device. While in general a complete analysis will be difficult to carry out, one important case in which it may be accomplished readily is the well-known small-signal case,[5–9] in which the levels of the signal components applied to the nonlinear element are much smaller than the level of the local oscillator.

The present paper uses the standard small-signal analysis to discuss the behavior of the simplest nonlinear reactor devices, used to illustrate the general energy relations in Part I, in which only one of the principal sidebands of the applied signal about the local oscillator frequency is allowed to carry a significant amount of power. The results in the small-signal case of course agree with the general energy relations, but in addition yield expressions for the gain, bandwidth, impedances, and sensitivity (to changes in the terminal admittances or in the local oscillator drive) for the various modulators, demodulators, and negative conductance amplifiers. These results may be of interest in connection with the various recent microwave devices of this type using *p-n* junctions and ferrites.[10–12] As in Part I, the analysis will be carried out for the nonlinear capacitor only, since the corresponding analysis for the nonlinear inductor is very similar.

* Original manuscript received by the IRE, December 20, 1957; revised manuscript received February 6, 1958.

† Bell Telephone Labs., Inc., Holmdel, N. J.

[1] J. M. Manley and H. E. Rowe, "Some general properties of nonlinear elements—Part I. General energy relations," Proc. IRE, vol. 44, pp. 904–913; July, 1956.

[2] M. T. Weiss, "Quantum derivation of energy relations analogous to those for nonlinear reactances," Proc. IRE, vol. 45, pp. 1012–1013; July, 1957.

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[7] H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers," McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.

[8] C. F. Edwards, "Frequency conversion by means of a nonlinear admittance," *Bell Sys. Tech. J.*, vol. 35, pp. 1403–1416; November, 1956.

[9] S. Duinker, "General properties of frequency-converting networks," Dissertation, Technical University of Delft, Netherlands; June, 1957.

[10] A. Uhlir, "Two-terminal *p-n* junction devices for frequency conversion and computation," Proc. IRE, vol. 44, pp. 1183–1191; September, 1956.

[11] H. Suhl, "Proposal for a ferromagnetic amplifier in the microwave range," *Phys. Rev.*, vol. 106, pp. 384–385; April 15, 1957.

[12] M. T. Weiss, "A solid-state microwave amplifier and oscillator using ferrites," *Phys. Rev.*, vol. 107, p. 317; July 1, 1957.

For convenience the frequencies involved in the two types of devices to be considered are shown in Fig. 1 (taken from Fig. 2, Part I), together with the corresponding general energy relations and modulator and demodulator power gains, given in Part I. The notation and terminology will remain the same. f_1 is the local oscillator frequency, f_l the lowest signal frequency (normally small compared to the local oscillator frequency for modulators and demodulators), $f_\pm = f_1 \pm f_l$. For modulators the signal input is at f_l, the signal output at either f_+ or f_-; for demodulators the signal input is at either f_+ or f_-, the signal output at f_l. The W's represent power flowing into the nonlinear element at the corresponding frequencies indicated by the subscripts. G_p represents power gain, defined as the ratio of power delivered to a specified load impedance to the power absorbed by the input of a transducer,[13] the subscripts indicate the direction of travel of the signal. The names inverting and noninverting were chosen because these devices respectively do and do not invert the signal spectrum when used as modulators or demodulators. A detailed discussion of the operation of these devices is given in Part I.

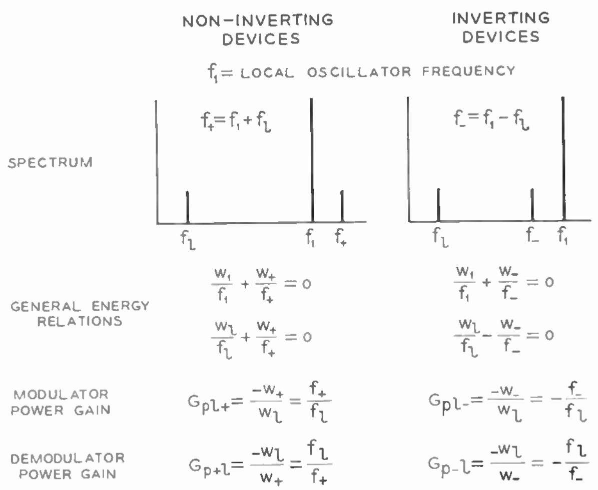

Fig. 1—Signal and local oscillator frequencies, general energy relations, and power gains for nonlinear reactor devices.

Finally, the general energy relations of Part I for nonlinear reactors are shown to hold true for any lossless linear variable reactor if, assuming the reactor to be varied mechanically, both electrical and mechanical power at every frequency are included in the analysis.

Small-Signal Analysis[6-9]

In this section the well-known small-signal analysis is summarized briefly for a nonlinear capacitor. The charge q will be some function of the voltage v,

$$q = f(v). \tag{1}$$

Let q_1 and v_1 be the carrier components of charge and voltage, consisting of fundamental components and harmonics of the local oscillator frequency f_1. Then in the absence of signal

$$q_1 = f(v_1). \tag{2}$$

Let the signal components of charge and voltage be given by δq and δv, small compared to the carrier components q_1 and v_1. Then,

$$\delta q = f'(v_1) \cdot \delta v \qquad f'(v) = \frac{df(v)}{dv}. \tag{3}$$

Now $f'(v_1)$, which may be thought of as an equivalent linear time-varying capacitance, will be periodic, of fundamental frequency f_1. Therefore, it may be written as a Fourier series,

$$f'(v_1) = \sum_{n=-\infty}^{\infty} C_n e^{jnx} \qquad x = 2\pi f_1 t \tag{4}$$

$$C_n = \frac{1}{2\pi}\int_0^{2\pi} f'(v_1)e^{-jnx}dx$$

$$C_n = C_{-n}^*.\text{[14]} \tag{5}$$

Thus, as far as the signal components δq and δv are concerned, the nonlinear capacitor may be replaced by an equivalent linear variable capacitor whose variation with time is determined by the nonlinear characteristic and by the local oscillator waveform. The origin of time has been arbitrary up to now; we can always choose it in such a way that

$$C_1 = C_{-1} = \text{positive real}, \tag{6}$$

and in the following we will assume that this has been done.

For our special case we now assume that the nonlinear capacitor is terminated in an ideal filter such that signal voltages at only the three frequencies f_l, $f_+ = f_1 + f_l$, and $f_- = f_1 - f_l$ are allowed to exist across the nonlinear capacitor. Then,

$$\delta v = V_+^* e^{-j(x+y)} + V_-^* e^{-j(x-y)} + V_l^* e^{-jy} + V_l e^{jy} + V_- e^{j(x-y)} + V_+ e^{j(x+y)} \tag{7}$$

$$\delta q = Q_+^* e^{-j(x+y)} + Q_-^* e^{-j(x-y)} + Q_l^* e^{-jy} + Q_l e^{jy} + Q_- e^{j(x-y)} + Q_+ e^{j(x+y)} + \cdots \tag{8}$$

$$x = 2\pi f_1 t \qquad y = 2\pi f_l t. \tag{9}$$

As indicated, δq will contain additional components at all of the frequencies $mf_1 \pm f_l$, $m = 2, 3, \cdots$. However, since the ideal filter suppresses the voltages at these frequencies, no power can flow except at the principal signal frequencies f_l, $f_1 \pm f_l$, and so these other components play no part in the present analysis. Substituting (4), (7), and (8) into (3), we find that the final results for the components of interest can be written in the usual matrix form

[13] ASA C42, "American Standard Definitions of Electrical Terms," August, 1953.

[14] The * indicates the complex conjugate.

$$\begin{bmatrix} Q_-^* \\ Q_l \\ Q_+ \end{bmatrix} = \begin{bmatrix} C_0 & C_{-1} & C_{-2} \\ C_1 & C_0 & C_{-1} \\ C_2 & C_1 & C_0 \end{bmatrix} \begin{bmatrix} V_-^* \\ V_l \\ V_+ \end{bmatrix}. \quad (10)$$

Since $I = j2\pi f \cdot Q$ and $I^* = -j2\pi f \cdot Q^*$, we have from (10) the well-known relation

$$\begin{bmatrix} I_-^* \\ I_l \\ I_+ \end{bmatrix} = \begin{bmatrix} -j2\pi f_- C_0 & -j2\pi f_- C_{-1} & -j2\pi f_- C_{-2} \\ j2\pi f_l C_1 & j2\pi f_l C_0 & j2\pi f_l C_{-1} \\ j2\pi f_+ C_2 & j2\pi f_+ C_1 & j2\pi f_+ C_0 \end{bmatrix} \begin{bmatrix} V_-^* \\ V_l \\ V_+ \end{bmatrix} \cdot (11)$$

Eq. (11) describes the small-signal behavior of the nonlinear capacitor when all three signal frequencies are present. For the noninverting devices of Fig. 1, we assume the voltage at the frequency $f_- = f_1 - f_l$ is suppressed by the ideal filter. Setting $V_-^* = 0$ in (11), and noting (6),

Noninverting case

$$\begin{bmatrix} I_l \\ I_+ \end{bmatrix} = \begin{bmatrix} j2\pi f_l C_0 & j2\pi f_l C_1 \\ j2\pi f_+ C_1 & j2\pi f_+ C_0 \end{bmatrix} \begin{bmatrix} V_l \\ V_+ \end{bmatrix}. \quad (12)$$

In the alternate case the voltage at the frequency $f_+ = f_1 + f_l$ is suppressed, and setting $V_+ = 0$ in (11),

Inverting case

$$\begin{bmatrix} I_l \\ I_-^* \end{bmatrix} = \begin{bmatrix} j2\pi f_l C_0 & j2\pi f_l C_1 \\ -j2\pi f_- C_1 & -j2\pi f_- C_0 \end{bmatrix} \begin{bmatrix} V_l \\ V_-^* \end{bmatrix}. \quad (13)$$

Since C_0 and C_1 are real, the admittance matrices of (12) and (13) are pure imaginary.

Gain and Terminal Admittances for a Four-Pole With a Pure Imaginary Admittance Matrix

Consider a linear four-pole characterized by a pure imaginary admittance matrix, such as those of (12) and (13).

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} jB_{11} & jB_{12} \\ jB_{21} & jB_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix}. \quad (14)$$

The nonlinear capacitor may be treated as such a four-pole; the two signal frequencies are assumed to be brought out at two different ports by the ideal filter associated with the nonlinear element. The powers flowing into the network at the two ports are

$$W_1 = \tfrac{1}{2}\,\mathrm{Re}\, V_1 I_1^*, \qquad W_2 = \tfrac{1}{2}\,\mathrm{Re}\, V_2 I_2^*. \quad (15)$$

Then from (14) we have

$$\frac{W_1}{B_{12}} + \frac{W_2}{B_{21}} = \frac{1}{2}\,\mathrm{Re}\begin{bmatrix} \frac{V_1^*}{B_{12}} & \frac{V_2^*}{B_{21}} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix}$$

$$= \frac{1}{2}\,\mathrm{Re}\begin{bmatrix} \frac{V_1^*}{B_{12}} & \frac{V_2^*}{B_{21}} \end{bmatrix} \begin{bmatrix} jB_{11} & jB_{12} \\ jB_{21} & jB_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = 0.$$

Therefore,

$$\frac{W_1}{B_{12}} + \frac{W_2}{B_{21}} = 0$$

$$G_{p12} = \frac{-W_2}{W_1} = \frac{B_{21}}{B_{12}} \qquad G_{p21} = \frac{-W_1}{W_2} = \frac{B_{12}}{B_{21}}. \quad (16)$$

From (12) and (13), therefore, we have for the nonlinear capacitor:

	Noninverting case	*Inverting case*
Energy relation	$\frac{W_l}{f_l} + \frac{W_+}{f_+} = 0$	$\frac{W_l}{f_l} - \frac{W_-}{f_-} = 0$
Modulator power gain	$G_{pl+} = \frac{-W_+}{W_l} = \frac{f_+}{f_l}$	$G_{pl-} = \frac{-W_-}{W_l} = -\frac{f_-}{f_l}$
Demodulator power gain	$G_{p+l} = \frac{-W_l}{W_+} = \frac{f_l}{f_+}$	$G_{p-l} = \frac{-W_l}{W_-} = -\frac{f_l}{f_-}$. (17)

Thus, the small-signal results agree with the general energy relations of Part I, shown in Fig. 1.

Next consider the four-pole terminated in admittances Y_1 and Y_2 and driven by current sources, as shown in Fig. 2. g_1 and g_2 represent the internal generator conductance and the load conductance, and B_1 and B_2 represent the terminal susceptances added for matching purposes. The input admittances are given by

$$Y_{1\,\mathrm{in}} = jB_{11} + \frac{B_{12}B_{21}}{g_2 + j(B_2 + B_{22})}$$

$$Y_{2\,\mathrm{in}} = jB_{22} + \frac{B_{12}B_{21}}{g_1 + j(B_1 + B_{11})}. \quad (18)$$

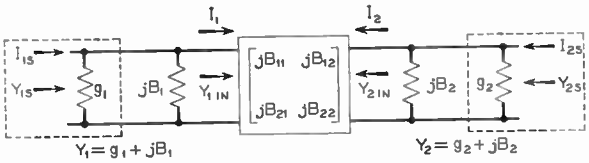

Fig. 2—Linear four-pole driven by current sources.

The admittances seen by the current sources are

$$Y_{1s} = Y_1 + Y_{1\,\mathrm{in}}$$

$$= g_1 + j(B_1 + B_{11}) + \frac{B_{12}B_{21}}{g_2 + j(B_2 + B_{22})}$$

$$Y_{2s} = Y_2 + Y_{2\,\mathrm{in}}$$

$$= g_2 + j(B_2 + B_{22}) + \frac{B_{12}B_{21}}{g_1 + j(B_1 + B_{11})}. \quad (19)$$

Next we determine the transducer gains G_t,[13] defined as the ratio of the power delivered to the load to the available power of the source, for various types of devices. We consider two cases separately.

Case 1

$$\underline{B_{12}B_{21} > 0.}$$

These conditions correspond to the noninverting nonlinear capacitor devices (12). From (16),

$$G_p > 0. \tag{20}$$

For all terminal conditions (18) show that

$$\text{Re } Y_{1\text{ in}} > 0 \qquad \text{Re } Y_{2\text{ in}} > 0, \tag{21}$$

so that these devices are always stable.

We next determine the transducer gains for the signal input and output at opposite ports (corresponding to the transducer gains for the noninverting modulator and demodulator). From (16) the power delivered to the load equals the power gain G_p times the power entering the input port of the four-pole. The latter is found in the usual way in terms of the admittances (18) and (19). Thus,

$$G_{t12} = G_{p12} \frac{4 \text{ Re } \dfrac{Y_{1\text{ in}}}{g_1}}{\left|\dfrac{Y_{1s}}{g_1}\right|^2}$$

$$G_{t21} = G_{p21} \frac{4 \text{ Re } \dfrac{Y_{2\text{ in}}}{g_2}}{\left|\dfrac{Y_{2s}}{g_2}\right|^2}, \tag{22}$$

where the corresponding power gains G_p are given in (16), or for the nonlinear capacitor in (17). From (20) and (21),

$$G_t > 0. \tag{23}$$

The maximum value of G_t, defined as the maximum available gain, is attained for conjugate match at both ports; the terminal admittances must then satisfy the relation

$$[g_1 + j(B_1 + B_{11})] \cdot [g_2 - j(B_2 + B_{22})] = B_{12}B_{21}$$

or

$$\angle [g_1 + j(B_1 + B_{11})] = \angle [g_2 + j(B_2 + B_{22})]$$

$$| g_1 + j(B_1 + B_{11}) | \cdot | g_2 + j(B_2 + B_{22}) | = B_{12}B_{21}. \tag{24}$$

In this case (22) becomes $G_t = G_p$, and thus the maximum available gain is equal to the power gain.

Eq. (24) indicates that a wide class of possible terminal admittances will yield a conjugate match at input and output ports. Indeed, for any output admittance whatever, a corresponding input admittance can be found that will match both input and output, and vice versa. The reason for this is easily seen. A four-pole of the special type given in (14) (all terms pure imaginary) must satisfy the general energy relation of (16) or (17). Since the power gains are independent of the terminal admittances, the transducer gain is maximized and made equal to the power gain by adjusting the generator (load) admittance for match, for any load (generator) admittance. The choice from the infinitely many possible pairs of terminal admittances must be made on grounds other than gain, such as maximizing the bandwidth of the device.

Case 2

$$\underline{B_{12}B_{21} < 0.}$$

These conditions correspond to the inverting nonlinear capacitor devices (13). From (16),

$$G_p < 0. \tag{25}$$

Eq. (18) shows that for all terminal conditions

$$\text{Re } Y_{1\text{ in}} < 0 \qquad \text{Re } Y_{2\text{ in}} < 0, \tag{26}$$

so these devices are potentially unstable. Under certain terminal conditions they will oscillate, so that a small-signal analysis is invalid. If the device is stable, the transducer gains G_{t12} and G_{t21} (corresponding to the inverting modulator and demodulator transducer gains) are again given by (22).

Since from (25) and (26) both G_p and Re Y_{in} are negative, (22) shows that

$$G_t > 0 \tag{27}$$

as before. In this case, since the input conductance will always be negative (26), match is never possible, and the transducer gain can be made as large as desired by operating close enough to instability, making the factor $|Y_s/g|^2$ in the denominator of (22) small enough.

The operation discussed so far has assumed a generator connected to one port of the four-pole and a load connected to the other. In the nonlinear capacitor devices the separate ports, of course, correspond to the different signal frequencies; the gains correspond to operation as a modulator or demodulator. However, in the present case, ($B_{12}B_{21} < 0$, corresponding to the inverting case), another type of operation is of interest in which gain at a single port (or signal frequency) is obtained by utilizing the negative input conductance of the device.

Assume that a circulator[15] is connected between the generator and the four-pole (or nonlinear capacitor) as shown in Fig. 3, so that the incident and the reflected waves may be separated easily. W_i represents the power in the incident wave, equal to the available power of the source; W_r represents the power in the reflected wave, transmitted to the load by the circulator. W_1 represents the power entering port 1 and W_2 the power entering port 2, as before (both will be negative in the present

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

case). The negative conductance transducer gain at port 1, considering W_r as the useful output, is given by

$$G_{t11} = \frac{W_r}{W_i}.$$

Since the input conductance is negative, the reflection coefficient is greater than unity, and power gain is obtained.

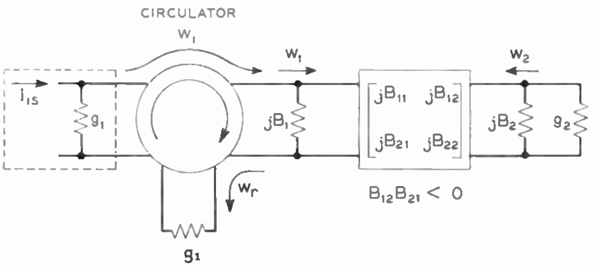

Fig. 3—Linear four-pole operated as a negative conductance amplifier.

From previous discussion the power and transducer gains considering $-W_2$ as the useful output are defined by

$$G_{p12} = \frac{-W_2}{W_1} \qquad G_{t12} = \frac{-W_2}{W_i},$$

given in (16) and (22). Conservation of power requires that

$$W_i = W_r + W_1.$$

Then,

$$G_{t11} = 1 - \frac{W_1}{W_i} = 1 - \frac{G_{t12}}{G_{p12}} \tag{28}$$

with a similar expression for G_{t22}, the corresponding negative conductance gain at port 2. From (22) we finally obtain

$$G_{t11} = 1 - \frac{4 \operatorname{Re} \dfrac{Y_{1\,\mathrm{in}}}{g_1}}{\left|\dfrac{Y_{1s}}{g_1}\right|^2}$$

$$G_{t22} = 1 - \frac{4 \operatorname{Re} \dfrac{Y_{2\,\mathrm{in}}}{g_2}}{\left|\dfrac{Y_{2s}}{g_2}\right|^2}. \tag{29}$$

Since Re $Y_{\mathrm{in}} < 0$ (26), (29) shows that

$$G_t > 1. \tag{30}$$

By making $|Y_s/g|^2$, in the denominator of the second term, sufficiently small the gain can be made as large as desired.

Eq. (29) applies equally well to the first case discussed above ($B_{12}B_{21} > 0$, corresponding to the noninverting case). Here, since Re $Y_{\mathrm{in}} > 0$, (29) states that $0 < G_t < 1$. Of course we do not expect to obtain gain in this way from a device with a positive input conductance.

Nonlinear Capacitor Modulators and Demodulators

Noninverting Case

Consider a nonlinear capacitor whose terminal admittances consist of a parallel combination of conductance and inductive susceptance, as shown in Fig. 4. Here the two signal frequencies f_l and f_+ are assumed to be brought out at separate terminals by the ideal filter associated with the nonlinear element. The g's represent the internal generator conductance and the load conductance.

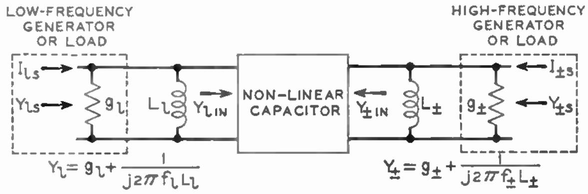

Fig. 4—Nonlinear capacitor with parallel resonant terminal networks.

We now select the terminal admittances to yield a symmetric single-peaked response curve of maximum gain and bandwidth. The terminal admittances must be matched at midband (24); to obtain a symmetric characteristic they are made parallel resonant at midband with their associated self-susceptances of the nonlinear capacitor matrix. Thus,

$$\angle[g_1 + j(B_1 + B_{11})] = \angle[g_2 + j(B_2 + B_{22})] = 0$$

$$B_1 = -B_{11} \qquad B_2 = -B_{22}$$

$$g_1g_2 = B_{12}B_{21}. \tag{31}$$

Denoting the midband frequencies by the subscript m,

$$f_{\substack{l\\+}m} = \text{midband frequency},$$

$$f_{\substack{l\\+}} = f_{\substack{l\\+}m} + \delta f,$$

$$\delta f = 0 \text{ at midband}. \tag{32}$$

The terminal inductances L_l and L_+ and the terminal conductances g_l and g_+ in Fig. 4 are given by

$$L_l = \frac{1}{4\pi^2 f_{lm}^2 C_0} \qquad L_+ = \frac{1}{4\pi^2 f_{+m}^2 C_0};$$

$$g_l g_+ = 4\pi^2 C_1^2 f_{lm} f_{+m}. \tag{33}$$

We may then write $Y_{\substack{l\\+}s} = Y_{\substack{l\\+}} + Y_{\substack{l\\+}\mathrm{in}}$, the admittances seen by the current sources in Fig. 4, in the following convenient form.

$$\frac{Y_{ls}}{g_l} = 1 + jQ_lF_l + \frac{\left(1+\frac{\delta f}{f_{lm}}\right)\left(1+\frac{\delta f}{f_{+m}}\right)}{1+jQ_+F_+}$$

$$\frac{Y_{+s}}{g_+} = 1 + jQ_+F_+ + \frac{\left(1+\frac{\delta f}{f_{lm}}\right)\left(1+\frac{\delta f}{f_{+m}}\right)}{1+jQ_lf_l}, \quad (34)$$

where

$$Q_l = \frac{2\pi f_{lm}C_0}{g_l} \qquad Q_+ = \frac{2\pi f_{+m}C_0}{g_+}$$

$$F_l = \frac{f_l}{f_{lm}} - \frac{f_{lm}}{f_l} = \left(1+\frac{\delta f}{f_{lm}}\right) - \frac{1}{\left(1+\frac{\delta f}{f_{lm}}\right)}$$

$$F_+ = \frac{f_+}{f_{+m}} - \frac{f_{+m}}{f_+} = \left(1+\frac{\delta f}{f_{+m}}\right) - \frac{1}{\left(1+\frac{\delta f}{f_{+m}}\right)}. \quad (35)$$

From (33),

$$Q_lQ_+ = \left(\frac{C_0}{C_1}\right)^2. \quad (36)$$

[16]

Maximum bandwidth is obtained for equal load conductances.

$$g_l = g_+ = g = 2\pi C_1\sqrt{f_{lm}f_{+m}}. \quad (37)$$

By choosing other matched terminal admittances of the form shown in Fig. 4, in accordance with (24), symmetric double-peaked response curves and a wide variety of asymmetric response curves may be obtained. However, the present choice permits the simplest discussion.

In further discussion only the narrow-band case will be considered. Here,

$$\left|\frac{\delta f}{f_{lm}}\right| \ll 1$$

$$\left|\frac{\delta f}{f_{+m}}\right| \ll 1. \quad (38)$$

Thus we may set the numerator of the second term on the right-hand side of the two equations in (34) equal to 1. The input admittance of the nonlinear capacitor now may be represented by the equivalent circuit shown in Fig. 5, consisting of a parallel resonant circuit and a series resonant circuit in parallel. For the special terminal admittances of (33) both circuits are resonant at midband and the two resistors are equal; the restriction on the terminal conductances in (33) requires that the product of the separate bandwidths of the parallel and series resonant circuits of the equivalent circuit of Fig. 5 must be a constant. For equal terminal conductances, (37), these two bandwidths become equal. This condition can be seen to maximize the bandwidth over which the input power to the nonlinear capacitor remains constant, and hence to result in the greatest operating bandwidth for the nonlinear capacitor used as a modulator or as a demodulator.

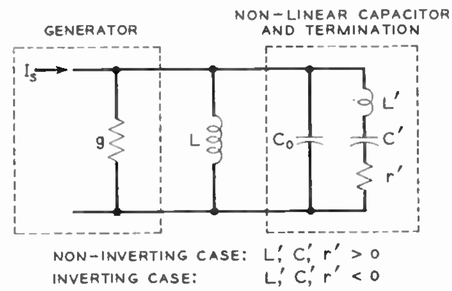

Fig. 5—Equivalent circuit for input admittance of nonlinear capacitor.

Making the terminal conductances equal, (37), and using the narrow-band approximation (38), (34)–(36) become

$$\frac{Y_{ls}}{g} = \frac{Y_{+s}}{g} = (1+jx) + \frac{1}{(1+jx)}$$

$$x = \frac{4\pi C_0}{g}\,\delta f = 2\,\frac{C_0}{C_1}\,\frac{1}{\sqrt{f_{lm}f_{+m}}}\,\delta f. \quad (39)$$

The modulator and demodulator transducer gains G_t are now determined in the narrow-band case from (22) and (39),

$$G_t = G_p\,\frac{4\ \mathrm{Re}\,\frac{1}{1+jx}}{\left|1+jx+\frac{1}{1+jx}\right|^2} = G_p\,\frac{1}{1+\left(\frac{x}{\sqrt{2}}\right)^4} \quad (40)$$

with corresponding subscripts on G_t and G_p. The power gains G_p are given by (17) and (32) as

$$G_{pl+} = \frac{f_+}{f_l} = \frac{f_{+m}}{f_{lm}}\,\frac{1+\frac{\delta f}{f_{+m}}}{1+\frac{\delta f}{f_{lm}}}$$

$$G_{p+l} = \frac{f_l}{f_+} = \frac{f_{lm}}{f_{+m}}\,\frac{1+\frac{\delta f}{f_{lm}}}{1+\frac{\delta f}{f_{+m}}}. \quad (41)$$

In the narrow-band approximation we neglect the frequency variation of G_p due to the second factor of (41), and substituting into (40) finally obtain for G_t

[16] The Q's of (34)–(36) and (51)–(53) should not be confused with the Q's of (8) and (10).

$$G_{tl+} = \frac{f_{+m}}{f_{lm}} \cdot F(x) \qquad G_{t+l} = \frac{f_{lm}}{f_{+m}} \cdot F(x)$$

$$F(x) = \frac{1}{1 + \left(\frac{x}{\sqrt{2}}\right)^4} \qquad x = 2\frac{C_0}{C_1}\frac{\delta f}{\sqrt{f_{lm}f_{+m}}}, \qquad (42)$$

for the modulator and demodulator gains, respectively. At midband (42) becomes

$$\left.\begin{aligned} G_{tl+} &= G_{pl+} = \frac{f_{+m}}{f_{lm}} \\ G_{t+l} &= G_{p+l} = \frac{f_{lm}}{f_{+m}} \end{aligned}\right\} \text{ for } x = 0. \qquad (43)$$

The 3-db bandwidth for both modulator and demodulator now is determined by setting $G_t = \frac{1}{2}G_p$.

$$F(x) = \tfrac{1}{2}$$
$$x = \pm\sqrt{2}. \qquad (44)$$

Then, from (42),

$$\delta f = \pm \frac{1}{2}\frac{C_1}{C_0}\sqrt{2f_{lm}f_{+m}} \qquad (45)$$

so that the 3-db bandwidth B for both the noninverting modulator and demodulator becomes

$$B = \frac{C_1}{C_0}\sqrt{2f_{lm}f_{+m}}$$
$$\frac{B}{f_{lm}} = \frac{C_1}{C_0}\sqrt{2\frac{f_{+m}}{f_{lm}}}. \qquad (46)$$

The present narrow-band discussion depends on having a small fractional bandwidth at the low-signal frequency, as stated in (38). Therefore, from (46) we must have

$$\frac{C_1}{C_0} \ll \sqrt{\frac{1}{2}\frac{f_{lm}}{f_{+m}}}. \qquad (47)$$

The ratio C_1/C_0 may be taken as a figure of merit for the nonlinear element. If the element becomes linear this quantity approaches 0; the maximum value it can attain for a nonlinear element whose q-v characteristic must always have a positive slope is 1, although this maximum value can be attained only for a highly idealized nonlinear capacitor.[17] Eq. (46) shows that even for a very modest amount of nonlinearity the fractional bandwidth at the low-signal frequency will be large for the usual case where the high-signal frequency is much greater than the low-signal frequency, $(f_+/f_l) \gg 1$. In fact the narrow-band analysis will be invalid for many cases of practical interest, although it gives a general indication of the response. Of course we may always return to (34) and perform an exact calculation for the cases of interest, but this will not be done here.

Finally, we consider the sensitivity of the device at midband to small changes in either of the terminal conductances or in the nonlinear capacitor parameter C_1. The sensitivity S is defined as the fractional change in G_t divided by the corresponding fractional change in the parameter causing the variation in gain.[18] Thus, zero sensitivity is desirable, and a high sensitivity undesirable in a device. In the present case the device is matched at midband, and so

$$S = \frac{dG_t/G_t}{dg/g} = \frac{dG_t/G_t}{dC_1/C_1} = 0. \qquad (48)$$

Let us consider briefly one numerical example. Let

$$f_{lm} = 100 \text{ mc} \qquad f_{+m} = 20{,}000 \text{ mc}$$
$$f_1 = 19{,}900 \text{ mc, local oscillator frequency}$$

Then the modulator gain at mid-band is

$$G_{tl+} = 200 \qquad 10 \log G_{tl+} = +23 \text{ db}.$$

The demodulator gain is

$$G_{t+l} = \frac{1}{200} \qquad 10 \log G_{t+l} = -23 \text{ db}.$$

The 3-db bandwidth is given by

$$B = 2000\frac{C_1}{C_0} \text{ mc} \qquad \left(\frac{B}{100}\right)_{mc} = 20\frac{C_1}{C_0}.$$

For a value of $C_1/C_0 = 1/20$, corresponding to a very small nonlinearity, the fractional bandwidth at the low-signal frequency, given by the narrow-band approximation, equals unity. Thus, for moderate nonlinearity we expect wide bandwidths from these devices.

Inverting Case

We now repeat the above analysis for the inverting case. Again consider the nonlinear capacitor terminated at the signal frequencies f_l and f_- as shown in Fig. 4. The terminal admittances are now chosen to yield a symmetric single-peaked response curve; they are again made parallel resonant at midband with their associated self-susceptances of the nonlinear capacitor matrix, as in (31). Now, however, the input conductance of the nonlinear capacitor is negative and match is impossible, and so the terminal conductances are no longer restricted by a condition such as the last equation in (31). Denoting the midband frequencies by f_{lm} and f_{-m} as before,

$$f_l = f_{lm} + \delta f$$
$$f_- = f_{-m} - \delta f$$
$$\delta f = 0 \text{ at midband}. \qquad (49)$$

[17] Torrey and Whitmer, *op. cit.*, pp. 410–411. The proof here is given for a nonlinear conductance, but applies equally well to the present case of a nonlinear capacitance.

[18] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., New York, N. Y.; 1945.

Then the terminal inductances L_l and L_- in Fig. 4 are given by

$$L_l = \frac{1}{4\pi^2 f_{lm}^2 C_0} \qquad L_- = \frac{1}{4\pi^2 f_{-m}^2 C_0}. \tag{50}$$

The admittances seen by the current sources in Fig. 4, $\underline{Y}_{l_s} = \underline{Y}_l + \underline{Y}_{l\,\mathrm{in}}$, may then be written

$$\frac{Y_{l_s}}{g_l} = 1 + jQ_l F_l - \frac{\alpha\left(1 + \frac{\delta f}{f_{lm}}\right)\left(1 - \frac{\delta f}{f_{-m}}\right)}{1 - jQ_- F_-}$$

$$\frac{Y_{-_s}}{g_-} = 1 + jQ_- F_- - \frac{\alpha\left(1 + \frac{\delta f}{f_{lm}}\right)\left(1 - \frac{\delta f}{f_{-m}}\right)}{1 - jQ_l F_l}, \tag{51}$$

where

$$Q_l = \frac{2\pi f_{lm} C_0}{g_l} \qquad Q_- = \frac{2\pi f_{-m} C_0}{g_-}$$

$$F_l = \frac{f_l}{f_{lm}} - \frac{f_{lm}}{f_l} = \left(1 + \frac{\delta f}{f_{lm}}\right) - \frac{1}{\left(1 + \frac{\delta f}{f_{lm}}\right)}$$

$$F_- = \frac{f_-}{f_{-m}} - \frac{f_{-m}}{f_-} = \left(1 - \frac{\delta f}{f_{-m}}\right) - \frac{1}{\left(1 - \frac{\delta f}{f_{-m}}\right)} \tag{52}$$

$$\alpha = \frac{4\pi^2 f_{lm} f_{-m} C_1^2}{g_l g_-} = Q_l Q_- \left(\frac{C_1}{C_0}\right)^2. \tag{53}$$

As discussed below, the parameter α determines the midband gain. For a fixed value of α, maximum bandwidth is again obtained for equal load conductances, as in the noninverting case.

$$g_l = g_- = g = 2\pi C_1 \sqrt{\frac{f_{lm} f_{-m}}{\alpha}}. \tag{54}$$

Considering again the narrow-band case,

$$\left|\frac{\delta f}{f_{lm}}\right| \ll 1$$

$$\left|\frac{\delta f}{f_{-m}}\right| \ll 1, \tag{55}$$

we set the numerator of the second term on the right-hand side of (51) equal to 1. Thus, the input admittance of the nonlinear capacitor is again represented by the equivalent circuit of Fig. 5, except that all three elements of the series resonant circuit are negative. For the terminal admittances of (50) both circuits are resonant at midband. In this case, the parameter α determines whether this network is stable or unstable,

$$0 < \alpha < 1 \text{ stable}$$

$$\alpha > 1 \text{ unstable}, \tag{56}$$

as shown by a Nyquist diagram. While this method of studying the stability is not strictly valid, it is assumed to give a reliable indication of stability.

For a fixed value of α the product of the separate bandwidths of the parallel and series circuits of Fig. 5 must be a constant. For equal terminal conductances (54) these two bandwidths become equal. This condition can be seen to maximize the bandwidth over which the power returned by the nonlinear capacitor to the source remains constant, and hence to yield the greatest operating bandwidth for the inverting modulator or demodulator.

For equal terminal conductances (54), in the narrow-band case (55) the admittances (51)–(53) become

$$\frac{Y_{l_s}}{g} = \frac{Y_{-_s}}{g} = (1 + jx) - \frac{\alpha}{(1 + jx)}$$

$$x = \frac{4\pi C_0}{g}\delta f = 2\frac{C_0}{C_1}\sqrt{\frac{\alpha}{f_{lm} f_{-m}}}\,\delta f. \tag{57}$$

Under stable operating conditions, $0 < \alpha < 1$, the modulator and demodulator transducer gains are given by (22) and (57);

$$G_t = G_p \frac{4\,\mathrm{Re}\,\frac{-\alpha}{1 + jx}}{\left|1 + jx - \frac{\alpha}{1 + jx}\right|^2} = -G_p \frac{4\alpha}{(1-\alpha)^2} \cdot \frac{1 + x^2}{1 + \frac{(3+\alpha^2)}{(1-\alpha)^2}x^2 + \frac{(3+2\alpha)}{(1-\alpha)^2}x^4 + \frac{1}{(1-\alpha)^2}x^6}. \tag{58}$$

Making the usual narrow-band approximations for G_p (17), and noting that G_p is negative, we obtain for the modulator and demodulator gains

$$G_{tl-} = \frac{f_{-m}}{f_{lm}} \frac{4\alpha}{(1-\alpha)^2} G(x, \alpha),$$

$$G_{t-l} = \frac{f_{lm}}{f_{-m}} \frac{4\alpha}{(1-\alpha)^2} G(x, \alpha)$$

$$G(x, \alpha) = \frac{1 + x^2}{1 + \frac{(3+\alpha^2)}{(1-\alpha)^2}x^2 + \frac{(3+2\alpha)}{(1-\alpha)^2}x^4 + \frac{1}{(1-\alpha)^2}x^6},$$

$$x = 2\frac{C_0}{C_1}\sqrt{\frac{\alpha}{f_{lm} f_{-m}}}\,\delta f. \tag{59}$$

The midband gains are no longer equal to the ratio of output-to-input frequency, but become

$$\left.\begin{aligned} G_{tl-} &= \frac{f_{-m}}{f_{lm}} \cdot \frac{4\alpha}{(1-\alpha)^2} \\ G_{t-l} &= \frac{f_{lm}}{f_{-m}} \cdot \frac{4\alpha}{(1-\alpha)^2} \end{aligned}\right\} \text{ for } x = 0. \tag{60}$$

The factor containing α varies from 0 to ∞ as α varies from 0 to 1, so that the midband transducer gain may be as large as we please. The 3-db bandwidth in both cases is again found by solving

$$G(x, \alpha) = \tfrac{1}{2} \tag{61}$$

for x and substituting into (59) to determine the bandwidth as a function of α. This cannot be done in closed form because of the complexity of $G(x, \alpha)$, so we will study two special cases to illustrate the general behavior.

First, consider the behavior close to instability, where α is almost equal to 1. Then, the solution to (61) is found to be

$$x = \pm \tfrac{1}{2}(1 - \alpha). \tag{62}$$

From (59),

$$\delta f = \pm \frac{1}{4} \frac{C_1}{C_0} \sqrt{f_{lm} f_{-m}}(1 - \alpha), \tag{63}$$

so the 3-db bandwidth for both the inverting modulator and demodulator becomes

$$\left.\begin{aligned} B &= \frac{1}{2} \frac{C_1}{C_0} \sqrt{f_{lm} f_{-m}}(1 - \alpha) \\ \frac{B}{f_{lm}} &= \frac{1}{2} \frac{C_1}{C_0} \sqrt{\frac{f_{-m}}{f_{lm}}}(1 - \alpha) \end{aligned}\right\} \text{ for } (1 - \alpha) \ll 1. \tag{64}$$

This is similar to (46) for the noninverting case, except for the factor $(1-\alpha)$. From (64) and (60) we see that as $\alpha \rightarrow 1$ and instability is approached, the gain increases to infinity and the bandwidth decreases to zero. We have

$$\left.\begin{aligned} \sqrt{G_{tl-}} \cdot B &= \frac{C_1}{C_0} f_{-m} \\ \sqrt{G_{t-l}} \cdot B &= \frac{C_1}{C_0} f_{lm} \end{aligned}\right\} \text{ for } (1 - \alpha) \ll 1, \tag{65}$$

showing the way in which bandwidth may be exchanged for gain.

Second, consider the behavior when the midband gains are equal to the ratio of output to input frequency, as in the noninverting case. From (60),

$$\frac{4\alpha}{(1 - \alpha)^2} = 1 \qquad \alpha = 0.172. \tag{66}$$

The solution to (61) is then,

$$x = \pm 0.513, \tag{67}$$

so that from (59),

$$\delta f = \pm \frac{1.24}{2} \frac{C_1}{C_0} \sqrt{f_{lm} f_{-m}} \tag{68}$$

and the 3-db bandwidth for both modulator and demodulator becomes

$$\left.\begin{aligned} B &= 1.24 \frac{C_1}{C_0} \sqrt{f_{lm} f_{-m}} \\ \frac{B}{f_{lm}} &= 1.24 \frac{C_1}{C_0} \sqrt{\frac{f_{-m}}{f_{lm}}} \end{aligned}\right\} \text{ for } \alpha = 0.172. \tag{69}$$

Comparison with (46) shows, for equal midband gains and signal frequencies, the inverting device will have slightly less bandwidth than the noninverting device.

Finally, the sensitivities at midband to small changes in either of the terminal conductances or in the parameter C_1 are given by

$$S_g = \frac{dG_t/G_t}{dg/g} = -\frac{1 + \alpha}{1 - \alpha}, \qquad S_{C_1} = \frac{dG_t/G_t}{dC_1/C_1} = -2S_g. \tag{70}$$

As $\alpha \rightarrow 1$ and instability is approached the sensitivity $\rightarrow \infty$.

Let us take as an example the same signal frequencies used in the previous section to illustrate the noninverting modulator and demodulator, and consider the behavior of the inverting modulator and demodulator for several values of α.

$f_{lm} = 100$ mc $f_{-m} = 20{,}000$ mc
$f_l = 20{,}100$ mc, local oscillator frequency

α	Modulator Gain 10 log G_{tl-}	Demodulator Gain 10 log G_{t-l}	Bandwidth B	Sensitivity (terminal conductance) S_g
0.172	+23 db	−23 db	1752 C_1/C_0 mc	−1.414
0.9	+48.5 db	+2.5 db	70.7 C_1/C_0 mc	−19
0.99	+69 db	+23 db	7.07 C_1/C_0 mc	−199

In the first case, $\alpha = 0.172$, the modulator and demodulator gains are simply the ratio of output to input frequency, as in the noninverting case. The bandwidth of the inverting modulator and demodulator is slightly less than that of the corresponding noninverting device. By operating close enough to instability the midband gain can be made as large as desired. However, as $\alpha \rightarrow 1$ the bandwidth will become very small and the sensitivity very large. Thus, for $\alpha = 0.99$ an infinitesimal decrease of x per cent in either source or load conductance will cause an increase in gain of $199x$ per cent; a 1 per cent decrease in either terminal conductance or a $\frac{1}{2}$ per cent increase in C_1 (caused perhaps by a slight variation in the local oscillator drive applied to the nonlinear element) will cause the device to become unstable.

The possibility of exchanging bandwidth for gain in the inverting case may permit us to use to better advantage a nonlinear element with a large nonlinearity as an inverting modulator or demodulator than as a noninverting device; in the above example, if C_1/C_0 is close to 1, substantially greater gains at moderate bandwidths can be obtained if a moderate sensitivity is not objectionable. However, we see that a demodulator with high gain will have a very narrow band and a high sensitivity, and such a device would not appear to be too practical.

Nonlinear Capacitor Negative Conductance Amplifiers

We now consider the use of an inverting nonlinear capacitor device as a single-frequency amplifier, by utilizing its negative input conductance. A circulator is assumed connected at the input port corresponding to the frequency to be amplified, as shown in Fig. 3, so that the incident and the (amplified) reflected waves may be separated. The discussion in the last section regarding the choice of terminal admittances to yield a symmetric single-peaked over-all response curve of maximum gain and bandwidth applies equally well to the present case. Thus (49)–(57) remain valid for the negative conductance amplifier; and from (29) and (57), or from (28) and (59) we have

$$G_{tll} = G_{t--} = 1 + \frac{4\alpha}{(1-\alpha)^2} G(x, \alpha) \qquad (71)$$

with $G(x, \alpha)$ as given in (59). The response is the same at both the low and the high signal frequency. At midband the gains become

$$G_{tll} = G_{t--} = 1 + \frac{4\alpha}{(1-\alpha)^2} = \left(\frac{1+\alpha}{1-\alpha}\right)^2 \text{ for } x = 0. \qquad (72)$$

This type of operation will be of interest only when α is close to 1 so that substantial gains can be obtained; under these conditions the term unity in (71) and (72) may be neglected compared to the second term, so that the 3-db bandwidths are given approximately by

$$G(x, \alpha) = \tfrac{1}{2}, \qquad (73)$$

which is the same as (61) for the modulator-demodulator bandwidths. Therefore, the bandwidth as a negative conductance amplifier is the same as (64),

$$B = \frac{1}{2}\frac{C_1}{C_0}\sqrt{f_{lm}f_{-m}}(1-\alpha) \quad \text{for } (1-\alpha) \ll 1; \quad (74)$$

and the gain-bandwidth product is

$$\sqrt{G_t}\cdot B = \frac{C_1}{C_0}\sqrt{f_{lm}f_{-m}} \quad \text{for } (1-\alpha) \ll 1. \qquad (75)$$

The midband sensitivities to small variations in either terminal conductance or in C_1 are

$$S_g = \frac{dG_t/G_t}{dg/g} = -\frac{4\alpha}{1+\alpha}\cdot\frac{1}{1-\alpha} \to -\frac{2}{1-\alpha} \quad \text{for } (1-\alpha) \ll 1,$$

$$S_{C_1} = -2S_g, \qquad (76)$$

the same as (70) for α close to 1.

In considering a numerical example we choose signal frequencies different than those used in discussing modulators and demodulators, since there is no reason now for one of the signal components to be a low (IF) frequency.

$f_{lm} = 10{,}000$ mc $f_{-m} = 20{,}000$ mc
$f_1 = 30{,}000$ mc, local oscillator frequency

α	Negative Conductance Gain $10 \log G_{tll} = 10 \log G_{t--}$	Bandwidth B	Sensitivity (terminal conductance) S_g
0.9	+25.6 db	707 C_1/C_0 mc	−18.95
0.99	+46.0 db	70.7 C_1/C_0 mc	−199.0

Once again as $\alpha \to 1$ the gain and sensitivity approach infinity, the bandwidth approaches zero. If the nonlinear element has a reasonably large nonlinearity, moderate gains and bandwidths at either signal frequency can be obtained, with of course a moderate sensitivity to variations in the terminal conductances or in the parameter C_1.

General Energy Relations for Linear Variable Reactors

Since the general energy relations of Part I for nonlinear elements are valid for arbitrary levels at the various frequencies present, they of course must hold when the signal levels are small, as shown in the special case considered here. This has been shown in general for the small-signal case, including all of the frequencies mf_1, $mf_1 \pm f_0$, by using the general matrix equations of the equivalent linear variable reactor.[9]

We may show, in a simple way, that these general energy relations apply to any lossless linear variable reactor varied periodically in time, for arbitrary signal levels. Consider linear inductors and capacitors that are varied mechanically and have no mechanical or electrical losses. Then (24) and (25) of Part I[1] will remain valid if W_{mn} is now taken to be the sum of both the electrical and mechanical power supplied to the variable reactor at the frequency $|mf_1+nf_0|$. For convenience let the fundamental frequency at which the reactance is varied be f_1, and the fundamental frequency of the applied electrical signal f_0. Then for $n \neq 0$, W_{mn} will contain only electrical energy; mechanical energy can occur only for $n=0$.

First we note that such a variable reactor may be replaced by an equivalent nonlinear reactor, plus an associated local oscillator, which for small signals is identical as far as the signal-frequency terminals are concerned. The nonlinear characteristic and the local oscillator driving waveform, consisting of the frequencies mf_1, must be chosen so that the equivalent time-varying reactance is identical with that of the original variable reactor, and so that the signal components are small compared to the local oscillator component. There are, of course, a great many ways of doing this, and consequently, there is not necessarily any correspondence whatever between the mechanical powers in the variable reactor and the local oscillator powers in the equivalent nonlinear reactor at the frequencies mf_1; further, the limit on the signal levels in order that they will be small compared to the local oscillator level will obviously depend on the choice of equivalent nonlinear element and local oscillator waveform.

The equivalent nonlinear reactor is, of course, governed by the general energy relations (24) and (25) of Part I. However, (25) contains only signal-frequency terms, since the $n=0$ terms $W_{m,0}$ are absent. Since the two devices are equivalent as far as the signal frequencies are concerned, (25) must apply also to the linear variable reactor for small signals. But since (25) is linear in the $W_{m,n}$ and since the variable reactor is linear in the electrical signals, (25) is valid for arbitrary signal levels. The first energy relation (24) may be obtained now from (25) and the conservation of electrical and mechanical energy.

Thus, the general energy relations of Part I for nonlinear reactors apply to linear variable reactors as well if the $W_{m,n}$ are suitably interpreted to include mechanical as well as electrical power.[19]

Discussion

The derivation of the general energy relations for nonlinear capacitors and inductors in Part I depended only on having a single-valued nonlinear characteristic; consequently, these relations tell us nothing about how the amount of nonlinearity affects the operation of the various devices. The present small-signal analysis shows, for two types of nonlinear reactor devices, the limitations imposed by the shape of the nonlinear characteristic. While the energy relations and the gains are independent of this characteristic, the bandwidths are limited and depend on the amount of nonlinearity; as the nonlinearity goes to zero so does the bandwidth. In contrast, for a nonlinear resistor the gain depends on the amount of nonlinearity and must be less than 1, while the bandwidth is limited only by the external circuits.

In the usual case of widely separated signal frequencies the noninverting modulator will have a high gain and the demodulator a high loss. Only a relatively small amount of nonlinearity will be required to attain a bandwidth equal to the low-signal frequency. Since this is a stable device and yields maximum gain for matched terminal admittances, the gain cannot be increased at the expense of bandwidth, as is possible with the inverting device.

The inverting device is potentially unstable and has a negative input conductance; consequently, it can never be matched. Its gain as a modulator or demodulator may be varied from 0 to ∞ by changing its terminal admittances; the ratio of modulator to demodulator gain remains the same as for the noninverting device. The price of high gain is narrow bandwidth and high sensitivity to changes in the terminal admittances or to changes in C_1 (due to variations in the local oscillator drive). If the modulator and demodulator gains are made the same as in the corresponding noninverting case, so that the modulator gain and demodulator loss are equal, the bandwidth will be almost as large as in the corresponding noninverting modulator and demodulator, so that again only a relatively small amount of nonlinearity is required to yield a bandwidth equal to the low-signal frequency. If a larger nonlinearity is available, then substantially greater gains can be obtained at moderately wide bandwidths and moderate sensitivities. If, however, the gains are increased to the point where the demodulator has a high gain, the bandwidth will be severely limited and the sensitivity very large.

As a negative conductance amplifier, a nonlinear capacitor in the inverting case can provide amplification at a single frequency with moderate gain, bandwidth, and sensitivity; the larger the nonlinearity the larger the bandwidth. Here again gain and bandwidth may be exchanged within wide limits, but very high gains are accompanied by narrow bandwidth and high sensitivity.

Various extensions in the present analysis may be made. First, terminal admittances other than the simple ones studied here may be of interest; increased bandwidth can be obtained if a double-peaked characteristic is acceptable. If several of these devices are cascaded the usual stagger tuning techniques may be of use. Next, the present narrow-band analysis may be extended to the wide-band case in a straightforward manner. Finally, the case where both sidebands $f_1 \pm f_l$ are present may be treated in a similar manner;[20] since the terminations at all three signal frequencies now must be specified, the analysis contains several additional parameters.[21]

Acknowledgment

The author wishes to thank J. M. Manley and M. T. Weiss for many helpful discussions and for their detailed suggestions during the preparation of this paper.

[19] A. E. Siegman, private communication.

[20] M. E. Hines, forthcoming paper.

[21] A related paper by S. Bloom and K. K. N. Chang, "Theory of parametric amplification using nonlinear reactances," *RCA Rev.*, vol. 18, pp. 578–593; December, 1957, came to the author's attention after the present paper had been submitted. It considers the gain and bandwidth of an inverting negative conductance amplifier whose nonlinear element has a quadratic characteristic under different terminal conditions.

Very Low-Noise Traveling-Wave Amplifier*

E. W. KINAMAN†, MEMBER, IRE, AND M. MAGID†

Summary—Recent improvements in design and techniques are described which have lowered the noise figure of a developmental traveling-wave amplifier from 9 db to 6 db. Some selected tubes have had noise figures as low as 4.8 db. The improved tube is now commercially available as the RCA-6861 low-noise traveling-wave tube.

A general explanation of beam noise is presented and the presently employed noise-reducing schemes and theories are discussed in relation to the 6861. The specific factors contributing to the very low-noise figures are evaluated, including low helix loss, low QC with high circuit impedance, maximum beam diameter with minimum intercepted current, a flexible low-noise gun, and, most important, a smooth and highly emissive dense oxide cathode operating at about 600°C.

Application considerations are also discussed, including phase sensitivity, life, saturation effects (modulation and harmonic generation), allowable voltage variations, and typical noise figure, gain, and match over the 2700-to-3500-mc frequency range of the tube.

FOR many years, there has been a need for a broadband rf preamplifier tube operating at microwave frequencies to amplify very weak signals in receiver applications. The usefulness of such a tube is measured by the minimum detectable signal strength which, in turn, is determined by the amount of noise the tube adds to the signal.

The RCA-6861 (Fig. 1), an *S*-band, low-noise, traveling-wave-tube amplifier suitable for such applications is described. The 6861 has a gain of 25 db and maximum power output of several milliwatts. Its average noise figure is below the presently accepted theoretical limit. Nine of these tubes have shown tube noise figures at 5 db or less; the lowest noise figure associated with the tube being 4.8 db. From Fig. 2 it can be seen that the average tube noise figure is 5.5 db. This graph was compiled from the last 85 tubes, 53 of which could be operated at the normal voltage and current conditions. The lowest noise figure measured on each tube is indicated on the graph and it can be seen that every tube was capable of at least a 6.3-db noise figure (6.1 db is the minimum noise figure as computed using the presently accepted classical theory including the circuit loss and space-charge factor of the 6861). The tube and components, shown in Fig. 1, are derived from a prototype described by Peter in 1952 [1] when an 8.5-db noise figure was reported at 3000 mc. Circuit and fabrication redesign as described herein have resulted in the first commercially available low-noise traveling-wave tube and some of its developmental variations covering the 2 to 4.3-kmc frequency range (Fig. 1 and Fig. 7).

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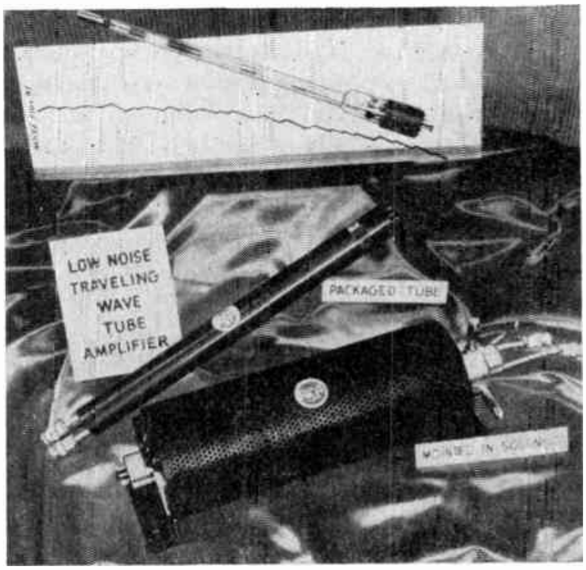


Fig. 1—Photograph of the RCA 6861 showing tube, capsule, and solenoid.

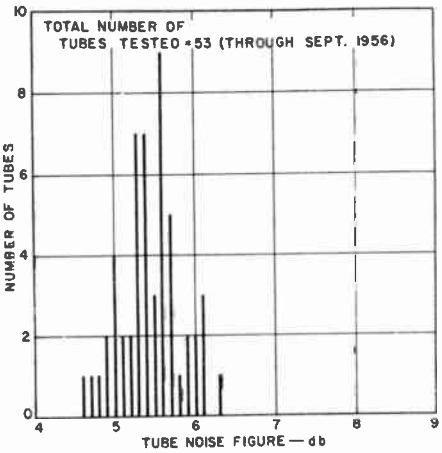


Fig. 2—Noise figures of various 6861 tubes.

NOISE FIGURE

The limiting sensitivity of a traveling-wave tube (or any receiver) is set by the amount of noise generated within the tube. The most commonly utilized formula describing the sensitivity of a traveling-wave tube is noise figure. This is numerically equal to the db difference between the input and output signal-to-noise power ratio when they are expressed in db. Or, stated another way (which leads to simplified measurement techniques), it is the ratio of the output noise power of a tube to the output of a noiseless tube having the same gain and bandwidth. (The input of the tube is terminated in a load at 290°K in these definitions.)

When no beam current is flowing in this traveling-wave tube, the tube acts as a microwave attenuator having an attenuation of 65 to 100 db. When rated beam current is drawn, the tube is an amplifier having a gain of 25 db and a terminal noise figure[1] of 6 db. The electron beam used in the traveling-wave tube to amplify microwave signals also amplifies a spectrum of its own inherent noise. If the beam were operated at the velocity required to produce microwave-signal amplification without regard to noise-reduction techniques, the tube noise figure would probably be 20 to 40 db. Although such a tube might perform well as an amplifier, it could not detect very low signal levels. Therefore, the electron gun used in low-noise traveling-wave tubes is specifically designed to minimize the contribution of beam noise to the output signal.

General Explanation of Noise

An understanding of the function of this gun may be aided by a brief discussion of noise theory. Electrons are emitted randomly from the cathode and move away from it under the influence of the accelerating electric field of the anode, exhibiting minute random variations in density and velocity. These random variations are the sources of noise in the beam. The motion of the individual electrons, which constitute the moving beam, is elastic in nature. When perturbed, the electrons oscillate at their plasma frequency. This oscillation is propagated along the beam in the form of space-charge waves. Similarly, the current and velocity fluctuations produce space-charge waves which are 90° out of phase with each other. It was originally thought that the cathode current fluctuations (the assumption here is that the mean square fluctuation in the convection current from a temperature limited cathode is $\overline{i^2}=2eT_c\Delta f$, full "shot" noise) would be almost completely cushioned if the cathode were operated under space-charge-limited conditions. Thus, the noise was considered as essentially one standing wave on the beam (resulting from the assumption of "Rack" velocity excitation at the cathode—Rack calculated the fluctuations about the average velocity of a current I_0 as

$$\overline{V^2} = (4-\pi)\frac{e}{m}\frac{kT_c\Delta f}{I_0}.$$

This line of thinking logically led to methods for reducing the coupling and magnitude of the beam noise based on only one standing wave of noise.

Noise Reduction Methods

Jump Gun

An important contribution to the art was the development of the "velocity-jump" principle described by Watkins [2]. In this method, the noise arising from the beam velocity fluctuations coupled to the helix is reduced by rapid acceleration of the beam in a very short distance. This technique produces a standing wave of noise on the beam. The helix was placed to couple a minimum amount of noise from the beam, thereby causing significant noise reduction. It had been suggested that successive jumps would lower the noise figure even more. Further work pointed out that current fluctuations are not completely damped by the potential minimum in front of the cathode and, more important, that the same device which deamplifies the velocity fluctuations increases the contribution from current fluctuations. Consequently, the gun cannot reduce the noise figure below a minimum value proportional to the product of the velocity and current fluctuations at the cathode.

Three-Region Gun

The three-region gun [1], [3], [4] increases the average electron velocity smoothly rather than by discrete potential jumps. The standing-wave ratio of noise current and the position of the noise-current minimum in the helix-input region can be adjusted by variation of the gun electrode potentials to compensate for a considerable range of frequencies and beam currents. This adjustability is important because minimum noise figure is obtained only for definite optimum values of these parameters. The three-region gun is used in the 6861 and can be operated over the entire frequency range from 2000 to 4300 mc by proper adjustment of the grid voltages. The beam current and tube gain are not materially affected by the voltage adjustment for minimum noise.

A further advantage of the three-region gun is that it operates without sharp potential discontinuities along the beam. This advantage is important because such discontinuities produce electrostatic lenses [4] which can increase the noise of the beam and, therefore, the noise figure of the traveling-wave tube.

Comparison of Results with Theory

Calculations for the 6861 based on first-order theory [5], [6], [9] indicate a theoretical minimum tube noise figure of 5.8 db[2] for a cathode temperature of 650°C. (When helix circuit loss and space charge for the 6861 [listed in Fig. 4] are considered, this figure rises to approximately 6.1 db.) This minimum noise figure is based on the assumption that the velocity distribution is that computed by Rack, the current fluctuation is full shot noise, and no correlation exists between the two fluctuations. Continued investigations resulted in predictions of noise figures ranging from 0 to 6 db, each with seemingly logical support of its conclusion. An

[1] The terminal noise figure includes both tube noise and the loss of the input coupler and cable. Unless otherwise noted, noise figures given in this paper are tube noise figures, which are approximately 0.4 db less than the terminal noise figures.

[2] Assuming no correlation, space charge, or helix loss, Pierce and Danielson [9] calculated the minimum noise figure as

$$1+(4-\pi)\frac{T_{\text{cathode}}}{290°\text{K}}$$

(Temperature is in absolute units).

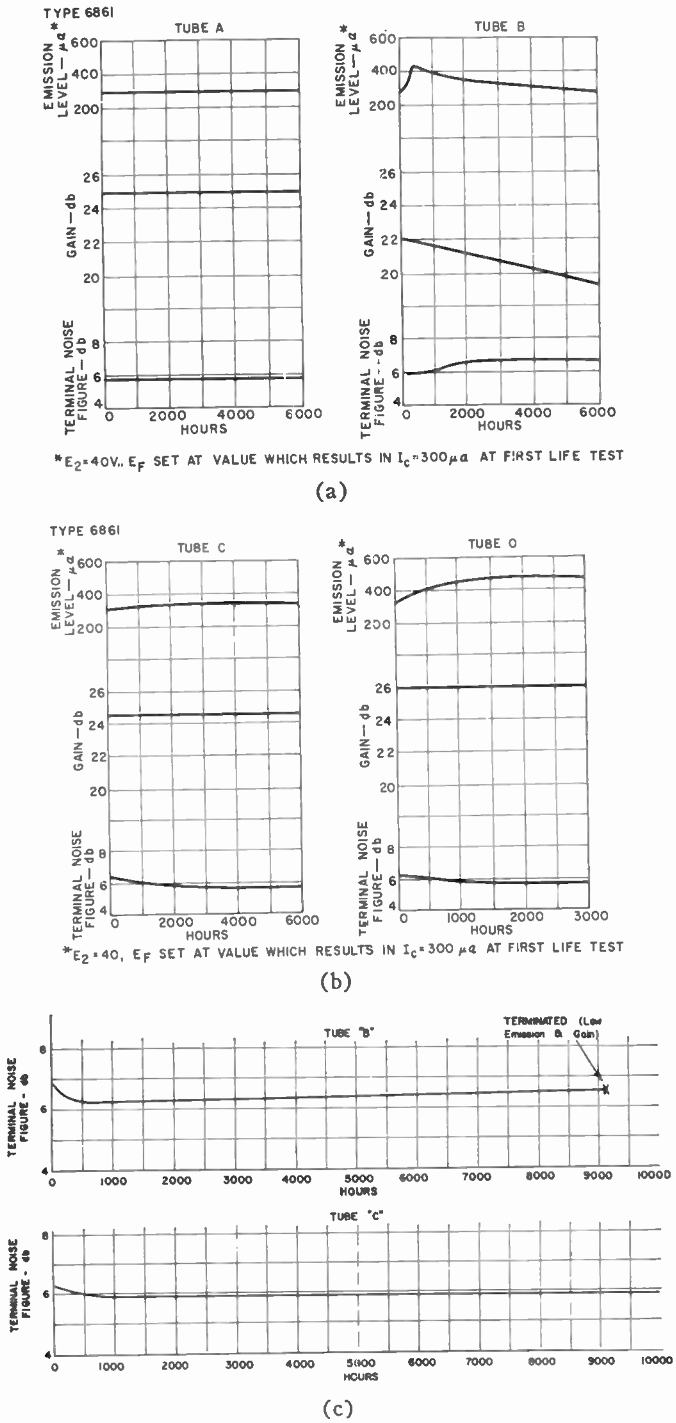


Fig. 3—Life characteristics.

earlier theory of Watkins [7] stated that the current fluctuations are smoothed by the space-charge-operated cathode to such an extent that a noise figure of 4.7 db is theoretically possible. Calculations based on this theory do not indicate any major reduction from the noise-figure value mentioned above for the 6861.

Life-test results indicate that a major factor in the low-noise performance of the 6861 is the ratio of the temperature-limited or available cathode current to the space-charge-limited or beam current. The fluctuations of a quantity proportional to this ratio with tube life seem to show up directly as noise-figure fluctuations, as shown in Fig. 3(a) and Fig. 3(b). For example, some tubes do not reach optimum performance for several hundred hours (depending upon their activation schedule), during which time the available current at operating cathode temperature is increasing and the noise figure is decreasing. Tubes whose noise figure increases slightly also show decreasing electron availability. The extra electrons available in tubes which improve on life may find their use in smoothing the current fluctuations, or the increased ratio of available to beam current may result in more uniform emission; both factors result in reduced noise figure. The highly emissive cathodes used in these tubes show ratios of available to beam current of 10 to 20 at the operating brightness temperature of 600 to 650°C.

Bloom [8] recently pointed out that full correlation of the velocity and current fluctuations at the potential minimum could result in a "0" db noise figure. However, no experimental evidence of correlation has yet been acknowledged in low-noise traveling-wave tubes.

An analysis of the effect on the noise figure by the higher-order space-charge [4] modes suggests that tubes should be designed for high "γb" (*i.e.*, large-beam, low-voltage operation). Since the cathode noise is distributed in many modes on the beam, Beam has pointed out that at high "γb" a larger percentage of the total noise power is in the higher order modes which do not couple as strongly to the helix as the lower order modes. This would appear to reduce the expected minimum noise figure, perhaps by as much as several db.

Description of Tube

Structural Details

A schematic diagram of the 6861 is shown in Fig. 4. The electron gun consists of a heater assembly, an oxide-coated cathode having a diameter of 0.040 inch, and a series of nichrome grids. During fabrication, the grids are mounted on a precision jig and their aperture edges are automatically "pinched" into 3 supporting pyrex glass beads. The cathode and heater assemblies are then welded in place to complete the gun.

The helix, made of 0.005-inch tungsten wire, is wound to 88 turns per inch on molybdenum mandrels in a precision helix-winding machine. The spacing between the end turns is increased exponentially for matching purposes by a cam-actuated mechanism. The eight-inch-long helix is fired on the mandrel to "set" the winding and is then silver-plated to improve conductivity. Three ceramic rods (which include lumped, density-tapered, thin-aquadag-film attenuators) are assembled around the helix and clamped in place. Because these precision rods extend more than an inch beyond the helix input, they are used to align the gun and helix with the "drift tube" (grid No. 4). Subsequent clamping of these rods eliminates the need for further precision alignment between the gun and the helix.

TWT PARAMETERS

Helix Voltage	$V_0=375$ volts
Beam Current	$I_0=150$ μa
Center Frequency	$f_0=3100$ mc
	$k_a=0.07$
	$\gamma_a=1.84$
Dielectric Loading Factor	$DLF=0.75$
Beam-to-Helix Ratio	$b/a=0.5$
Gain Parameter	$C=0.018$
Space Charge Parameter	$Q=4$
Space Charge Factor	$QC=0.072$
Active Circuit Wave Lengths	$N=49.3$
Attenuation Parameter	$d=0.06$
Initial Loss Parameter	$A=-7.9$
Gain Parameter	$B=43$
Total Loss	$L=-6.3$ db
Calculated Gain	$G=A+BCN-L=24$ db

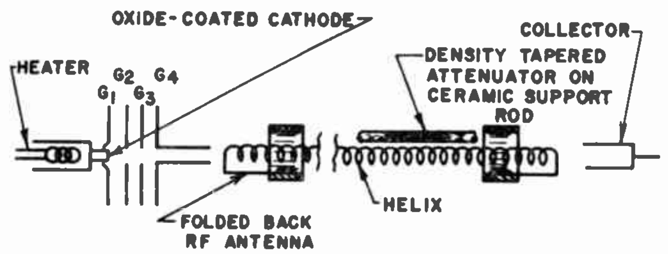


Fig. 4—Schematic diagram and twt parameters.

The collector is made of Kovar to facilitate sealing to the glass envelope, and includes a copper tubulation which permits adequate flushing during sealing. This tubulation is subsequently "pinched off."

The entire assembly is mounted in a precision-ground hard-glass envelope (or bulb). The helix assembly incorporates centering rings and flat springs which hold the assembly tightly in the center of the bulb. The finished tube includes an octal base and a collector-pinch-off protector cap.

Coaxial couplers are matched to the tube, as shown in Fig. 5, and the entire unit is assembled in an anodized aluminum capsule. The unit is then inserted into a solenoid which supplies a magnetic field of 500 to 550 Gauss of sufficient length and precision; the tube, as shown in Fig. 1, is then ready for use. Alignment for minimum helix-current interception is accomplished by means of the solenoid focusing gear. The tube is adjusted for minimum noise figure by adjustment of the helix voltage for maximum gain and the other elements for minimum noise output.

Specific Considerations of Low-Noise Tube Design

As yet it is not apparent what exactly to expect theoretically in the way of a minimum noise figure for the 6861. The steps followed in obtaining the present results on the 6861 are quite apparent. The helix and beam were designed (see Fig. 4) to minimize the space-charge factor, QC, which in practical tubes should be as low as possible [5], [9]. The low current and small beam diameter have an additional advantage, due to the lessened space-charge depression of gun voltages in the beam, of providing a noise swr of better definition

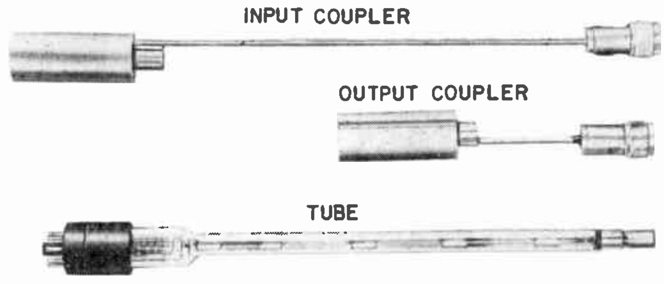


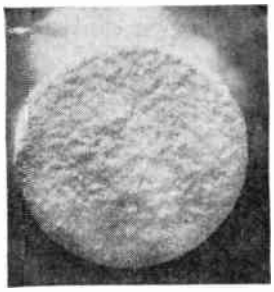
Fig. 5—Tube and matching coaxial couplers.

(increased electron synchronization), without resorting to the complexity of a hollow beam and center gun electrodes. The gain per inch was made as high as feasible by increasing the helix impedance and the beam-helix coupling to minimize the effect of the helix loss. (These changes are now noted as being in the same direction suggested for minimization of the noise figure by optimization of the space-charge modes [4].) The beam diameter was chosen to provide a satisfactory compromise between maximum helix-beam coupling and minimum intercepted current. The small helix diameter used requires unusual care in the alignment of the cathode, gun, and helix because the slightest misalignment causes high helix current from the beam and, in turn, high noise. The electron gun was empirically derived to provide an approximately exponential change of beam impedance [3], [5] in a minimum distance.

It has been pointed out [8] that optimization of the electron gun is not dependent on the magnitude or correlation of the velocity and current fluctuations arising from the region of minimum cathode potential. In other words, the minimum noise figure is not limited by the well designed electron gun, but is a function of the cathode alone. The noise performance of the cathode is, therefore, of prime importance. Although the cathode emits enough electrons to supply the rated beam current at a brightness temperature of 550 to 600°C, its operation is temperature-limited. A temperature of 600 to 650°C provides adequately uniform emission for minimum noise and comparatively long life.

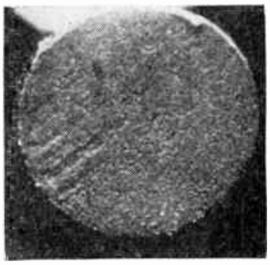
The high emission and successful life of the cathode is the result of techniques developed to activate the cathode without materially changing the surface from its original sprayed condition (see Fig. 6). Low tube pressure and cathode temperature are maintained during every phase of tube degassing and activation, and a high flushing rate is used during stem sealing. Once this had been mastered further concentration on spraying techniques resulted in repeatably low tube noise figures. The N109 nickel cathode base is sprayed to a thickness of 0.7 mil with a very wet mixture of triple-carbonate material, binder, and amyl-acetate. The use of a "wet" spray provides coating density above 1.3 grams/cc and maximum surface smoothness. The spray jigs are precisely designed, and include such features as 0.001-

TYPICAL OXIDE-COATED CATHODE AFTER SPRAYING AND BEFORE USE IN TWT



OXIDE-COATED CATHODE PICTURED BEFORE USE IN TWT (RESULTS 6 db NF)

THIS CATHODE RUN IN A TWT, WAS TESTED 760 HOURS WITH THE OPTIMUM NF DEGRADING 1 db. SEVERE WARPING, CRACKING AND GLAZING ARE QUITE EVIDENT.



IN 21 HOURS OF TUBE LIFE THE EMISSION DROPPED 70% AND NF DEGRADED 8 db. NOTE IN PARTICULAR THE THICKNESS OF COATING

Fig. 6—Typical oxide cathodes photographs.

inch air pockets around the blank to minimize side spray. To keep a gas-free product, vacuum firing and storage of parts was introduced. This had the added advantage of resulting in less but more careful parts handling.

Fig. 6 shows enlarged photographs (approximately 40×) of the coating on several cathodes. Realizing the shades of the coating are largely a function of the photography techniques, the smooth dense cathodes in the upper half can be contrasted to the solidified cracked residues below them. Careful attention, via standard techniques, to the processing of the tube and its parts eliminated this type of result.

Test Results

The 6861 is designed to operate with fixed voltages over the entire frequency range of 2700 to 3500 mc. At all frequencies within this band, the terminal noise figure is less than 7 db, while the average terminal noise figure at all frequencies within the band is 6 db. Variation of the average noise figure with frequency is shown in Fig. 7. The gain function is approximately the inverse of the noise-figure function, and the average gain is 25 db. Because a perfect match to the helix and the attenuator is not attained at all frequencies; the noise figure and gain curves have a more jagged fine-frequency response. Typically, the peaks and valleys are within ±0.2 db of the average noise figure characteristics and ±0.8 db of the average gain curve.

The 6861 is matched to a 50-ohm coaxial line (Type *N* connectors) by the use of cavity-type couplers. These

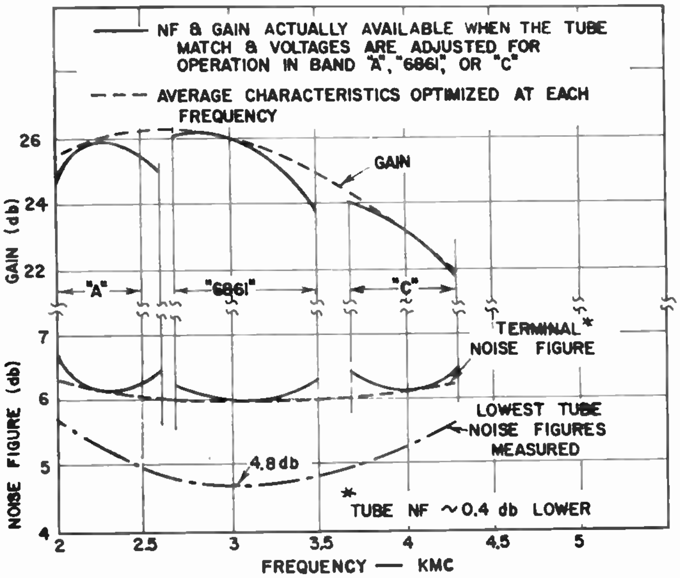


Fig. 7—Broad-band characteristics of the *S*-band low-noise twt.

couplers are factory-adjusted for a vswr of less than 1.6 on the tube input and less than 1.9 on the tube output in the frequency range from 2600 to 3600 mc.

Matching

The voltage standing-wave ratio of the 6861 when no voltages are applied is known as the "cold vswr." In general, the vswr increases as the electron beam current is increased. This "hot vswr" is a direct function of gain and can be attributed to reflections of the amplified signal at a discontinuity along the slow-wave structure (usually the lumped attenuator) and to a minor degree by the mere presence of the beam. The transfer of input signal energy to the helix is a function of the "cold vswr." Consequently, the noise figure of the 6861 is not increased by the "hot mismatch." The mechanism which causes the mismatch to change with beam current eventually causes the tube to break into oscillation at beam currents much higher than the rated value. The tube is designed for stability at the rated beam current with a short circuit of variable phase placed on the tube input. (The output impedance has but a small effect on stability.)

The tube has been successfully matched outside the frequency range of 2700 to 3500 mc, and has performed satisfactorily over a very broad frequency band. Tube noise figures of less than 5.3 db have been recorded for every region selected in the range from 2000 to 4300 mc as shown in Fig. 7. The limitation at this time on full-band operation (2000 to 4300 mc) is primarily due to considerations of matching and fixed-voltage operation.

Voltage Variation

The noise figure of the 6861 is relatively insensitive to voltage variations as shown in Fig. 8. The helix voltage is the most sensitive with respect to noise figure, while the collector voltage is least sensitive. The noise

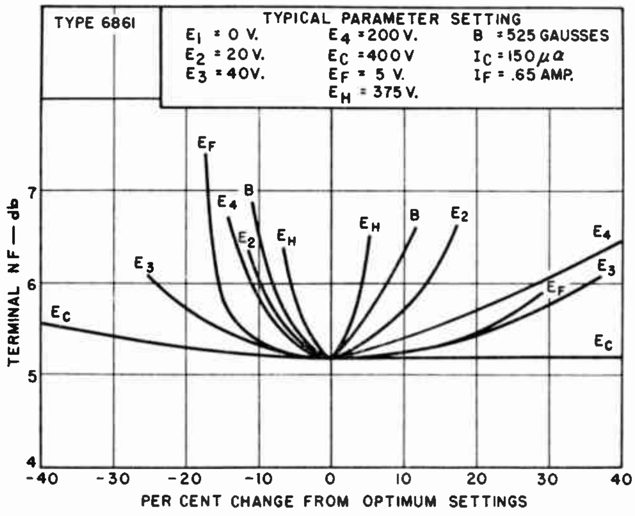


Fig. 8—Noise figure variation with parameter adjustment.

figure remains within 1 db of the minimum value for a change of 5 per cent in any one parameter. A change in grid-No. 2 or heater voltage changes the beam current, the noise figure, and the gain. Similarly, the gain drops 1.5 db for a 15-volt change in helix voltage. The noise figure is also dependent on the magnitude of the surrounding magnetic field. Because a solenoid's resistance increases as its temperature rises, the magnetic field intensity decreases with time for a fixed coil voltage. Consequently, a current regulator could be provided in the magnet-field power supply, depending upon system requirements.

In Fig. 9, the gain and noise figure are shown as functions of the grid-No. 2 voltage, with all other voltages remaining constant. The tube is well suited for AGC control because the noise figure remains within 1 db of the minimum value for gain changes of ± 4 db. The gain sensitivity for changes in grid-No. 2 voltage is approximately 1 db/volt.

Saturation Characteristics

The characteristics of the 6861 described thus far are those pertaining to low-level operation; *i.e.*, when the signal input is less than -35 dbm. At higher input signals, the power output begins to saturate and the gain drops off, as shown in Fig. 10(a). At a power input of -20 dbm, the gain is 5 db below the low-level value and the power output is maximized at 1 milliwatt in a typical tube. At still higher power inputs, the power output fluctuates under 1 milliwatt and the gain gradually decreases. Peak input powers as high as 250 watts with an average input power of 1 watt cause no degradation in performance. Furthermore, the output power does not exceed 1 milliwatt and hence the tube acts as a limiter for protection of sensitive succeeding stages (*e.g.*, mixers employing crystals). At high input-power values the tube actually becomes an attenuator having a value approaching the cold insertion loss.

A possible application of nonlinear operation is in frequency multiplication. As shown in Fig. 10(c), a signal

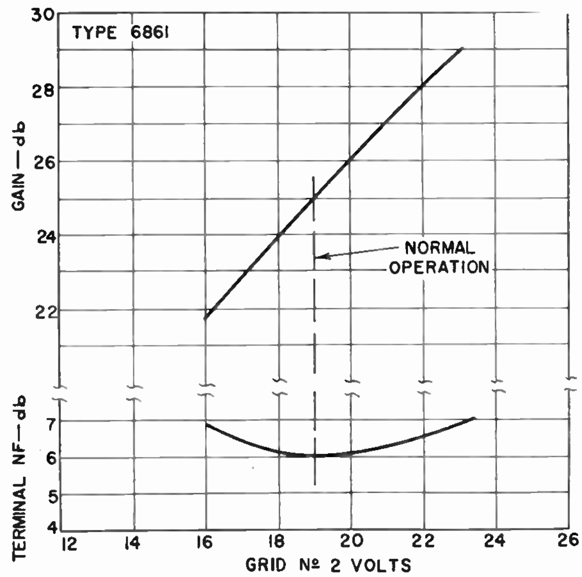


Fig. 9—Noise figure and gain vs E_2 voltage.

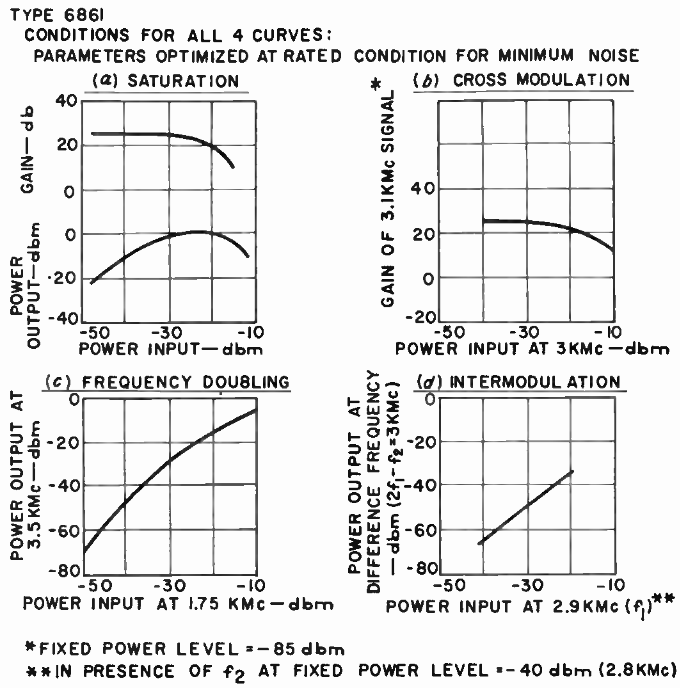


Fig. 10—Saturation characteristics.

below the operating band at 1750 mc was frequency doubled in the 6861 with 3-db gain at a certain optimum input. The conversion gain decreases around the optimum value of power input because lower powers result in less tube nonlinearity and higher levels result in reduced gain due to saturation effects.

Extraneous signals produce little cross-modulation as long as the cumulative effect of the input powers does not drive the tube into the nonlinear region. For instance, Fig. 10(b) indicates that a 3100-mc interfering signal has little effect on low-level operation until the interfering signal level approaches -30 dbm. Above this power input, the interfering signal causes a gradual reduction of the gain of the desired signal. Similarly, intermodulation distortion is small except near saturation, as shown in Fig. 10(d).

Phase Sensitivity

The output signal lags with respect to the input signal due to transit-time effects in the traveling-wave tube. This phase shift is constant until the input power approaches the saturation level; beyond this point, the relative phase shift increases with increasing power input. Ripple from the power supplies may also modulate the rf signal. Certain elements are more critical than others, as shown in Fig. 11. The voltages listed indicate the amount of rms voltage required to produce a peak-to-peak change in phase of 1° in a 3000-mc signal. In addition, measurements taken on three tubes show that the tube phase shift changes about 10° as the power input is raised from the noise level to the saturation level. The exact value is a function of the tube and operating frequency.

Tube Electrode	Typical Operating DC Volts	Approx. RMS Ripple Volts For Peak-to-Peak Phase Shift of 1°
Grid No. 1	0	0.1
Grid No. 2	20	0.1
Grid No. 3	40	0.5
Grid No. 4	200	3.5
Helix	375	0.024
Collector	400	6.7

Fig. 11—Table of phase characteristics.

Life

The life problem, as well as noise-figure reproducibility, was solved when the cathode chemistry and tube processing were worked out in detail. To date four of the life test tubes are being life-tested to end of life; one tube has passed 10,000 hours without any perceptible change (tube *C* in Fig. 3). Tube *B* was taken off life at 9200 hours since its gain and emission were fairly low—the noise figure rose only slightly with life (heater reoptimized to compensate for the reduced emission). Evaluation of this tube showed that the emissive area on the cathode had decreased to a small area about the cathode center—it is theorized that this tube was not processed properly. Tube *D* has shown a decrease in noise figure (from 6.3 to 5.8 db) in the first 3000 hours of life, as the emission level of its cathode gradually increases. (The emission level does not represent the available current but is merely a fraction of it.)

The characteristics of tube *A*, on the other hand, have remained perfectly constant through 6000 hours. All life tests are run above rated conditions to accelerate any life defects (*i.e.*, the beam current is set 50 μa and the cathode temperature 50° above the rated operating conditions). These curves indicate that the life of the 6861 is at least 10,000 hours; other considerations predict it will undoubtedly be much longer.

Acknowledgment

The authors extend their thanks to H. K. Jenny, P. R. Wakefield, W. Beam, and R. G. Talpey for their guidance; to A. Hogg, W. Johnson, and E. Stefanowicz for their major contributions to the success of the 6861; and to the many others associated with the design, fabrication, and testing of the low-noise tube.

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New Applications of Impedance Networks as Analog Computers for Electronic Space Charge and for Semiconductor Diffusion Problems*

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Summary—Starting from a general partial differential equation of the second order, which includes the equations of Laplace, of Poisson, and those of semiconductor diffusion problems, an equivalent equation with finite differences is discussed. This leads to resistance and to resistance-capacitance networks. Resistance chains are applied to high vacuum diodes of one-dimensional character (plane, circular, and spherical systems). Results are within about a per cent of the exact solutions. Resistance-capacitance chains and networks are applied to semiconductor diffusion problems with space and surface recombination, yielding known and new results pertaining to *p-n* diodes. A plane resistance network is applied to a triode with electronic space charge and yields the anode current vs grid voltage curve within a few per cents of the published value. These results may lead to considerable savings in the design of new tubes.

I. Introduction

In many physical and technical problems, solutions of the partial differential equations of Laplace, or of Poisson, are required, contingent on specified boundary conditions. In most cases, no exact solution is known, therefore, only numerical solutions are possible. Similar problems arise with the diffusion of carriers through semiconductors.

We start from the partial differential equation:

$$\frac{\partial^2 P}{\partial z^2} + \frac{\nu}{r}\frac{\partial P}{\partial r} + \frac{\partial^2 P}{\partial r^2} + f(P) = 0 \tag{1}$$

in two-dimensional problems with rectangular coordinates. We have $\nu = 0$ in a plane, and in problems of rotational symmetry round the z axis, we have $\nu = 1$. Transforming the differential equation (1) into an equation of finite differences and referring to Fig. 1,

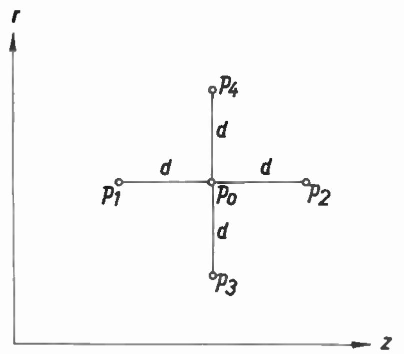

Fig. —Five adjacent maze points, spaced d, of a two-dimensional potential field of coordinates r and z.

$$\left\{\begin{aligned} P_1 &= P_0 - \frac{1}{1!}\frac{\partial P}{\partial z}d + \frac{1}{2!}\frac{\partial^2 P}{\partial z^2}d^2 - \frac{1}{3!}\frac{\partial^3 P}{\partial z^3}d^3 + \frac{1}{4!}\frac{\partial^4 P}{\partial z^4}d^4 - \cdots \\ P_2 &= P_0 + \frac{1}{1!}\frac{\partial P}{\partial z}d + \frac{1}{2!}\frac{\partial^2 P}{\partial z^2}d^2 + \frac{1}{3!}\frac{\partial^3 P}{\partial z^3}d^3 + \frac{1}{4!}\frac{\partial^4 P}{\partial z^4}d^4 + \cdots \\ P_3 &= P_0 - \frac{1}{1!}\frac{\partial P}{\partial r}d + \frac{1}{2!}\frac{\partial^2 P}{\partial r^2}d^2 - \frac{1}{3!}\frac{\partial^3 P}{\partial r^3}d^3 + \frac{1}{4!}\frac{\partial^4 P}{\partial r^4}d^4 - \cdots \\ P_4 &= P_0 + \frac{1}{1!}\frac{\partial P}{\partial r}d + \frac{1}{2!}\frac{\partial^2 P}{\partial r^2}d^2 + \frac{1}{3!}\frac{\partial^3 P}{\partial r^3}d^3 + \frac{1}{4!}\frac{\partial^4 P}{\partial r^4}d^4 + \cdots . \end{aligned}\right. \tag{2}$$

Hence:

$$\frac{\partial^2 P}{\partial z^2} + \frac{\nu}{r}\frac{\partial P}{\partial r} + \frac{\partial^2 P}{\partial r^2} + f(P) = \sum_{i=1}^{4}\frac{P_i - P_0}{d^2} + \frac{\nu}{r}\frac{P_4 - P_3}{2d} + f(P) + F(P),$$

$$F(P) = -\left(\frac{\nu}{r}\frac{1}{3!}\frac{\partial^3 P}{\partial r^3}d^2 + \frac{2}{4!}\frac{\partial^4 P}{\partial z^4}d^2 + \frac{2}{4!}\frac{\partial^4 P}{\partial r^4}d^2 + \cdots\right). \tag{3}$$

Replacing r by md, where d is the maze width, we obtain

$$\frac{P_1 - P_0}{d^2} + \frac{P_2 - P_0}{d^2} + \frac{P_3 - P_0}{d^2}\left(1 - \frac{\nu}{2m}\right) + \frac{P_4 - P_0}{d^2}\left(1 + \frac{\nu}{2m}\right) + f(P) + F(p) = 0. \tag{4}$$

Now consider the resistance network of Fig. 2. Application of Kirchhoff's law to the center point yields

$$\sum_{i=1}^{4}\frac{V_i - V_0}{R_i} + i_0 = 0. \tag{5}$$

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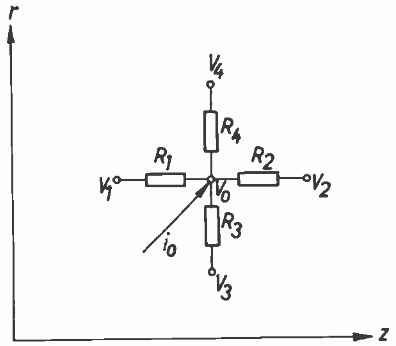

Fig. 2—Resistance network with resistances R_1, R_2, R_3, and R_4 between adjacent maze points. At the maze point of potential V_0 a source current i_0 is fed into the network.

The electric potentials V of the network points are proportional to the corresponding P values of the problem in hand

$$k_0P = V. \tag{6}$$

The resistances R_i have to be determined so that (4) is satisfied in each network point. Multiplication of (4) with $A(r, z)$ and identification of the corresponding parts of (4) and (5) yield the set of values (see Fig. 2):

$$R_1 = \frac{d^2}{A(r, z)}$$

$$R_2 = \frac{d^2}{A(r, z)}$$

$$R_3 = \frac{d^2}{A(r, z)\left(1 - \dfrac{\nu}{2m}\right)}$$

$$R_4 = \frac{d^2}{A(r, z)\left(1 + \dfrac{\nu}{2m}\right)}. \tag{7}$$

Furthermore, we obtain the source current, $i_0 = A(r, z) \times \cdot k_0f(P)$. Eq. (7) shows that all network resistances of Fig. 2, situated horizontally, are equal if $A(r, z)$ is independent of z. From the identity relation for the vertically situated resistances of Fig. 2 (R_3 at $m+1$ must be equal to R_4 at m) we obtain [1].

$$\frac{A_{m+1}}{A_m} = \frac{(m + 1)(2m + \nu)}{(2m + 2 - \nu)m}. \tag{8}$$

Eq. (8) yields,

$$\begin{aligned} &\text{if } \nu = 0, \quad A_m = A \text{ independent of } m, \\ &\text{if } \nu = 1, \quad A_m = mA. \end{aligned} \tag{9}$$

These relationships fully determine the resistances of the analog resistance network.

Special considerations are required at the *borders* of a network. If the border traverses the points 4, 0, 3 of Fig. 2, we may assume that the network is continued beyond this border. If we now replace R_3 and R_4, respectively, by two parallel resistances of values $2R_3$ and $2R_4$ and moreover replace i_0 by two currents of values $i_0/2$, the difference equation for P_0 remains satisfied. Now we may dissect the network along 3–0–4 and discard each one value $2R_3$, $2R_4$, and $i_0/2$. Then, beyond the border, no network exists and (4) is satisfied. Thus, in order to comply with a straight border, the border resistances should be doubled and the source currents at the border points should be halved.

By replacing a differential equation with an equation of finite differences, an error arises. This error is determined by the expression $F(P)$ of (3). The corresponding relative error is

$$F_{\text{rel}} = \frac{F(P)}{\nabla^2P + f(P)}. \tag{10}$$

This relative error is approximately proportional to the square of the maze distance d. If d is small, the number of mazes becomes large, thus requiring more resistances, contacts, etc. In each network point, the value of F_{rel} may be evaluated. By the application of a corrected source current $i_0' = k_0A_m(F(P) + f(P))$, this error F_{rel} may be very nearly eliminated.

A further error arises from inevitable tolerance values of the applied resistances. This error may be evaluated approximately and may be minimized by the application of properly adjusted resistances [2], [3].

II. Resistance Chains for the Solution of One-Dimensional Problems

In these cases, z may be dropped from (1) and (3). If $\nu=0$, we have a plane one-dimensional problem; if $\nu=1$ we have a problem of circular symmetry; and if $\nu=2$, we have a problem of spherical symmetry. As a case of practical importance, consider $\nu=1$. By (8) we obtain

$$\frac{A_m}{m} = \frac{A_{m+1}}{m + 1} = A.$$

Hence, by (7)

$$R_m = \frac{1}{mA\left(1 - \dfrac{1}{2m}\right)}; \quad R_{m+1} = \frac{1}{mA\left(1 + \dfrac{1}{2m}\right)}.$$

The value of m is equal to r/d. The first resistance of the chain is $R_1 = 2/A = R_0$. The subsequent resistances are, respectively,

$$R_0, \frac{R_0}{3}, \frac{R_0}{5}, \frac{R_0}{7}, \cdots \frac{R_0}{2m - 1}.$$

We now apply this resistance chain to the problem of space-charge distribution in a circular cylindrical high vacuum diode. In this case,

$$f(P) = \frac{\rho}{\epsilon_0} = \frac{s}{\epsilon_0\sqrt{2e_0}\sqrt{V}} = \frac{c}{\sqrt{V}}. \tag{11}$$

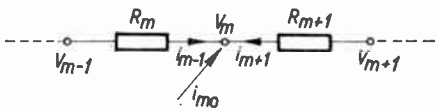

Fig. 3—Part of a resistance chain with resistances R_m and R_{m+1} between adjacent points and with a source current i_{mo} fed into the chain at the point of potential V_m.

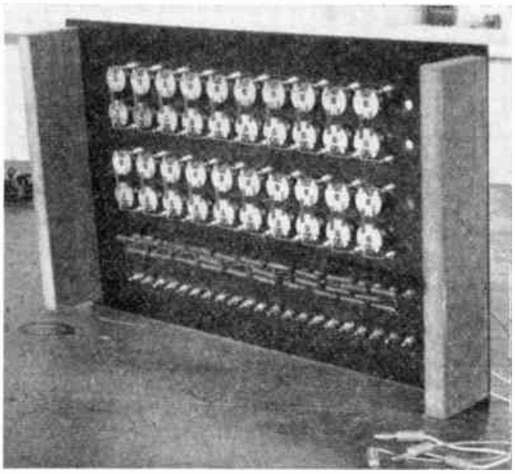

Fig. 4—Photograph of a resistance chain with source current units.

Here, ρ is the local space-charge density (coulomb per m^3), $\epsilon_0=(36\pi\ 10^9)^{-1}$ (coulomb)^2/m joule, s is the local current density (Amp/m^2), e_0 is the ratio of absolute charge to mass of an electron at rest ($1.76\ 10^{11}$ coulomb/kg), V is the local potential, and c is a constant. We have the relation, $sr=s_cr_c$, where r is the local radius and s_c is the current density at the cathode surface of radius r_c. Referring to Fig. 3, we obtain

$$i_{m0} = -\frac{2md^2}{R_0}\frac{\rho}{\epsilon_0} = -\frac{2d^2r_c}{\epsilon_0\sqrt{2e_0}R_0\,d}\frac{s_c}{\sqrt{V_m}}$$

$$= -\frac{c_1}{\sqrt{V_m}}. \tag{12}$$

The solution of this space-charge problem, as of most of the other problems to be solved here, is based on iteration. The first approximation for V_m is obtained by omitting space charge. The resulting V values are substituted in (12) and the corresponding source currents i_{m0} are fed into the network. Thereby, the second approximation for V_m is obtained, etc. After few iterations, the solution converges toward final V_m values, thus providing the approximate solution of the problem in hand. If the linear model dimensions are made n times larger, the relevant space charge is divided by n^2. Hence, the source currents also must be divided by n^2 in this case.

A practical network is shown in Fig. 4. The value of R_0 is 4.5 kilo-ohms and $nd=2$ cm. The currents are fed into the network by means of sources of very high internal resistance. Hereby, the individual currents have no influence on the adjacent current values. The starting velocity of electrons at the cathode (radius r_c) of the circular cylindrical high vacuum diode under consideration expressed in volts is 0.2 times the anode potential with respect to the cathode. The dimensions are $r_c=0.4$ mm, r_a (anode radius) $=2.4$ mm. In the model network r_c is represented by two mazes and the anode radius r_a by 12 mazes. The anode potential is 10 volts, assuming k_0 of (6) to be unity, and the cathode potential is 2 volts. The potential values without space charge are shown in Table I. The values V_{mod} obtained from the model network are in very good coincidence with the values V_{calc} obtained by exact calculations, the error being less than 1 per cent.

TABLE I

r/r_c	1.5	2	2.5	3	3.5	4	4.5	5	5.5
V_{calc}	2.265	3.87	5.115	6.13	7.0	7.73	8.40	8.98	9.52
V_{mod}	2.28	3.85	5.10	6.13	7.0	7.75	8.40	9.0	9.52

In applying iteration to the solution of the space-charge problem, errors may occur at the electrodes. Referring to Fig. 5, the resistance R corresponds to a maze width d. The source current i_{10} should represent the space charge symmetrically to point 1, the source current i_{20} symmetrically to point 2, etc., as indicated in Fig. 5 by the shaded areas. Obviously, the space charge adjacent to the electrode E is not accounted for by any model current. It may be taken into account by increasing i_{10} properly. In our case this increase is by about a factor 1.25. This correction factor may be determined experimentally in solving a specific space-charge problem by iteration and comparing the values at the model network with the calculated values. Instead of correcting i_{10}, we may alter the resistance adjacent to the electrode E, by inserting $R/2$ instead of R and by feeding i_{10} at a distance $d/2$ from E. This yields correct results only in cases in which V is approximately linear adjacent to E.

An approximate method of calculating the potential values of the problem in hand was published by Page and Adams [4]. This method is lengthy and intricate. The model network yields results in good agreement with calculated values after four iterations (see Fig. 6).

The iteration method was also successfully applied to a plane high vacuum diode with electrons, emitted by the cathode with a Maxwellian velocity distribution, as well as to further problems [1].

III. Resistance-Capacitance Chains

Here these chains are applied to semiconductor junction p-n diode diffusion problems. The differential equation for the hole density p in an n semiconductor is as follows [5]

$$\frac{\partial^2 p}{\partial z^2} + \frac{p_n - p}{L_p^2} + \frac{1}{D_p}\frac{\partial(p_n - p)}{\partial t} = 0. \tag{13}$$

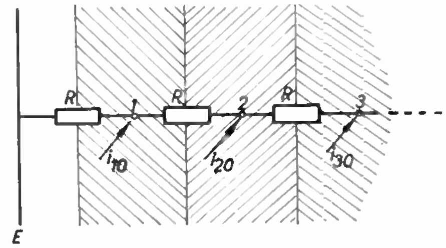

Fig. 5—Representation of the field, adjacent to an electrode E, by elements permitting an elementary corrective treatment of the source currents.

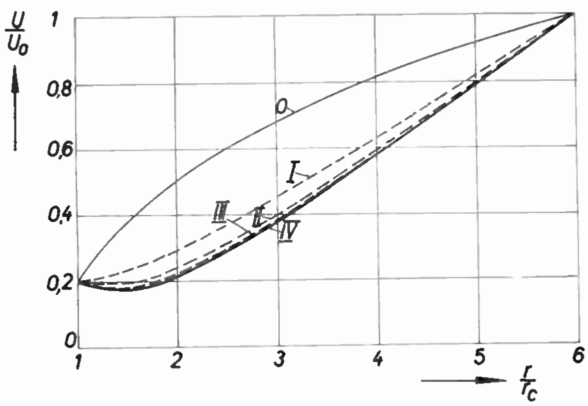

Fig. 6—Potential curves resulting from an iterative resistance chain solution of the space-charge problem connected with the flow of electrons in a high vacuum cylindrical diode. Curve marked O is without space charge. Curves marked I to IV are successive iterative solutions, the final one (IV) coincides with the calculated curve. Horizontal axis: ratio of radius r to cathode radius r_c.

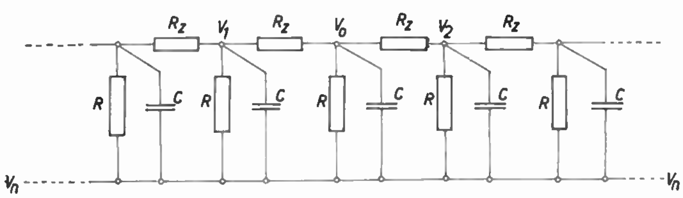

Fig. 7—Part of a resistance-capacitance chain.

We consider only the n semiconductor, starting from the end of the p-n depletion layer at $z=0$ to the diode termination at $z=l$, where there is an ohmic contact electrode. In (13), p_n is the value of p at $z=l$. Transformation of (13) into an equation of finite differences yields

$$\frac{(p_1 - p_0)k_0}{R_0} + \frac{(p_2 - p_0)k_0}{R_0} + \frac{(p_n - p_0)k_0 d^2}{R_0 L_p^2} + \frac{\partial}{\partial t}\frac{(p_n - p_0)k_0 d^2}{R_0 D_p} = 0. \quad (14)$$

Herein, L_p is the diffusion length of the holes in the n material and D_p their diffusion coefficient. At the point V_0 of the model chain we have, according to Kirchhoff's law (see Fig. 7),

$$\frac{V_1 - V_0}{R_z} + \frac{V_2 - V_0}{R_z} + \frac{V_n - V_0}{R} + C\frac{\partial(V_n - V_0)}{\partial t} = 0. \quad (15)$$

By a comparison of (14) and (15), we obtain, in Table II, the following analogy equations [6]–[8].

TABLE II

Semiconductor Diode		Model Chain
pk_0	=	V
R_0	=	R_z
$R_0 \frac{L_p^2}{d^2}$	=	R
$\frac{d^2}{R_0 D_p}$	=	C

At $z=0$, the depletion layer ends. The boundary condition is, if $U_1 \ll U_0$

$$p(z=0, t) = p_n \exp\left(\frac{eU_0}{kT}\right)\left[1 + \frac{eU_1}{kT}\exp(j\omega t)\right]. \quad (16)$$

In (16), the voltage at the diode's ohmic contact terminals is equal to $U_0 + U_1 \exp(j\omega t)$, $j = +\sqrt{-1}$, ω is the angular frequency, k is the Boltzmann constant ($1.38 \; 10^{-23}$ joule/°Kelvin), T is the temperature in °Kelvin, and e is the amount of electronic charge. According to Table II, we have $p(z=0, t) = V/k_0$ and $p_n = V_n/k_0$ (see Fig. 7). The hole density is p_n at the end of the diode; *i.e.*, at $z=l$. The differential equation (13) has to be solved under these two boundary conditions.

The electric current density s, caused by diffusion, is

$$s = -eD_p\frac{\partial p}{\partial z} \approx eD_p\frac{p_1 - p_0}{d} = \frac{eD_pR_0(V_1 - V_0)}{k_0 d R_z}. \quad (17)$$

From this equation, the diode characteristic; *i.e.*, the current as dependent on diode voltage, may be calculated. The assumed total diode length l is 450 microns.

Using the numerical values

$T = 300$°K, $D_p = 5 \; 10^{-3}$ m²/s, $L_p = 5 \; 10^{-4}$ m, $p_n = 10^{16}$ m^{-3}, $U_0 = 0.119$ volt, and $d = 50$ microns, we obtain

F_{rel} [relative error, see (10)] $= 0.085$ per cent, $k_0 = 10^{-16}$ m³ volt, $R_0 = R_z = 3506.1$ ohms ± 0.5 per cent, $R = 350.61$ kilo-ohms ± 0.5 per cent, $C = 142.8$ pF, $V(z=0) = 101$ volts, $V_n = p_n k_0 = 1$ volt.

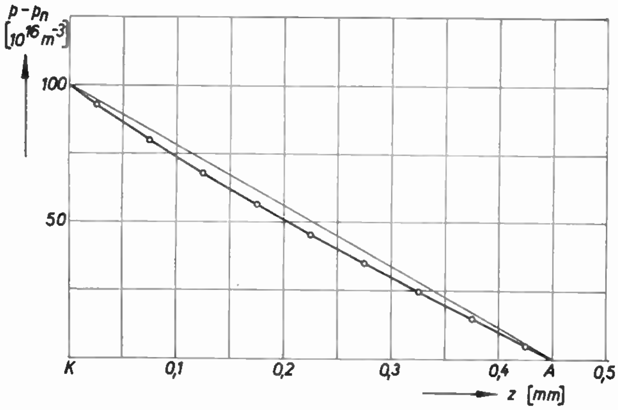

Fig. 8—Hole density (vertical scale) in the n material of a p-n junction diode as found from the chain of Fig. 7. At the left (K) is the border of the depletion layer and at the right (A) is the diode's electrode.

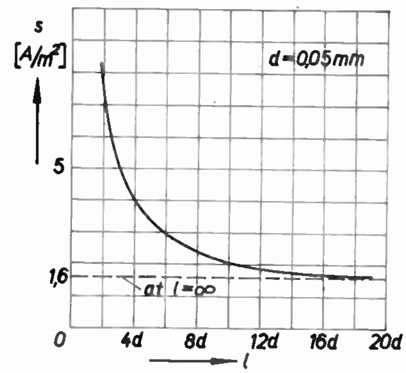

Fig. 9—Current density at the border of the depletion layer of a p-n junction diode (vertical scale) as dependent on the length l of the n material, expressed in maze distances d of the model chain.

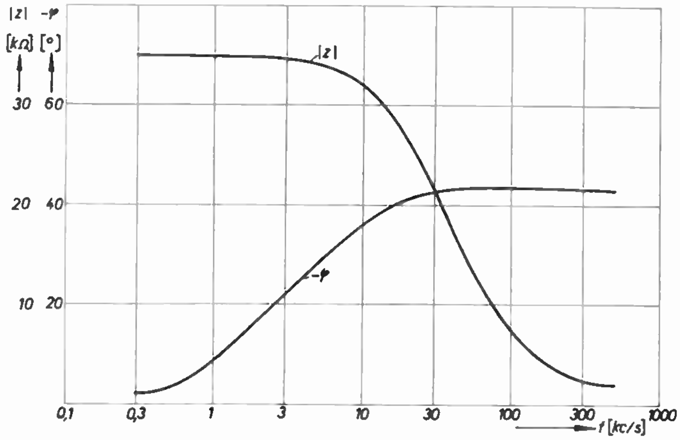

Fig. 10—Input impedance $Z \exp(j\phi)$ of a resistance-capacitance chain of Fig. 7 as dependent on frequency f.

The solution of (13), if $\partial/\partial t=0$ and $U_1=0$; *i.e.*, the stationary solution, is under the above boundary conditions

$$p(z) = p_n + \frac{p(z=0) - p_n}{[1 - \exp(-2l/L_p)]} \times [\exp(-z/L_p) - \exp(-2l/L_p)\exp(z/L_p)]. \quad (18)$$

The corresponding potential $V(z)$ as dependent on z is obtained by multiplying $p(z)$ by k_0. This potential $V(z)$ was measured on the model chain of Fig. 7, using the above values and short-circuiting the chain's end (Fig. 7). The result is shown in Fig. 8. In measuring $V(z)$ on the chain, a bridge circuit showing errors less than 0.1 per cent was used. The difference between measured and calculated values, according to (18), was less than 0.3 per cent. Doubling the maze width d led to errors less than 0.43 per cent.

From $p(z)$ of (18), the current density s may be evaluated by (17) under the above conditions. Again, the measured values on the chain were compared with values obtained by direct calculation from (18) and (17). The differences were less than about 2 per cent, this being the maximum error of the meters used for this purpose (see Fig. 9).

Finally, the alternating current density s_a may be evaluated by using (17) in the solution of (13), under nonstationary conditions, and applying the boundary conditions,

$$p(z=0) = p_n \exp\left(\frac{eU_0}{kT}\right)\frac{eU_1}{kT}\exp(j\omega t)$$

and $p(z=l)=0$. The solution is

$$s_a = \frac{e^2 p_n D_p \exp(eU_0/kT)}{L_p kT}\sqrt{1+j\omega\tau_p} \times \operatorname{cth}\left(\frac{l}{L_p}\sqrt{1+j\omega\tau_p}\right)U_1\exp(j\omega t). \quad (19)$$

Herein, $\tau_p = L_p^2/D_p$ is the lifetime of the minority p carriers in the n material of the diode. By multiplying s_a by the diode's cross-sectional area A, we obtain the alternating diode current i_a. Dividing i_a by $U_1 \exp(j\omega t)$ yields the diode's ac admittance $(1/Z)\exp(-j\phi)$. Again, this admittance may be obtained by measuring the admittance value at the input end of the chain of Fig. 7, short-circuiting its far end. The results of the chain measurements are shown in Fig. 10. These results coincide within 2 per cent with the values found from (19). These 2 per cent constitute the maximum error of the measuring instruments. Finally, the square root of the mean square value of s_a according to (19) is shown as dependent on frequency in Fig. 11 according to measurements on the chain. Here again, coincidence with calculated values is within 2 per cent.

IV. Resistance Networks Applied to Two-Dimensional Electronic Space-Charge Problems

Now we investigate the space charge in a high vacuum triode, which may be approximately represented by a plane electrode system. The dimensions and symbols,

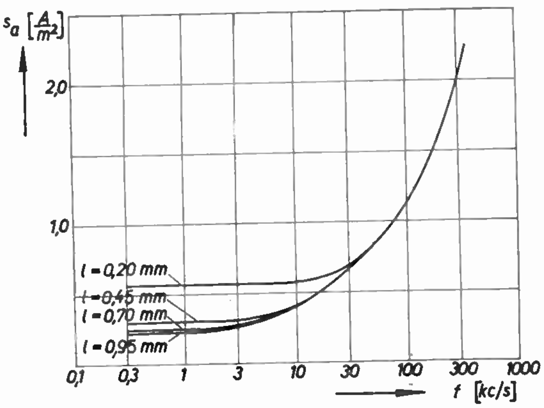

Fig. 11—Alternating current density s_a at the border of the depletion layer of the *p-n* junction diode of Fig. 7 as dependent on the frequency f at several lengths l of the n material of the diode.

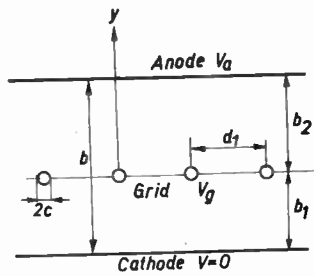

Fig. 12—Nomenclature of plane triode system.

attached to the plane triode system, are shown in Fig. 12. As a special triode system, the European tube type EL6 was selected [9]. This is a pentode with an oval cathode and an oval first grid wound of wires approximately parallel to the greater part of the surface of the cathode. The two supporting rods of the first grid are adjacent to the curved ends of the oval cross section of the cathode. Therefore, by the negative voltage of the rods, these curved ends of the cross section of the cathode will not partake in the emission of the cathode. Only the flat parts of the cathode are actively emitting electrons. Under these conditions, this central part, together with the parallel grid wires, may be approximately represented by a plane system according to Fig. 12. The equivalent dimensions of this plane triode system are (see Fig. 12) $2c=0.06$ mm, $b_1=d_1=0.6$ mm, $b_2=1.2$ mm, $b=b_1+b_2=1.8$ mm. The actual pentode system is replaced by a triode, the anode of which approximately has the screen grid's situation. In the model, the dimension b_2 is represented by 16 maze widths d, b by $24d$, $2c$ by $0.8d$. The linear magnification of triode to model is 266.7. By this large magnification, the error according to (10) is practically insignificant.

Special attention is given to the representation of the grid wires in the model. As $c=0.4d$, the distance between the wire surface and the adjacent next maze point is $0.6d$. Referring to Fig. 13, it is obvious that the grid

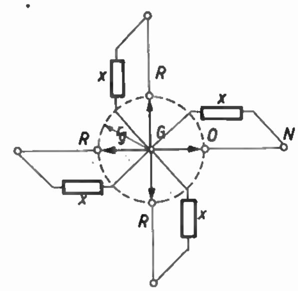

Fig. 13—Representation of a grid wire in the resistance network model of a plane triode. The wire radius is r_g and is equal to $0.4d$, where d is the maze distance, equal to GN in the figure. By connecting a suitable parallel resistance X to the network resistance R, the wire surface may be simulated.

wire surface (radius r_g round G) may be represented by connecting four extra resistances X in parallel to the network resistances R. If $X=R(0.6/0.4)=1.5R$, the "distance" ON of Fig. 13 apparently will be $0.6d$, where d is the distance GN. In this way, small wire radii may be adequately represented in the model [10] [11].

In the model network, the resistances R between adjacent maze points are 3.5 kilo-ohms with a tolerance of ± 0.3 per cent. At the model boundaries, the resistances are corrected according to Section I.

The electrostatic field was calculated according to a method published by M. J. O. Strutt [13]. Hereby, equipotential curves are found, which coincide almost exactly with the grid wire circumferences, whereas most other formulas do not comply with this requirement.

In the present case, the function $f(P)$ of (1) is given by

$$f(P) = \frac{\rho}{\epsilon_0} = \frac{s}{\epsilon_0\sqrt{2e_0}\sqrt{P}}\,. \tag{20}$$

Here, s is the local current density and P is the local potential. The corresponding source current of (5) is

$$i_0 = \frac{d^2}{\epsilon_0 R}\,\rho = \frac{d^2}{\epsilon_0 R}\,\frac{s}{\sqrt{2e_0}\sqrt{V}}\,. \tag{21}$$

In (21) the resistance R is the uniform value of R_i of (5) in the present case. The initial difficulty in the triode case is that the local current density s is *not known*. Hence, it would seem, that (21) cannot be applied and that the iteration process cannot be carried out.

This difficulty however, was overcome by some preliminary experiments. In these, a cylindrical diode system was represented properly, including boundary corrections, on the plane resistance model network. The case of zero cathode emission velocity and zero electric field strength at the cathode was considered, the theoretical solution of this case being well known. The local current density is variable. Instead of this true *variable* current density, a *constant* current density s was as-

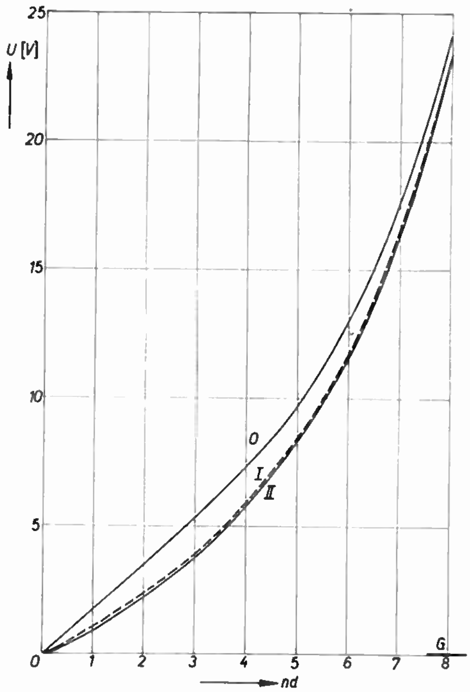

Fig. 14—Potential distribution obtained from network model measurements for a plane triode system. Vertical scale: potential between cathode and grid along a line perpendicular to the cathode through the center point between two adjacent grid wires. Horizontal scale: distance from cathode (left) expressed in maze distances d. Curve marked O: without space charge. Curves marked I and II: first and second (final) approximation of potential with full electronic space charge.

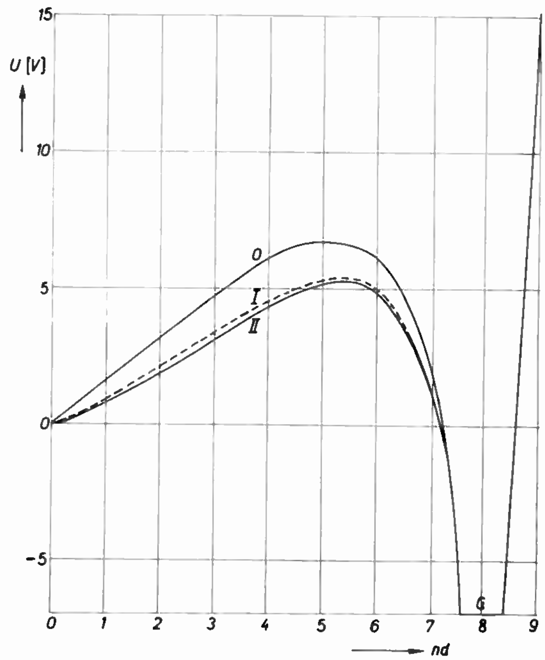

Fig. 15—Similar to Fig. 14, except for a line perpendicular to the cathode through the center of a grid wire.

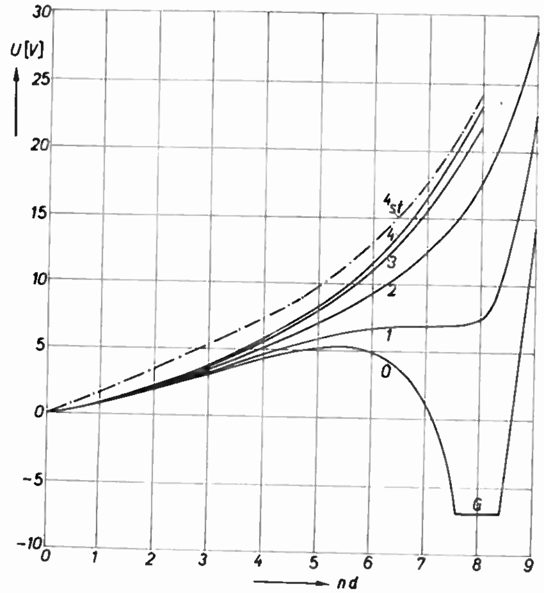

Fig. 16—Potential curves for a plane triode system with full electronic space charge (curves 0 to 4) and without space charge (curve 4_{st}). The curves are along lines perpendicular to the cathode and (curve 0) through the center of a grid wire; (curve 1), through a point between adjacent grid wire centers, distant d from wire center; (curve 2), distant 2 d from wire center; (curve 3), distant 3 d from wire center; and (curve 4), distant 4 d from wire center. The curve 4_{st} is equivalent to 4 without space charge.

sumed in the entire diode field. The constant value of s was chosen in between the known exact values of s at the cathode and at the anode. The iteration process of Section I and Section II was then applied, using the assumed constant current density in (21), which remained unaltered throughout the iterations. After two successive iterations, a resulting potential field was found, which coincided well within 1 per cent with the theoretically known exact field [10], [11].

From these preliminary experiments, it was concluded that any reasonable constant current density may be used, instead of the exact local current density, in carrying through the iteration process and still yield a correct final potential field. This important result provides the key to our triode problem. The anode current density of the triode system is 32.7 ma/cm². Hereby, in our model, we obtain $i_0 \approx 100/\sqrt{V}$. Carrying out the iteration process with source currents, calculated by this equation, yields the potential distribution, shown in Fig. 14 through Fig. 17, including space charge [10] [11].

From these potential distributions, and especially from Fig. 17, the triode's characteristic curve, anode current vs grid volts at constant anode volts, may be evaluated. From Fig. 17, it is seen that the curve U_{d1}, representing the potential, including space charge at a distance d from the cathode (one maze distance), is very nearly parallel to the one volt line. The *mean* potential value, obtained from the U_{d1}-curve is indicated by the dashed $\overline{V}_{kd}$-line in Fig. 17. This mean value $\overline{V}_{kd}$ is now used in calculating the triode characteristic.

Prior to this determination, the situation of the potential minimum, which is adjacent to the cathode sur-

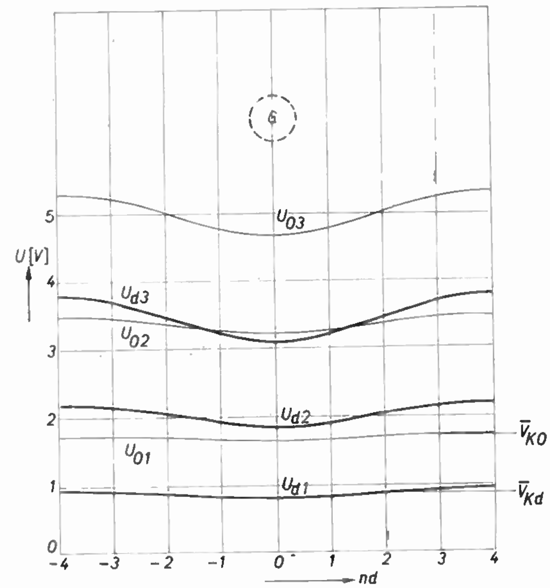

Fig. 17—Potential distribution in a plane triode system including full electronic space charge along lines paralled to cathode surface. G marks the grid wire. The curves marked U_{d1}, U_{d2}, U_{d3} give the potentials at a distance d, $2d$ and $3d$ from the cathode surface, including full space charge. The curves marked U_{01}, U_{02}, U_{03} give the corresponding potentials without space charge. The dashed line marked $\overline{V}_{kd}$ is the mean value of U_{d1} and $\overline{V}_{k0}$ is the mean value of U_{01}.

face in the actual triode, should be considered. A simple calculation shows that in our triode, at the above anode current density, the potential minimum is at a distance of approximately 2.6 10^{-2} mm from the cathode surface, whereas the grid wire centers are at a distance of 0.6 mm from the cathode surface. Under these conditions, the triode characteristic may be approximately calculated, assuming electron emission at zero emission velocity and zero electric field strength at the cathode.

By this result, we obtain approximately

$$s_a = 2.34\ 10^{-6} \frac{\overline{V_{kd}}^{3/2}}{d^2} \tag{22}$$

where s_a is the anode current density. From Fig. 17 we obtain $s_a = 33.3$ ma/cm^2 and an anode current $I_a = 81.5$ ma at a grid voltage of -7 v and an anode voltage of 250 v. Exactly similar calculations were carried out at the same anode voltage and at grid voltages of -5 and -10 v, yielding anode currents of 117.5 and 39.0 ma. These values are in very satisfactory agreement with published values by tube manufacturers. The tube's transconductance also may be found in this way. Again the values obtained from model measurements are in very satisfactory agreement with published values by the tube manufacturers. The differences are below 2 per cent.

The above method of obtaining a triode's characteristic curve and values from model measurements is the first one, as far as the authors are aware, that yields

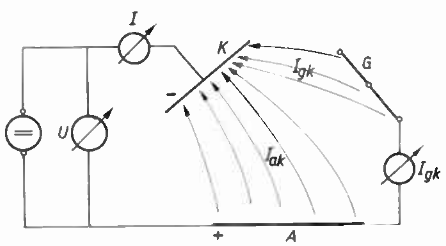

Fig. 18—Shows connections for measuring the capacitance values of the triode, including space charge, using the resistance network model. K is the cathode, G is the grid, and A is the anode.

entirely satisfactory results in cases akin to the EL6 tube. Other methods, based on formulas derived from the electrostatic fields and corrected for space charge, are in error for as much as 50 per cent in similar cases, as was shown by the authors [11].

This network model also can be applied in the determination of interelectrode capacitances including electronic space charge. The procedure is shown in Fig. 18. To the complete model, with full source currents simulating space charge, according to the final iteration, a dc or ac voltage U is applied between the cathode K and the anode A (see Fig. 18). Then the currents I and I_{gk} are measured (see Fig. 18). Obviously, $I = I_{ak} + I_{gk}$. Hence, $I_{ak} = I - I_{gk}$. Now the three triode capacitances are given by

$$C_{gk} = k_c I_{gk}/U,$$
$$C_{ak} = k_c I_{ak}/U,$$
$$C_{ga} = k_c I_{ga}/U.$$

Here, the factor k_c depends on the model dimensions and its determination is shown presently. In order to determine I_{ag}, we must interchange the electrodes K and G of Fig. 18. Thereby, the current I becomes equal to $I_{gk} + I_{ag}$ and hence, $I_{ag} = I - I_{gk}$ in this case. By carrying out these measurements without space charge, that is, with zero source currents at the model, we obtain the equivalents of the *electrostatic* interelectrode capacitances of the triode. These are either given by the tube manufacturer or else may be easily measured at the triode. Thus the factor k_c may be determined. The error caused by the grid supports, etc., is assumed to be insignificant. Furthermore, the ratio of the space charge to the electrostatic capacitances is found directly from the model measurements. In our case, for C_{kg} the ratio is found to be 1.20 and for C_{ga} it is 1.08. In the measured values of C_{ga} and C_{ka} the fact that electrons land on the anode is not accounted for.

V. Resistance-Capacitance Networks Applied to Semiconductor Junction Diodes

Most semiconductor junction diodes show circular cylindrical symmetry. Hitherto, the theoretical treatment refers to one-dimensional cases. Here the case of

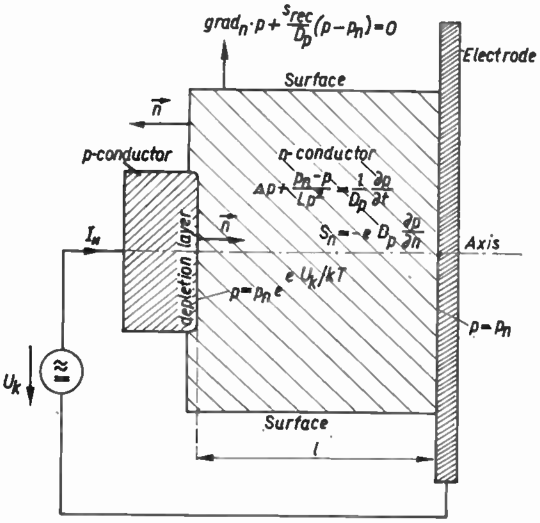

Fig. 19—Model of a p-n junction diode of circular cylindrical symmetry. The equations are derived from the laws of diffusion and of recombination of holes in the n conductor and on its surface. Symbols: p hole density, p_n hole density at the electrode, n direction normal to n-conductor surface, U_k voltage across diode, s_n current density normal to n-conductor surface, L_p diffusion path of holes, s_{rec} recombination velocity of holes, e the amount of electronic charge, and k Boltzmann's constant.

circular cylindrical symmetry is treated by means of a suitable model network.

The physical relationships of a p-n junction diode are shown in Fig. 19. Parts of the resistance-capacitance model network are shown in Fig. 20 and Fig. 21. Referring to (14), we derive Table III, showing equivalent quantities, physical and model (left and right, respectively). Table III is different for points on the axis of rotational symmetry and for points outside this axis. In Table III R_0 indicates the same quantity as in Table II, while d is the maze width of (14), and D_p the diffusion-coefficient of the holes in the n material.

At the junction diode's free surface (Fig. 19) the relationship [12]

$$\frac{\partial p}{\partial n} + \frac{s_{rec}}{D_p}(p - p_n) = 0$$

or

$$R_{0r} = R_0 \frac{D_p}{s_{rec} r}$$

expresses the recombination (here n is the direction normal outward to the surface). The recombination velocity at the surface is s_{rec} and r is the corresponding radius from the axis. At the depletion layer boundary, according to Fig. 19, $p = p_n \exp(eU_k/kT)$, where U_k is the voltage at the diode's terminals. At the diode electrode to the right in Fig. 19, we have $p = p_n$.

With these boundary conditions, the stationary hole flow in the junction diode was investigated, taking into account several recombination conditions (see Fig. 22).

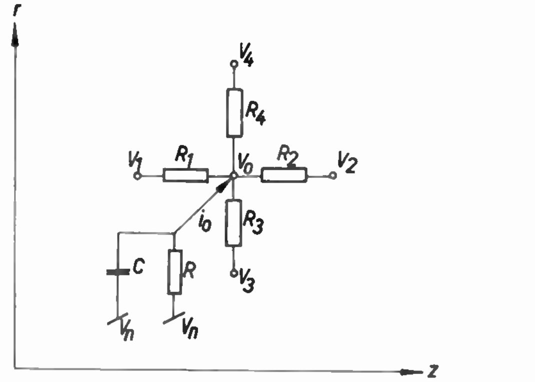

Fig. 20—Resistances and capacitance for a maze point outside the z axis of the resistance-capacitance network.

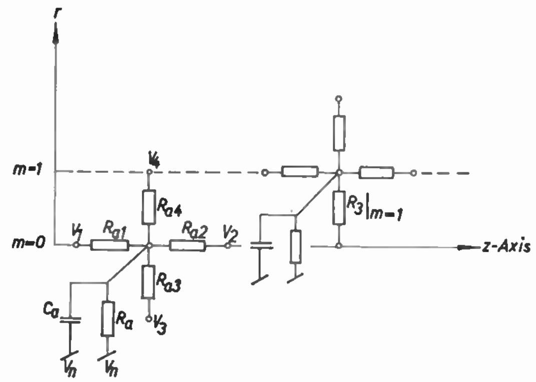

Fig. 21—Similar to Fig. 20, except for a maze point on the z axis.

TABLE III

Outside Axis	On Axis
$pk_0 = V$	$pk_0 = V$
$R_0 \frac{1}{m} = R_1$	$4R_0 = R_{a1}$
$R_0 \frac{1}{m} = R_2$	$4R_0 = R_{a2}$
$R_0 \frac{1}{m-1/2} = R_3$	$2R_0 = R_{a3}$
$R_0 \frac{1}{m+1/2} = R_4$	$2R_0 = R_{a4}$
$R_0 \frac{L_p^2}{md^2} = R$	$4R_0 \frac{L_p^2}{d^2} = R_a$
$\frac{md^2}{R_0 D_p} = C$	$\frac{d^2}{4R_0 D_p} = C_a$

VI. Acknowledgment

The work described in this paper was made possible by grants from the Batelle Memorial Institute at Geneva, from the Swiss Federation, and from Hasler Werke, Berne. The authors wish to express their thanks for these grants.

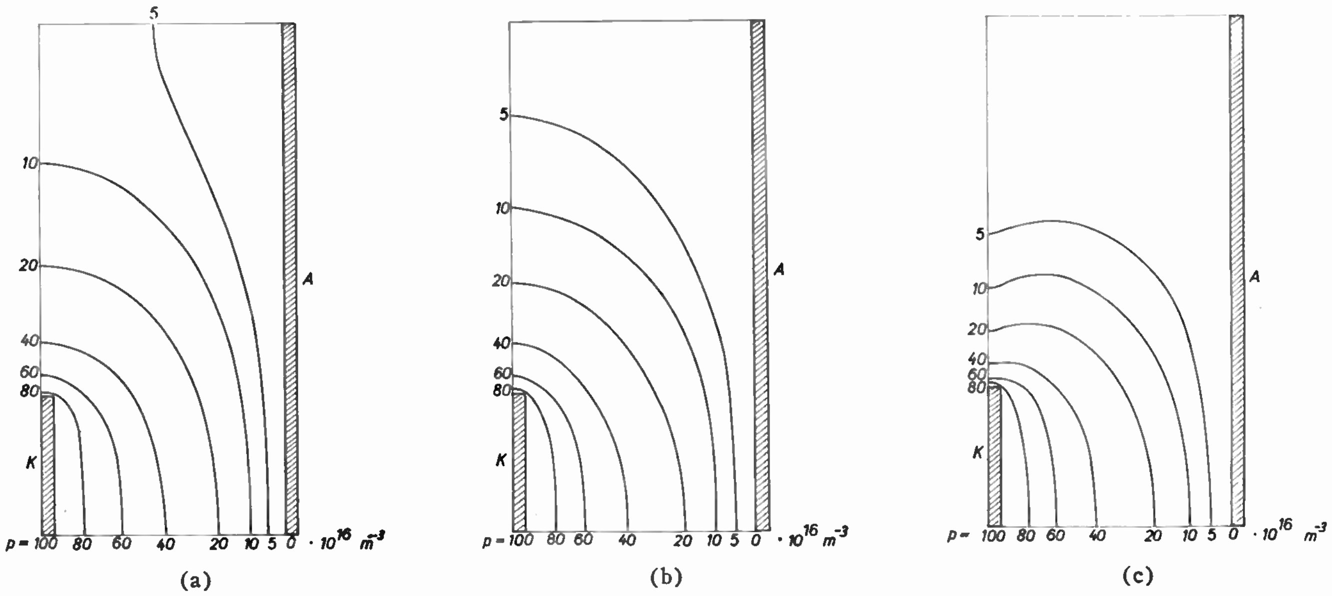

Fig. 22—Hole densities, obtained from measurements on the model network: (a) no space and no surface recombination; (b) only surface recombination; (c) surface as well as space recombination. To the left is the depletion layer K, to the right is the ohmic electrode A on the n-conductor surface.

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IRE Standards on Solid-State Devices: Methods of Testing Point-Contact Transistors for Large-Signal Applications, 1958*

58 IRE 28.S1

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1.0 General

THIS standard recommends and describes methods of measurement of the important characteristics of transistors in large-signal applications. Power amplifiers, pulse amplifiers, sinusoidal oscillators, multivibrator-type switches, and regenerative pulse generators are examples of this type of application.

Large-signal applications involve excursions of the operating point over large ranges of the transistor characteristics. Often, the operating point may move from a low-current cut-off region, through an essentially linear region, to a high-current saturation region. Since these two end regions correspond to the transistor being OFF or ON, respectively, the transistor can be considered as a switch.

Transistors generally fall into two classifications: 1) devices whose short-circuit forward-current transfer ratio α_{fb} is greater than unity, and 2) devices whose short-circuit forward-current transfer ratio α_{fb} is less than unity. Point-contact and point-junction transistors generally fit into the first classification, and the junction transistor into the latter. This standard considers the methods of test for point-contact transistors only.

Both point-contact and junction transistors are used in large-signal applications. In the case of point-contact transistors it is difficult to express the large-signal behavior in an analytical form which applies over the entire range of operations. This standard is based on piecewise linear approximations of the characteristics. The procedures for measuring the parameters of these approximations are specified.

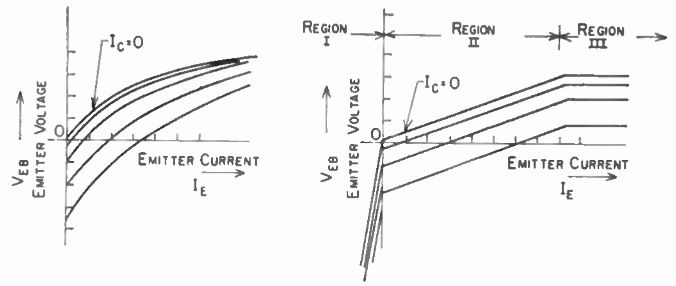
In the case of junction transistors, analytical expressions are available which are valid over the entire range of operation; however, it is again convenient to divide the characteristics into three regions for purposes of understanding the transistor behavior and making measurements. The differences between point-contact and junction transistors make it convenient to separate the methods of test.

2.0 Methods of Testing Point-Contact Transistors

In nonlinear applications, the device behavior must be described in all regions of operation. Pulse or switching operation is a special case of large-signal operation, and many of the tests described in this section apply directly to other large-signal applications such as Class B or Class C amplifiers.

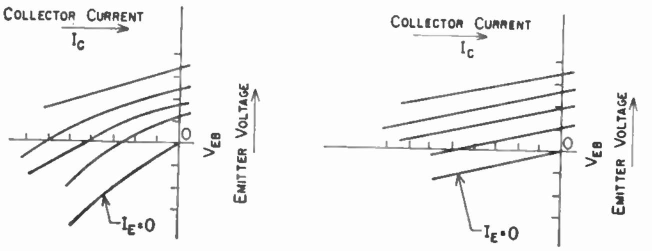
In small-signal applications, such as linear amplifiers, the transistor can be represented by a two-port network whose input is described by voltage v_i and current i_i and whose output is described by voltage v_0 and current i_0. There are six ways[1] of specifying this network in terms of v_i, i_i, v_0,, and i_0. The open-circuit impedance representation is most convenient, because point-contact transistors designed for switching applications are potentially unstable when short-circuit admittance measurements are attempted. A specific method of notation involving the letter subscript is used in this standard, consistent with "IRE Standards on Letter Symbols for Semiconductor Devices."[2]

[1] E. A. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. 2, pp. 133–151; 1935.
[2] "IRE Standards on Letter Symbols for Semiconductor Devices, 1956" (56 IRE 28.S1), Proc. IRE, vol. 44, pp. 934–937; July, 1956.



(a) (b)

Fig. 1—V_{EB} vs I_E with I_C as a parameter. (a) Typical input characteristic. (b) Idealized input characteristic and regional division.



(a) (b)

Fig. 2—V_{EB} vs I_E as a parameter. (a) Typical reverse-transfer characteristic. (b) Idealized reverse-transfer characteristic.

Four open-circuit impedance parameters can be measured,[3–5] as suggested by (1) and (2) that describe the point-contact transistor in small-signal applications.

$$v_i = z_i i_i + z_r i_0 \tag{1}$$

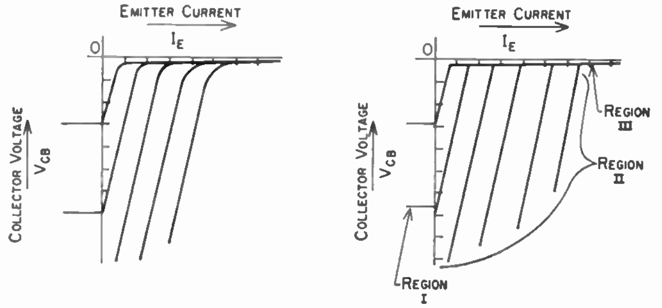
$$v_0 = z_f i_i + z_0 i_0. \tag{2}$$

In Figs. 1 through 4 are shown both typical and idealized input, output, and transfer static characteristics of point-contact transistors. Methods for obtaining static characteristics are covered in Section 2.0 of "IRE Standards on Solid-State Devices: Methods of Testing Transistors."[5] These characteristics can be divided into three regions[6] as follows:

Region I—a region in which both the emitter and collector junctions are biased in the reverse, or high-resistance, direction. Considered as a switch the transistor is in the OFF condition. This region may be designated "collector-current cutoff."

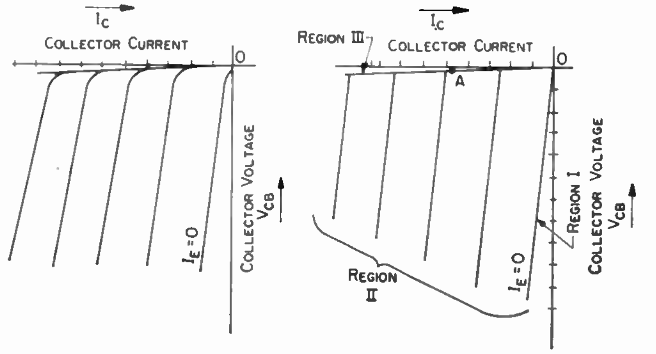
Region II—a region in which the emitter junction is biased in the forward direction and the collector junction is biased in the reverse direction. This region is usually called the "active region."

Region III—a region in which the emitter junction is biased in the forward direction and the collector current is limited primarily by the external collector circuit. Considered as a switch the transistor is in the ON condition.



(a) (b)

Fig. 3—V_{CB} vs I_E with I_C as a parameter. (a) Typical forward-transfer characteristic. (b) Idealized forward-transfer characteristic.



(a) (b)

Fig. 4—V_{CB} vs I_C with I_E as a parameter. (a) Typical output characteristic. (b) Idealized output characteristic.

Nonlinear effects complicate the analysis of circuits in which the operating point of the transistor traverses two or more of the above regions. To a first approximation, however, the switching or pulse behavior can be determined using the idealized characteristics shown in Figs. 1 through 4 in which linearity is assumed within each region. A set of dynamic and static parameters may be chosen that will describe the behavior of the point-contact transistor in all three regions. In circuit applications the design of transistor switching circuits can be facilitated by using negative-resistance characteristics.

Measurement of the following static, dc, and dynamic parameters may provide the information required for most large-signal applications. This listing may be expanded or abbreviated to conform with the requirements of any specific application.

Static Measurements

1) V_{EB} vs I_E curves with I_C as a parameter.
2) V_{EB} vs I_C curves with I_E as a parameter.
3) V_{CB} vs I_E curves with I_C as a parameter.
4) V_{CB} vs I_C curves with I_E as a parameter.

DC Measurements

5) $r_E(I_C^*, V_{EB}^*)$, the reverse resistance of the cutoff emitter with $I_C=0$.

Note: The symbology used herein, *i.e.*, $r_E(I_C^*,$

[3] R. F. Shea, *et al.*, "Principles of Transistor Circuits," John Wiley and Sons, Inc., New York, N. Y.; 1953.

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V_{EB}^*), denotes that the parameter r_E is measured for specified, constant bias values I_C and V_{EB}; hence r_E is a constant. On data sheets and test specifications for a given transistor wherein these bias conditions are fixed, the notation would reduce to the form $r_E(0, -20)$, referring to the constant bias conditions of measurement of $I_C^*=0$ ma, and $V_{EB}^*=-20$ v.

6a) $r_C(I_E^*, V_{CB}^*)$, the reverse resistance of the cut-off collector with $I_E=0$.

6b) $I_C(I_E^*, V_{CB}^*)$, the collector current with $I_E=0$.

Note:

$$I_C(0, V_{CB}^*) = \frac{V_{CB}^*}{V_C(0, V_{CB}^*)}.$$

This quantity is often called I_{CO}.

7) $I_C(I_E, V_{CB})$, for a specified increase in I_E at a constant V_{CB}, determines the dc current transfer ratio, α_{FB}.

Note:

$$\alpha_{FB} = \frac{I_C(I_E, V_{CB}) - I_C(0, V_{CB})}{I_E}.$$

8) $r_B(I_E^*, V_{CB}^*)$, the base resistance with $I_E=0$.

9) $V_{CB}(I_E^*, I_C^*)$, the voltage between collector and base at a specified I_E and I_C.

10) $V_{CE}(I_E^*, I_C^*)$, the voltage between collector and emitter at a specified I_E and I_C.

Dynamic Measurements

11) α_{fb}-vs-I_E curves with V_{CB} as a parameter.

12) Frequency response.

(a) α_{fb} vs frequency.

(b) Square-wave response

Rise time, t_r

Storage time, t_s

Fall time, t_f

Propagation delay time, t_d.

2.1 Region I DC Measurements

In Region I, both the input terminal pair and the output terminal pair are biased in the reverse direction. Consequently, relatively small currents are flowing in the emitter and collector circuits, and the transistor functions as a passive network. Since the collector current is essentially cut off, the device is analogous to a switch in the OFF condition. In this condition, the slope of the input static characteristic ($r_i=r_e+r_b$) is large compared to the slopes encountered in Regions II and III, and is essentially equivalent to the reverse resistance of a simple diode. Similarly, the slope of the output static characteristic ($r_0=r_c+r_b$) is essentially equivalent to the reverse resistance of a diode.

2.1.1 Reverse Resistance Measurements: For pulse applications, it is common practice to specify the dc voltage-to-current ratios at some specified operating point, rather than the small-signal r_i and r_0 in Region I.

$$r_E(I_C^*, V_{EB}^*) \equiv \frac{V_{EB}}{I_E} \text{ with } I_C = 0 \text{ and } V_{EB} \text{ specified} \quad (3)$$

$$r_C(I_E^*, V_{CB}^*) \equiv \frac{V_{CB}}{I_C} \text{ with } I_E = 0 \text{ and } V_{CB} \text{ specified.} \quad (4)$$

The relatively small effect of the Region I base resistance r_B is included in r_E and r_C in the above measurements. The measurement of r_E and r_C may be used to determine acceptability for a given application.

The reverse resistance r_i and r_0 also can be measured by oscillographic methods. These methods of test have the advantage of exhibiting any irregularities in the curves occurring in any specified operating range.

An alternative method of characterizing the collector is to measure the reverse collector current at a specified collector voltage. In this test, the emitter is open-circuited, $I_E=0$. The current $I_C(0, V_{CB}^*)$ is of particular interest in large-signal applications.

2.1.2 Base-Resistance Measurement: The open-circuit reverse transfer resistance, r_r, measured by small-signal methods in the active region cannot be used to determine the peak turning point. (The peak turning point is defined in Section 2.5.3.2.) For this purpose, the base resistance $r_B(I_E^*, V_{CB}^*)$ may be measured at $I_E=0$. This value of $r_B(0, V_{CB}^*)$ plus any external resistance R_B is multiplied by the reverse collector current $I_C(0, V_{CB}^*)$ to determine the approximate peak point. This peak point marks the boundary between Region I and II and influences the selection of trigger and bias magnitudes. A direct method of measuring the peak point is given in Section 2.5.3.3.

The base resistance $r_B(0, V_{CB}^*)$ may be determined at a specified value of V_{CB} by open-circuiting the emitter and measuring the emitter-to-base voltage V_{EB} with a sensitive dc electronic voltmeter. The base resistance may then be calculated by

$$r_B(0, V_{CB}^*) \equiv \frac{V_{EB}}{I_C}. \quad (5)$$

2.1.3 Precautions: In making dc measurements of the reverse resistance, the collector current, and base resistance in Region I, the following factors must be considered:

1) Care must be taken not to exceed maximum allowable current, voltage, and power ratings.
2) Under collector reverse-bias conditions, leakage currents between emitter and collector may be observed by an increase in emitter-to-base potential. This leakage current may adversely affect the operation of some circuits and invalidate (5).
3) The measured parameters are temperature-sensitive.

2.2 Region II DC Measurements

The low-frequency behavior of point-contact transistors in the active region can be determined from a set of static characteristics or alternatively by properly chosen open-circuit impedance measurements. For pulse characterization, particular attention must be paid to the boundaries of the active region.

2.2.1 Static Characteristics: The static characteristics of point-contact transistors are of the class which have in most cases a current as the quantity which is held constant. Pulse methods of obtaining static characteristics are particularly useful and are recommended for determining the large-signal behavior of point-contact transistors for two major reasons. The active behavior must be known in regions of operation where the maximum power dissipation of the device may be exceeded. In addition, thermal effects on temperature-sensitive parameters are minimized using pulse techniques particularly if sufficiently long repetition periods are employed as compared to the fundamental thermal time constant. For methods of measurement see Section 2.1 of "IRE Standards on Solid-State Devices: Methods of Testing Transistors."[5]

Precautions: In obtaining the static characteristics of point-contact transistors, it is sometimes found that irregularities and/or oscillations are exhibited. These anomalies arise from the device being open-circuit-unstable in some regions and short-circuit-unstable in others. If the exact transistor behavior in these regions is to be determined, methods must be devised to stabilize the device. In general, by providing the proper terminations, it is possible to suppress the oscillations and to trace out the static characteristics over the regions of interest.

2.2.2 Measurement of Open-Circuit Impedances: From the four static characteristic families, the low-frequency values of the two-port open-circuit impedances (r_i, r_r, r_f, and r_o) can be obtained as discussed in Section 3.5 of "IRE Standards on Solid-State Devices: Methods of Testing Transistors."[5] In order to determine trigger amplitude and duration requirements in pulse applications, it is of particular importance to know the variations of r_i and r_r for emitter currents in the microampere range. In this case it may be necessary to resort to a balance or null method measurement.

2.2.3 Measurement of the DC Short-Circuit Forward Current Transfer Ratio: The dc short-circuit forward current transfer ratio, α_{FB}, is frequently used as a "forming" objective in the fabrication of point-contact transistors. The magnitude of α_{FB} may be obtained directly from the output characteristic. In switching applications, the dc current transfer ratio for low voltages in the vicinity of Region III is a convenient measure of the current amplification of the device and is defined by

$$\alpha_{FB} \equiv \frac{I_C(I_E, V_{CB}) - I_C(0, V_{CB})}{I_E}. \tag{6}$$

2.3 Region III DC Measurements

In Region III, the collector current is limited primarily by the external collector circuit. Consequently, relatively large currents flow in the emitter and collector circuits which results in the transistor acting as a passive network. Under these conditions, the slope of the input static characteristic (Fig. 1) is essentially the forward characteristic of the emitter as a simple diode in series with the Region III base resistance. Similarly, the slope of the output static characteristic for low values of collector voltage (Fig. 4), is essentially the forward characteristic of the collector as a simple diode in series with the Region III base ressistance.

2.3.1 $V_{CB}(I_E^*, I_C^*)$: The $V_{CB}(I_E^*, I_C^*)$ parameter is the voltage from collector to base under large-signal current biasing conditions. The emitter and collector current values are proportioned so that the point of measurement for a typical acceptable device is near the knee of the output characteristic in Region III (see point A, Fig. 4). The value of $V_{CB}(I_E^*, I_C^*)$ measured gives an indication of the nearness of the saturation curve to the $V_{CB}=0$ axis, and of the dc forward-current transfer ratio of the device, and thereby provides a measure of the effectiveness of the device as a closed switch. This measurement may also be performed on a curve tracer by displaying the output characteristic with the specified value of I_E as an independent parameter. At the specified value of I_C, the V_{CB} may be measured from the curve.

2.3.2 $V_{CE}(I_E^*, I_C^*)$: The $V_{CE}(I_E^*, I_C^*)$ parameter is the voltage from collector to emitter under large-signal current biasing conditions. The emitter and collector current values are proportioned so that the point of measurement for a typical acceptable device is near the Region III boundary. Under these biasing conditions, all of the device parameters (r_e, r_b, r_c, and r_m) are quite small, and the current which flows is limited primarily by the external circuit elements.

The $V_{CE}(I_E^*, I_C^*)$ parameter may be obtained by adjusting the emitter and collector currents to the specified values, and measuring the voltage between the collector and emitter. This characteristic is useful in determining the power dissipation of the transistor in Region III. In some circuit applications, where small values of circuit resistance are used, the switch voltage drop may be quite high, and precautions should be taken to insure that the rated power dissipation of the device is not exceeded.

2.4 Dynamic Measurements

The short-circuit forward current transfer ratio, α_{fb}, of a transistor is generally considered to be a good figure of merit in large-signal applications. The magnitude of α_{fb} is a function of the operating point and of frequency.

2.4.1 α_{fb}-vs-I_E Measurements: The α_{fb}-vs-I_E characteristic curve may be measured by dc, ac, or sweep methods. Methods of test are described in Sections 3.5 and 3.6 of "IRE Standards on Solid-State Devices, Methods of Testing Transistors."[5]

2.4.2 $|\alpha_{fb}|$-vs-Frequency Measurements: In the active region, the emitter current in a conventional point-contact transistor is primarily minority-carrier flow into the semiconductor, and the collector current is chiefly majority-carrier flow out of the semiconductor. Current multiplication results when an emitted minority carrier arrives at the collector rectifying barrier and liberates several majority carriers. The time required for the minority carriers to travel from the emitter to the collector limits the frequency response of the device and introduces a phase lag. This effect is analogous to transit-time effects in electron tubes.

The frequency response can be measured in terms of $|\alpha_{fb}|$-vs-frequency curves or pulse response. The components of the pulse response are considered in Section 2.4.3. $|\alpha_{fb}|$-vs-frequency measurements can be made by point-by-point small-signal or by sweep techniques. Methods of test are described in Sections 3.5 and 3.6 of "IRE Standards on Solid-State Devices, Methods of Testing Transistors."[5] A figure of merit, the α_{fb} cutoff frequency, $f_{\alpha b}$, may be obtained from the $|\alpha_{fb}|$-vs-frequency curve. This frequency, $f_{\alpha b}$, is the frequency at which $|\alpha_{fb}|$ is $\sqrt{2}/2$ times its low-frequency magnitude.

2.4.3 Pulse Response Measurements: Application of a pulse to the input and observation of the rise and fall times in the output is an accepted method of studying the frequency response of a device or circuit. Fig. 5 shows a typical pulse response characteristic. The output pulse may be divided into intervals[2] as defined below.

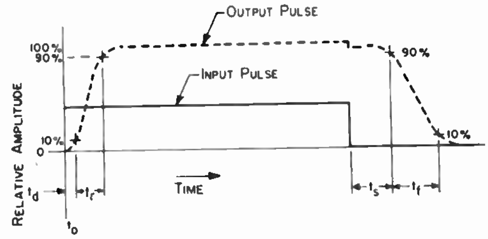
1) Propagation delay (t_d) is the time between the application of the input pulse and the time when the output pulse attains 10 per cent of its maximum amplitude.

2) Rise time (t_r) is the time duration during which the output pulse is increasing from 10 per cent to 90 per cent of its maximum amplitude.

3) Storage time (t_s) is the time between the end of the input pulse and the time when the output pulse has decreased to 90 per cent of its maximum amplitude.

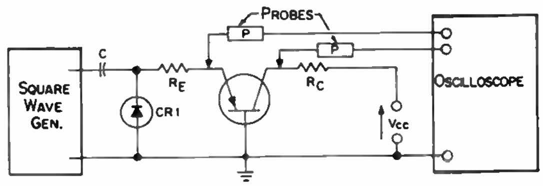
4) Fall time (t_f) is the time duration during which the output pulse is decreasing from 90 per cent to 10 per cent of its maximum amplitude.

After a pulse of injected carriers is applied to a point-contact transistor, minority carriers diffuse from the emitter to the collector and follow many paths of varying length. For this and other reasons, the collector current response to an applied pulse is not instantaneous. Immediately after an emitter pulse is applied and before the first carriers arrive at the collector, a small increase in collector current is often observed. This increase is produced by the voltage-divider action of the applied pulse across the emitter and base resistances. When the emitter current is suddenly terminated, the carriers in transit take different lengths of time to reach the collector, resulting in a finite fall time.



Definitions: t_d = propagation delay.
t_r = rise time.
t_s = storage time.
t_f = fall time.

Fig. 5—Pulse response.



Typical Operating Conditions

$V_{CC} = -15$ v	Pulse width ~4.0 µsec.
$R_E = 10{,}000\ \Omega$	Pulse amplitude ~10 v.
$R_C = 500\ \Omega$	Pulse rise time 0.01 µsec.
$CR\ 1$ = crystal diode	Pulse rep. freq. 5000 pps.

Fig. 6—Circuit for measuring pulse response.

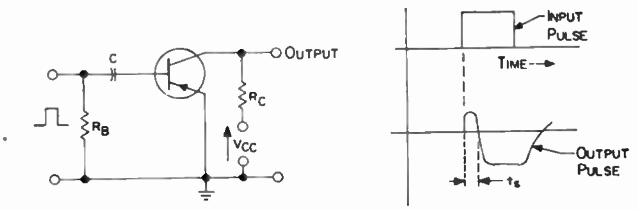


Fig. 7—Circuit diagram and pulse waveforms; storage-time measurement.

The propagation delay, rise, fall, and storage times cannot be calculated easily, and considerable variation may occur from one transistor to another. Storage time depends on the extent to which the transistor is driven into the current-saturation region (Region III) and on the duration of the input pulse.

Fig. 6 indicates a method of measuring pulse response. The diode clamps the base of the emitter square-wave signal at ground potential. Resistor R_E is large compared to the Region II input resistance of the transistor so that the signal generator approximates a constant-current source. Current-limiting resistor R_C must be small so that it will not appreciably affect the frequency response of the device under test.

Another circuit which is frequently used to measure the storage time is shown in Fig. 7. In the absence of an input pulse, the transistor is in the current-saturation region with unity saturation ($I_E = I_C$). With the application of a positive input pulse at the base, the collector voltage becomes more positive until the excess minority carrier density at the collector decreases to zero. The initial positive pulse is a result of "ohmic feed-through"; *i.e.*, initially the transistor acts as a passive device, and a portion of the input pulse also appears at the output terminals. The input pulse repetition-rate is sufficiently low to insure that the density of minority carriers will build up to a constant value between pulses. The amplitude of the input pulse must be sufficient to drive the transistor to collector-current cutoff.

The storage time in an actual switching circuit does not necessarily correspond to the value measured in the above circuit for three reasons: 1) the transistor is not always completely saturated, 2) the trigger pulses may differ in various circuits, and 3) the time that the transistor has been in the high-current condition depends on circuit and pulse arrangements.

2.4.4 Precautions: Visual display using an oscilloscope permits rapid measurements with an accuracy of about 5 per cent provided precautions are taken with respect to display bandwidth, repetition rate, and terminations.

2.4.4.1 Display Bandwidth: The test circuit and oscilloscope must have sufficient frequency response to pass many harmonics of the repetition rate if a faithful reproduction of the α_{fb}-vs-operating point or $|\alpha_{fb}|$-vs-frequency curves is to be observed. In general, the display bandwidth will be determined by the fastest rise or fall time of the characteristic to be displayed. In most cases, the display bandwidth (bw) will be adequate if:

$$\text{bw} = 2/t$$

where t is the fastest rise or fall time to be observed. Use of a sine-wave sweep and display of both the forward and reverse traces permits one to check for adequate bandwidth. Insufficient bandwidth may be revealed by hysteresis-like separation of the two traces.

2.4.4.2 Repetition Rate: Repetition rates less than about 25 cps will produce flicker and cause operator fatigue and measurement inaccuracies. The maximum repetition-rate limit is governed by display-bandwidth requirements and frequency-response characteristics. A low rate will minimize inaccuracies caused by inadequate display bandwidth and circuit frequency response.

2.4.4.3 Terminations: Terminations for the transistor and measuring circuit must be known and constant over the frequency band of interest. Instability may result if suitable resistance is not provided in series with the bias sources, particularly in the case of short-circuit-unstable point-contact transistors.

2.5 Negative-Resistance Characteristics

Transistors having short-circuit forward current transfer ratios in excess of unity may be employed in circuits to produce dynamic negative-resistance characteristics and hence may be used in switching-type circuits. The input characteristics of these transistors are useful for circuit investigation and for providing suitable specifications for the devices.

Since the performance of the devices in this type of application cannot be divorced from circuit parameters, these are included in the relationships among device parameters, terminal voltages and terminal currents given in the following section.

Precautions: In all of the following analyses, assumptions, and derivations specific to given regions, it is understood that the device-parameter values appearing in the equations for each region are assumed to be constant at the value pertinent to that region. Typical regional parameter values,[6] in ohms, are shown in Table I, for reference purposes only.

TABLE I

Region	Parameter			
	r_e	r_b	r_c	r_m
I	100,000	160	20,000	0
II	100	160	20,000	50,000
III	25	50	70	30

2.5.1 Open-Circuit-Stable Characteristics: The negative-resistance characteristics obtained for point-contact transistors looking between emitter or collector and a circuit common point are of the open-circuit-stable type[6-9] and have the form illustrated in Fig. 8 and Fig. 9, respectively.

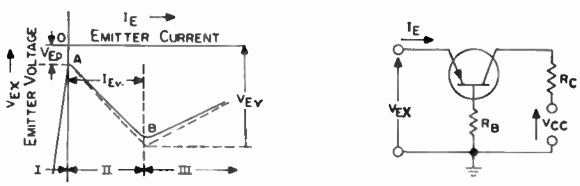


Fig. 8—Typical emitter negative-resistance characteristic.

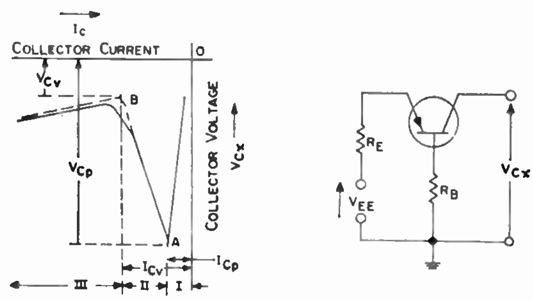


Fig. 9—Typical collector negative-resistance characteristic.

The device characteristics in each region may be considered approximately linear. These approximations are shown graphically in Figs. 1 through 4.

The general input-impedance relationship[10] for any

[7] B. G. Farley, "Dynamics of transistor negative resistance," Proc. IRE, vol. 40, pp. 1497–1508; November, 1952.
[8] E. Eberhart, R. O. Endres and R. P. Moore, "Counter circuits using transistors," *RCA Rev.*, vol. 10, pp. 459–476; December, 1949.
[9] A. W. Lo, "Transistor trigger circuits," Proc. IRE, vol. 40, pp. 1531–1541; November, 1952.
[10] L. C. Peterson, "Equivalent circuits of linear active four-terminal networks," *Bell Sys. Tech. J.*, vol. 27, pp. 593–622; October, 1948.

linear active two-port where the z's are the usual open-circuit loop impedances, may be applied for analysis within any region.

$$z_{\text{in}} = z_{11} - \frac{z_{12}z_{21}}{z_{22} + Z_C}. \tag{7}$$

2.5.1.1 Emitter Negative Resistance Characteristic (V_{EX} vs I_E): By application of (7), the following general relationship may be derived[6] for the circuit shown[11] in Fig. 8.

$$V_{EX} = \left[r_e + r_b + R_B - \frac{(r_b + R_B)(r_b + R_B + r_m)}{r_b + R_B + r_c + R_C}\right] I_E + \frac{V_{CC}(r_b + R_B)}{r_b + R_B + r_c + R_C}. \tag{8}$$

This general equation may be simplified by suitable approximations within each region. The device parameters appearing in the equations for each region will therefore be assumed to be constant within the region.

Region I: When a transistor is biased such that its operation is in Region I, the emitter and collector currents are in the reverse direction. Under this condition of operation the following assumptions can be made:

$$\begin{array}{l} \text{Region I} \\ \text{Device parameter values} \end{array} \begin{cases} r_b \ll R_B \\ r_e \gg (R_B + r_b) \\ r_m \approx 0. \end{cases}$$

Subject to the above assumptions, (8) becomes:

$$V_{EX} \approx r_e I_E + \frac{V_{CC} R_B}{R_B + r_c + R_C}. \tag{9}$$

This equation is that of a straight line, of slope or input impedance equal to r_e, the reverse resistance of the emitter diode, and an intercept on the voltage axis at $I_E = 0$:

$$V_{EXp} = \frac{V_{CC} R_B}{R_B + r_c + R_C}. \tag{10}$$

Region II: When a transistor is biased such that its operation is in Region II, the emitter current is in the forward direction, and the collector current is in the reverse direction. Under this condition of operation the only approximations which may be made are:

$$\begin{array}{l} \text{Region II} \\ \text{Device parameter values} \end{array} \begin{cases} r_b \ll R_B \\ r_e \ll R_B. \end{cases}$$

Subject to these assumptions, (8) becomes:

$$V_{EX} \approx \left[R_B - \frac{R_B(r_m + R_B)}{R_B + r_c + R_C}\right] I_E + \frac{V_{CC} R_B}{R_B + r_c + R_C}. \tag{11}$$

[11] Note that for a p-n-p transistor, V_{CC} is negative.

If $R_B + R_C$ is small in comparison with r_m and r_c, the further approximation may be made that:

$$\alpha_{fb} \approx \frac{R_B + r_m}{R_B + r_c + R_C}$$

and the relationship may be simplified to the form:

$$V_{EX} \approx R_B(1 - \alpha_{fb}) I_E + \frac{V_{CC} R_B}{R_B + r_c + R_C}. \tag{12}$$

The slope will be negative, provided α_{fb} is greater than unity, and the voltage intercept at $I_E = 0$ will be identical with that shown for Region I.

Region III: When a transistor is biased such that its operation is in Region III, the emitter current is in the forward direction, and the collector current is limited primarily by the external collector circuit. Under these conditions of operation the approximations will depend largely on the magnitude of the external resistances. If it is assumed that:

$$\begin{array}{l} \text{Region III} \\ \text{Device parameter values} \end{array} \begin{cases} r_c \ll R_C \\ r_b \ll R_B \\ r_c \ll R_B \\ r_m \ll R_B \\ r_e \ll R_B \end{cases}$$

then (8) becomes:

$$V_{EX} \approx \frac{R_B R_C}{R_B + R_C} I_E + \frac{V_{CC} R_B}{R_B + R_C}. \tag{13}$$

With the above assumptions the circuit is essentially independent of the device parameters. However, in many applications it is desired to focus attention upon those device parameters which will limit the maximum ON device current because of device dissipation limitations. The following assumptions may be made in order to focus attention upon the controlling device parameters:

$$\begin{array}{l} \text{Region III} \\ \text{Device parameter values} \end{array} \begin{cases} r_b \ll R_B \\ r_e \ll R_B. \end{cases}$$

In this case (8) becomes:

$$V_{EX} \approx \left[r_e + \frac{R_B(R_C + r_c - r_m)}{R_B + R_C}\right] I_E + \frac{V_{CC} R_B}{R_B + R_C}$$

and when $R_C = 0$:

$$V_{EX} \approx [r_e + r_c - r_m] I_E + V_{CC}. \tag{14}$$

Since all of the pertinent device parameters r_e, r_c, and r_m are quite small, the currents which flow may be quite high.

The emitter negative-resistance characteristic may be determined from (9), (11), and (13).

2.5.1.2 Collector Negative-Resistance Characteristic (V_{CX} *vs* I_C): The following general relationship may be derived[6] for the circuit shown in Fig. 9:

$$V_{CX} = \left[r_c + R_B + r_b - \frac{(r_b + R_B)(r_b + R_B + r_m)}{r_e + R_E + r_b + R_B} \right] I_C + \frac{V_{EE}(r_b + R_B + r_m)}{r_e + R_E + r_b + R_B}. \quad (15)$$

This general equation may be simplified by suitable approximations within each region. The device parameters are assumed to be constant within each region, at the value pertinent to that region.

Region I: Assumptions:

Region I

Device parameter values $\begin{cases} r_b \ll R_B \\ r_e \gg (R_B + r_b) \\ r_m \ll R_B \\ r_c \gg R_B \\ R_E = 0. \end{cases}$

Then (15) becomes:

$$V_{CX} \approx I_C r_c + \frac{V_{EE} R_B}{r_e + R_B}. \quad (16)$$

Region II: Assumptions:

Region II

Device parameter values $\begin{cases} r_b \ll R_B \\ r_e \ll R_B. \end{cases}$

In this case (15) becomes:

$$V_{CX} \approx \left[(r_c + R_B) - \frac{R_B(r_m + R_B)}{R_E + R_B} \right] I_C + \frac{V_{EE}(r_m + R_B)}{R_E + R_B}. \quad (17)$$

Region III: Assumptions:

Region III

Device parameter values $\begin{cases} r_b \ll R_B \\ r_c \ll R_B \\ r_m \ll R_B \\ r_e \ll R_B. \end{cases}$

Then (15) becomes:

$$V_{CX} \approx \frac{R_B R_E}{R_B + R_E} I_C + \frac{V_{EE} R_B}{R_B + R_E}. \quad (18)$$

Under somewhat modified assumptions made to focus attention upon the controlling device parameters in current saturation, the following equation may be derived:

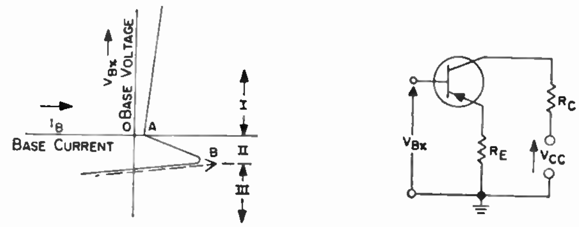


Fig. 10—Typical base negative-resistance characteristic.

Assumptions:

$$R_B \gg r_b, r_e$$
$$R_E = 0.$$

From (15)

$$V_{CX} \approx (r_e + r_c - r_m) I_C + V_{EE}. \quad (19)$$

The collector negative-resistance characteristic may be determined from (16)–(18).

2.5.2 Short-Circuit-Stable Characteristics: The negative-resistance characteristics obtained looking into the base terminal for point-contact transistors having short-circuit forward current transfer ratios in excess of unity are short-circuit stable, as shown[11] in Fig. 10. Again the characteristic may be divided into three regions, so that within each region the curve may be considered approximately linear. The following general relationship may be derived using (7):

$$V_{BX} = \left[r_b + R_E + r_e - \frac{(r_e + R_E)(r_e + R_E + r_m)}{r_e + R_E + r_c + R_C - r_m} \right] I_B + \frac{V_{CC}(r_e + R_E)}{r_e + R_E + r_c + R_C - r_m}. \quad (20)$$

Region I: Assumptions:

Region I

Device parameter values $\begin{cases} r_e \gg R_E \\ r_b \ll r_e \\ r_m = 0 \\ R_C = 0. \end{cases}$

With these assumptions (20) becomes:

$$V_{BX} \approx \left[\frac{r_e r_c}{r_e + r_c} \right] I_B + \frac{V_{CC} r_e}{r_e + r_c}. \quad (21)$$

Region II: Assumptions:

Region II $\quad r_c \gg r_e + R_E.$

Device parameter values

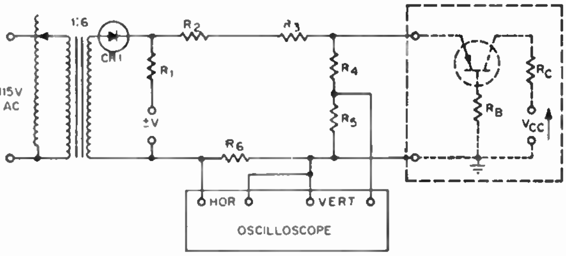


Fig. 11—Typical circuit for measurement of open-circuit-stable negative-resistance characteristics.

With these assumptions (20) becomes:

$$V_{BX} \approx \left[\frac{(r_c + R_C)(r_e + r_b + R_E) - r_m r_b}{r_c + R_C - r_m}\right] I_B + \frac{V_{CC}(r_e + R_E)}{r_c + R_C - r_m}. \quad (22)$$

Region III: Assumptions:

Region III $\quad r_m \ll r_e + r_c.$

Device parameter values

With these assumptions (20) becomes:

$$V_{BX} \approx \left[r_b + \frac{(r_e + R_E)(r_c + R_C)}{r_e + R_E + r_c + R_C}\right] I_B + \frac{V_{CC}(r_e + R_E)}{r_e + R_E + r_c + R_C}. \quad (23)$$

2.5.3 Methods of Measurement: The negative-resistance characteristics of point-contact transistors may be displayed by sweep methods. It is convenient to use power-line frequency sweeps for those circuits in which the external feedback circuit elements are purely resistive.

2.5.3.1 Open-Circuit-Stable Characteristics: A method of displaying the emitter open-circuit-stable negative resistance characteristic is shown[11] in Fig. 11. A similar circuit may be used to display collector negative-resistance characteristics. The ac input voltage, whose amplitude is controllable by means of an autotransformer, is impressed upon a low-capacitance step-up transformer and thence to the rectifying diode, CR 1, which is incorporated to prevent excessive reverse emitter currents and the consequent high voltages which would be developed from emitter to base electrodes. Bias to the diode may be supplied by the auxiliary voltage source V to control the point in the sine wave cycle where clipping will occur, and hence to choose the swept portion of the characteristic for which detailed information on behavior is desired. For instance, it may be desired to examine the characteristic more closely in the neighborhood of the peak or valley turning points. The resistance R_2+R_3 gives the signal impressed upon the transistor the characteristics of a constant-current source. The input current is determined by measuring the voltage drop across R_6. The input capacitance of the vertical amplifier of the oscilloscope is isolated from the transistor by the use of a high-impedance voltage-divider circuit, R_4 and R_5.

Care must be taken to minimize the stray circuit capacitance. Excessive capacitance from emitter to ground may result in oscillations over a portion of the negative-resistance characteristic in the active region. The series-feed resistance is divided into two or more resistors, with one of these connected as close to the transistor terminations as possible, and of sufficient magnitude to suppress the oscillations. The resistor R_6 should be as small as possible consistent with sensitivity requirements. Stray capacitance across R_6 should be minimized to reduce hysteresis in the observed characteristic.

If the external circuit feedback element from base electrode to circuit common point includes reactive components so that it is frequency dependent, modifications to this circuit are required. This is because the total impedance from base to circuit common point will then be a function of the rate of change of input current.

2.5.3.2 Turning Points: Peak and valley turning points in the negative-resistance characteristics shown in Fig. 8 and Fig. 9 are designated A and B respectively.

Peak Turning Points: The equations for the voltage and current coordinates of the peak turning points have been derived.[6]

Emitter Negative-Resistance Characteristic

$$V_{Ep} \approx \frac{R_B V_{CC}}{R_B + R_C + r_{co}}, \quad \text{where } r_{co} \equiv r_c \big|_{I_E=0}$$

$$I_E \big|_{V_{Ep}} = 0, \text{ by assumption.} \quad (24)$$

Collector Negative-Resistance Characteristic

$$V_{Cp} \approx \left[\frac{R_B + R_C + r_{co}}{R_B}\right] V_{EE}$$

$$I_C \big|_{V_{Cp}} \equiv I_{Cp} \approx \frac{V_{EE}}{R_B}. \quad (25)$$

Valley Turning Points: The equations for the valley turning points are:

Emitter Negative-Resistance Characteristic

$$V_{Ev} \approx \frac{V_{CC} R_B (1 - \alpha_{fb})}{R_B(1 - \alpha_{fb}) - \alpha_{fb} R_C} \quad (26)$$

$$I_E \big|_{V_{Ev}} \equiv I_{Ev} \approx \frac{V_{CC}}{R_B(1 - \alpha_{fb}) - \alpha_{fb} R_C} \quad (27)$$

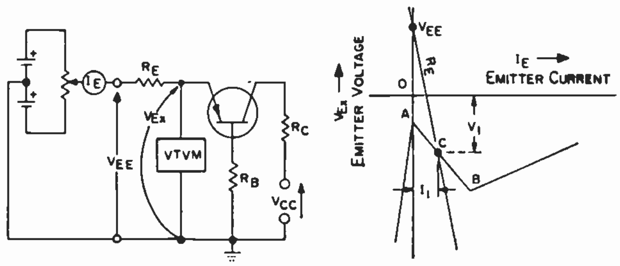


Fig. 12—Circuit for measurement of negative-resistance characteristic point-by-point method.

Collector Negative-Resistance Characteristic

$$V_{C_v} \approx \frac{V_{EE}R_B(1 - \alpha_{fb})}{R_B(1 - \alpha_{fb}) + R_E} \tag{28}$$

$$I_C\big|_{V_{C_v}} \equiv I_{C_v} \approx \frac{\alpha_{fb}V_{EE}}{R_B(\alpha_{fb} - 1) - R_E}. \tag{29}$$

2.5.3.3 Measurement Techniques: The circuit arrangement shown[11] in Fig. 12 may be used for dc examination of the negative-resistance characteristic close to the turning points. An illustrative example of the mechanism of operation is shown in the figure. A typical emitter input characteristic is shown with the transition point between Regions I and II (point A) and the transition point between Regions II and III (point B) being the peak and valley turning points respectively. The bias voltage V_{EE} applied to the emitter load resistance R_E causes intersection between the load line R_E and the characteristic at point C. DC monitoring meters will register the current and voltage I_1 and V_1, respectively, corresponding to the coordinates of the intersection point. This measurement must be correlated with the circuit requirements, and test limits may be placed upon the allowable variations in the coordinates of points A and B. Thus the actual measurements may consist simply of reading the currents and voltages of the points A and B without the necessity of plotting the entire characteristic.

Precautions: A double-peaked α_{fb}-vs-I_E characteristic invalidates the foregoing measurement technique, since this may result in the existence of a double valley which may not be indicated by dc measurements, or may be overlooked in the test procedures. Stray capacitance from the device terminals to ground must be minimized. A high-impedance electronic voltmeter must be used for voltage measurements to prevent erroneous current readings, and to keep the capacitance within acceptable limits. The information obtained is limited in its general applicability because the circuit elements R_B and R_C, the biasing potentials, and the device parameters determine the turning points. The limiting resistance R_E must be at least an order of magnitude greater than the negative slope for accuracy and stability of measurement. If a complex base impedance is to be employed the dc method of measurement is not directly applicable

Carrier-to-Noise Statistics for Various Carrier and Interference Characteristics*

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Summary—**Techniques are presented for the calculation of the statistical properties of the resultant carrier-to-noise ratios of systems subject to both additive and multiplicative noise. The cases considered are those in which the desired signal is either steady or exhibits Rayleigh fading while the interference consists of receiver noise plus an interfering signal which may be steady, Rayleigh fading, Gaussian fading, or Rayleigh fast fading with slow Gaussian fading of the median of the Rayleigh distribution. The results of the indicated calculations are presented graphically. A simple process is outlined for converting the data presented for use with any other signal strength.**

The utility of the data is demonstrated by a sample calculation the results of which are presented graphically. This calculation derives threshold data for the system as a function of desired signal power with the percentage of time that it operates in one of its two possible modes (fading or nonfading) as a parameter.

Attention is called to the technique demonstrated in the Appendix. This result appears quite useful in many joint or combination probability problems, especially those requiring numerical solutions.

Although this paper stemmed from work on line-of-sight microwave links, the results may be useful in other applications in the hf, vhf, and uhf regions.

Introduction

THE recent, widespread use of microwave links, as well as the planning for future installations, has led to a rather complete survey of the effects of various signal-interference combinations on the resultant carrier-to-noise ratios. The work lends itself to division into several parts: 1) the study of experimental data to determine the statistical properties of both the desired and undesired signals; 2) the determination of analytical procedures for combining the various signal-interference pairs; 3) the actual calculation and presentation of the data; and 4) a comparison with other work and an indication of the application of the data to a practical case. This paper will present only a very brief summary of the experimental background which has led to the various combinations to be considered. The interested reader will be provided with a starting point for further investigation of this phase.[1–3]

While most of the analytical work is straightforward, attention is called to the technique demonstrated in Appendix I. Although this result provides a very useful alternate method of solution to many probability problems, it does not appear to be either widely known or employed. The result consists of the substitution of single integration in the cumulative probability plane for a double integration in the probability density space. The possibility of such a substitution was called to the attention of the authors by B. F. Wheeler and S. B. Crosby of the RCA Laboratories in an unpublished memo.

The actual combinations of desired and interfering signals to be considered are listed for convenience in Table I. The numerical results are presented in several curves showing the per cent of the time that the carrier-to-noise will fall below the abscissa value vs the resultant carrier-to-noise ratio. The curves are plotted for the case of a steady signal 47 db above the receiver noise level or, with fading signals, for signal medians 47 db above the reference level.

Although the curves have been presented for a specific signal strength, they are almost directly applicable for any other signal strength. The justification for this generalization as well as specific information on how to accomplish it, are considered after the initial consideration of the various signal and interference combinations.

Analytical Expressions for Desired and Undesired Signals

Desired Signals—"Line-of-Sight" Paths

The analysis of experimental data[1–3] indicates that much of the time the signal, on good "line-of-sight" paths, is essentially constant. During periods where temperature inversions are possible such paths may exhibit a fast fading characteristic which may be reasonably approximated by a Rayleigh distribution having its median at the steady signal value.

Thus, the desired signal on a "line-of-sight" microwave path may be considered as either a steady signal or as a fast Rayleigh-fading signal.

Undesired Signals—"Line-of-Sight" Paths

If one assumes that good engineering practices have been followed in laying out the microwave links in an area, then the interfering signals on a line-of-sight path should consist mainly of signals propagated "over-the-horizon" from distant links. Referring again to the experimental data one finds that such signals are charac-

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[1] A. L. Durkee, "Results of a microwave propagation test on a 40 mile overland path," Proc. IRE, vol. 36, pp. 197–205; February, 1948. (See especially Fig. 8.)

[2] J. H. Chisholm, P. A. Portmann, J. T. de Bettencourt, and J. F. Roche, "Investigations of angular scattering and multipath properties of tropospheric propagation of short radio waves beyond the horizon," Proc. IRE, vol. 43, pp. 1317–1335; October, 1955.

[3] H. R. Mathwich, E. D. Nuttall, J. E. Pitman, and A. M. Randolph, "Propagation test on microwave communications systems," *Elec. Eng.*, vol. 75, pp. 1020–1025; November, 1956.

terized by a short-term (periods less than several minutes) fast fade which may be reasonably approximated by a Rayleigh distribution whose median is the median value over the period.

If longer observations are made of the median of the signal then one observes that the median may be approximated by a decibel Gaussian distribution having a standard deviation of about 8 or 8.5 db. The composite undesired signal or interference thus may be postulated as consisting of one of the following combinations (for completeness the case of interference from another "line-of-sight" station has been included).

1) A steady interfering signal plus a steady receiver noise. This covers the case of receiver noise only as well as the possibility of interference from another "line-of-sight" station during its nonfading periods.
2) A Rayleigh fading interfering signal (with a constant median) plus a steady receiver noise. This case would cover a "line-of-sight" interference when it was subject to fast fading as well as short-period observations of an "over-the-horizon" station.
3) A Rayleigh fading interfering signal whose median varies according to a Gaussian distribution having a standard deviation of 8.5 db plus a steady receiver noise. This is the normal, long-period observation case for "over-the-horizon" interfering signals.

In all cases the receiver noise and the external interfering signal have been combined by the direct addition of the receiver noise power (assumed constant) to the assumed interfering signal power. Table I contains the various combinations to be considered, allowing easy reference to the various curves.

In all cases the composite interference is generated by adding a constant receiver noise power to the external distribution.

TABLE I

Case	Signal	External Interference	Remarks
1	Steady	Steady	Fig. 1, carrier/noise never falls below given value.
2	Steady	Rayleigh fade	Fig. 1, curves for median interference values of −10, 0, +10, +20, and +30 db above receiver noise level.
3	Steady	Fast Rayleigh fade with slow fading Gaussian median.	Fig. 1, curves for the cases where the median exceeds the receiver noise for 20 per cent and for 56 per cent of the time respectively.
4	Rayleigh	Steady	Fig. 2.
5	Rayleigh	Rayleigh	Fig. 2, curves for median interference values of −10, 0, +10 and +20 db above receiver noise level.
6	Rayleigh	Fast Rayleigh with slow fading Gaussian median	Fig. 2, curve for case where median exceeds receiver noise level for 20 per cent of the time.
7	Rayleigh	Slow fading Gaussian only	Fig. 2, curve for case where the interfering signal exceeds receiver noise level for 20 per cent of the time.

Analytical Procedures

If both the desired and interfering signals are expressed in decibel form, then at any instant the carrier to interference "ratio" in decibels is equal to the difference in the two decibel levels.

Thus in the steady signal cases, probability of the carrier/interference (C/N) falling below any level is known if the probability of the interference exceeding any threshold level (T) is known, where S db $-T$ db $=(C/N)$ db; S db is the carrier expressed in db above the receiver noise level.

In the steady interference, fading signal cases the resultant probability of (C/N) db below any value is the probability of the signal falling below the level (T), where now T db $-I$ db $=(C/N)$ db; I db is the composite interfering signal expressed in db above the receiver noise.

When both the desired signal and the interference fade, then one must consider the joint probability of the signal and interference both varying so as to approach to within the (C/N) db of each other.

The procedure in all cases is then, first to change the variables so that the distributions appear in decibel form, and second, to determine the appropriate cumulative probability, often employing the result demonstrated in Appendix I. This result is as follows:

Given two independent probability densities, $p(x)$ and $p(y)$, together with some relation between x and y that effectively divides the $x-y$ plane into two parts (or given two probability densities $p(x)$ and $p_x(y)$ and some restriction on y that again divides the plane into two portions).

Then the probability of x and y falling within the closed region in the first case (or of y falling between the two restrictive values y_1 and y_2 in the second case) is equal to the area of the enclosed curve in the $P(x)-P(y)$ [or $P(x)-P_x(y)$] planes when x is allowed to vary from one extreme value to the other.

Consideration of Specific Cases (See Table I)

Case 4—Rayleigh Fading Signal—Steady Interference

This case is chosen for initial consideration since it requires only the cumulative probability of a Rayleigh distribution exceeding any level T. Since all work is in terms of decibels it is necessary to change the linear Rayleigh distribution:

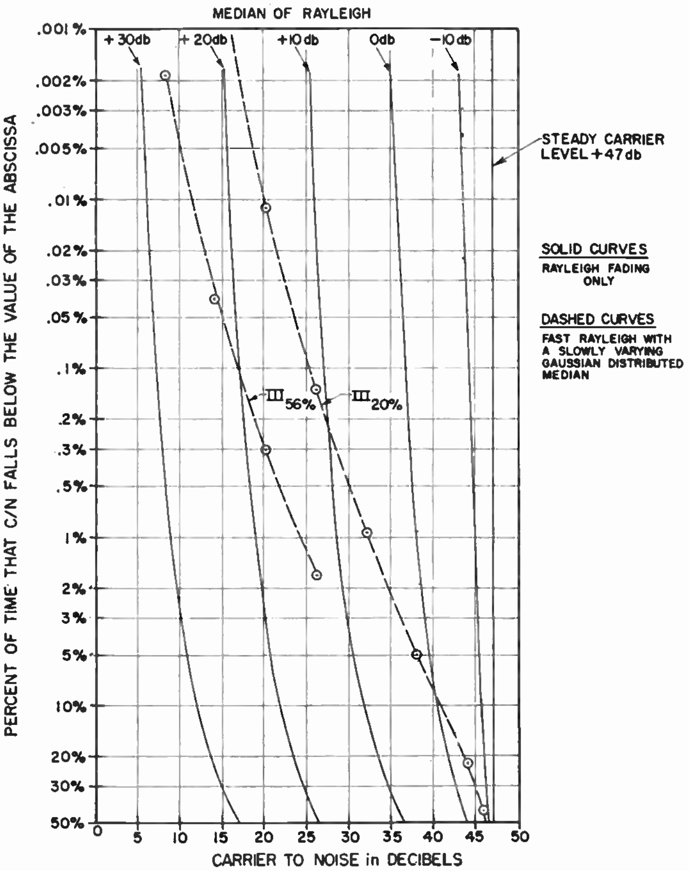


Fig. 1—Effects of interference on a steady carrier. (Carrier level 47 db above receiver noise.) Solid curves show the effect on the C/N ratio of the combination of receiver noise and a Rayleigh fading interference. The decibel values are the median value (with respect to the receiver noise level) of the Rayleigh interference.

The Case 3 curves consider the situation when the short-term median of the fast Rayleigh fade of the interference follows a slow Gaussian distribution ($\sigma=8.5$ db). The 20 per cent curve is for a Gaussian of mean equaling 7.15 db below the receiver noise level and the 56 per cent curve for a Gaussian mean 1.35 db above the receiver noise level. The percentage figures indicate the per cent of the time that the interfering signal exceeds the receiver noise level for these two mean values.

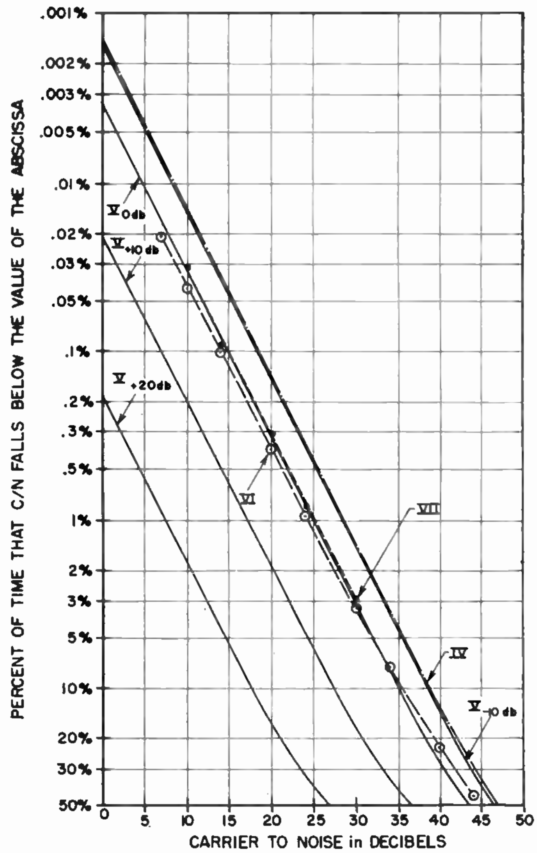


Fig. 2—The effect of various interferences on a Rayleigh fading signal. (The median of the signal is 47 db above the receiver noise level.) Case 4 is the case of receiver noise as the sole interference. Case 5, receiver noise plus a Rayleigh fading interference. Decibel levels are median of Rayleigh with respect to receiver noise. Case 6, receiver noise plus a fast Rayleigh fading interference whose short-term median follows a Gaussian distribution having a mean of −7.15 db and a standard deviation of 8.5 db. Case 7, receiver noise plus a signal exhibiting a Gaussian distribution having $\sigma=8.5$ db and a mean value 7.15 db below the receiver noise level.

$$p(r) = \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} \qquad P(r) = e^{-r^2/2\sigma}$$

to a decibel form. The change of variable for a probability density[4] involves recognition of the fact that the probability of the transformed value, g lying within the interval g, $g+dg$ must be equal to the probability of the initial variable, r lying within the interval r, $r+dr$. Therefore if $g=f(r)$ then

$$p(g) = \frac{p(r)}{\dfrac{d[f(r)]}{dr}}.$$

At the same time that the Rayleigh was converted into decibels, the relation between the median, M, of a linear Rayleigh and its standard deviation, σ, was employed to change the parameter from σ to M. The result of these transformations is (1) for the cumulative probability,

$$P_M(g > g_1) = \exp - 0.693[10^{g_1/10}][10^{-M/10}] \qquad (1)$$

where g is the signal level in decibels with respect to an arbitrary level, N, and M is the level in decibels, with respect to the same level, of the median of the Rayleigh.

In Case 4 we set $M=47$ db and derive the data for Curve IV of Fig. 2 by setting g equal to various threshold values T, as explained previously, and taking $1-P_M(g>g_1)=P_M(g<g_1)$.

Case 2—Steady Signal—Rayleigh Fading Interference

The addition of a constant receiver noise to a Rayleigh fading interference is handled by considering the level, N, as the receiver noise level. One then directly adds powers, which amounts to finding the power which must be supplied by the Rayleigh to reach any level g. The resultant probability for the composite interference exceeding g_1 is

[4] T. C. Fry, "Probability and Its Engineering Uses," D. Van Nostrand Co., Inc., New York, N. Y., sec. 63; 1928.

$$P_M(g > g_1) = \exp - 0.693[10^{g_1/10} - 1][10^{-M/10}]. \quad (2)$$

The data for the curves for Case 2 shown in Fig. 1 were derived by setting M (the median of the interference) successively to -10, 0, $+10$, $+20$, and $+30$ db with respect to the receiver noise level.

Case 3—Steady Signal—Composite Fading Interference

In this case the probability of Case 2, (2), has an M which instead of being a constant, as in Case 2, has a Gaussian distribution. By the definition of a Gaussian

$$P(M) = \frac{1}{\sigma\sqrt{2\pi}} \int_M^\infty \exp - \frac{(M - \mu)^2}{2\sigma^2} dM, \quad (3)$$

where σ is the standard deviation and μ the mean. From experimental data one finds that $\sigma = 8.5$ db is a reasonable value to choose. As the mean, μ, varies, the per cent of time that the median of the Rayleigh exceeds the receiver noise level varies. Numerical work has been carried out for $\mu = -7.15$ db and $\mu = +1.35$ db, leading to cases where the median of the Rayleigh exceeds the receiver noise level, N, for 20 per cent and 56.3 per cent of the time respectively. The actual calculation was done as follows: for any particular value of g, μ, and σ; $P(M)$ (from Tables[5]) and $P_M(g)$ [from (2)] were plotted as M varied from minus to plus infinity. This led to a curve in the $P(M)$-$P_M(g)$ plane. Now, according to Appendix I, the area under this curve should be the probability of the composite interference exceeding the level g. Fig. 3 shows a plot of this composite probability $P(\text{III})$ vs g for the cases of $\mu = -7.15$ db and $\mu = +1.35$ db.

From previous arguments the probability of the carrier/interference falling below any level, $(C/N)_{\text{db}}$, in this case, is just the probability of the interference exceeding any level, T where $S_{\text{db}} - T_{\text{db}} = (C/N)_{\text{db}}$.

Curves $\text{III}_{20 \text{ per cent}}$ and $\text{III}_{56 \text{ per cent}}$ of Fig. 1 present the resultant carrier/noise vs the percentage of the time that they are not exceeded. This figure offers a comparison between the fast fading, constant median interference curves of Case 2 and the fast fading, slowly varying median interference curves of Case 3.

Case 5—Rayleigh Fading Signal—Rayleigh Fading Interference

In this case we desire the resultant signal-to-noise ratio when a Rayleigh fading signal is combined with a composite interference made up of a Rayleigh fading interfering signal plus receiver noise. The form of these signals is such that one can integrate directly. This may be done either with probability densities or, utilizing the result from the Appendix, with the cumulative probabilities. Eq. (1) gives the form for the cumulative probability of the signal and (2) for the cumulative probability of the composite interference. If R_1 is the signal-to-noise ratio, g is the interference level, M_s is the

[5] "Tables of Normal Probability Functions," Applied Mathematics Series, vol. 23, Natl. Bur. Stand., Washington, D. C.

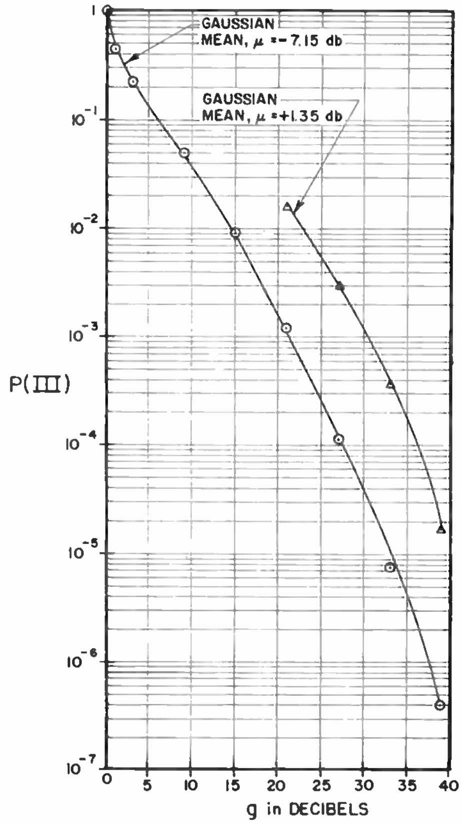


Fig. 3—Probability of a Rayleigh-Gaussian signal exceeding a given level, g. The signal under consideration consists of the addition of the powers of a steady level, N, and a variable signal. The variable signal has a short-term Rayleigh distribution whose short-term median follows a Gaussian distribution of standard deviation 8.5 db. The two curves are for Gaussian mean values of -7.15 db and $+1.35$ db. All decibel values are with respect to the level N.

median of the signal Rayleigh distribution, and M_I is the median of the interfering Rayleigh distribution (all expressed in decibels with respect to the receiver noise level) then

$$P\{(S - g) < R\} = \int_{-\infty}^{+\infty} \int_{-\infty}^{R_1+g} p(g)p(S)dSdg, \quad (4)$$

where

$$p(S) = 0.1595\ 10^{(S-M_s)/10} \exp - 0.693\ 10^{(S-M_s)/10}, \quad (5)$$

$$p(g) = 0.1595\ 10^{(g-M_I)/10} \cdot \exp - 0.693(10^{g/10} - 1)(10^{-M_I/10}). \quad (6)$$

The integration is simplified if, after the inner integration, the change of variable

$$A = 10^{g/10}; \quad dg = \frac{10dA}{2.3A}$$

is introduced.

The resultant is shown as

$$P\{(S-g)<R_1\} = 1 - \frac{\exp - 0.693\, 10^{(R_1-M_s)/10}}{1+10^{(R_1-M_s+M_I)/10}}. \quad (7)$$

Curves of carrier-to-noise for interfering medians of −10, 0, +10, +20 db are shown as curves V_{-10db}, V_0, V_{+10db}, and V_{+20db} of Fig. 2.

Case 6—Rayleigh Fading Signal—Composite Fading Interference

In this case the results for the interference from Case 3 are combined with the Rayleigh fading signal in a numerical integration. The curve of P(III) vs g of Fig. 3 yields the probability of the assumed interference exceeding any given level, while (1) of Case 4 yields the probability of the signal falling below any chosen decibel level. A level is assumed, $R_{db}=S_{db}-g_{db}$ and $P(s)$ vs $P(g)$ plotted as g varies from 0 to ∞. According to the Appendix the area under the curve is the probability of this level not being exceeded. A curve of carrier-to-noise vs percentage of time below the ordinate for $\mu=-7.15$ db (the median of the interfering signal exceeds the received noise 20 per cent of the time) is presented as Curve VI of Fig. 2.

Case 7—Rayleigh Fading Signal—Slow Gaussian Fading Interference

This case was considered to determine the effect of neglecting the fast fade of the interfering signal. The procedure is the same as Case 6 except that now one substitutes

$$P(q) = \frac{1}{\sigma\sqrt{2\pi}}\int_h^\infty \exp - \frac{(h-\mu)^2}{2\sigma^2}\, dh, \quad (8)$$

for $P(g)$ in Case 6. In (8), $h=10 \log (10^{q/10}-1)$ where q is the interfering signal level under consideration. Eq. (8) comes directly from the definition of a gaussian when provision is made for the effect of the added receiver noise. As in Case 3 values of $P(q)$ may be obtained from published tables of normal probability functions. $P(s)$ and $P(q)$ are plotted for various assumed carrier/noise ratios. Curve VII of Fig. 2 shows the result for a mean of −7.15 db and a standard deviation of 8.5 db, *i.e.*, for the case where the slowly varying interfering signal exceeds the receiver noise level for 20 per cent of the time.

Conversion of the Curves for Use With Other Signal Strengths

Cases 2 and 3 (Table I)

These cases consider a steady signal, S, and a composite interference. Consideration of (2) shows that, for a given interference median M_I, $P_M(g>g_1)$ depends only on g. The carrier-to-noise, in this case, is $R_{db}=S_{db}-g_{db}$. Therefore if R_{db} and S_{db} are both changed by an amount Δ, g_{db} and hence $P_M(g>g_1)$ will be unchanged. Fig. 1 is plotted for the case where $S_{db}=47$. If any other signal strength is desired one simply relabels the abscissa by adding Δ to all the present carrier-to-noise values, where $\Delta=S_{new}-47$.

Case 4 (Table I)

In this case the interference is constant and the signal fades. Consideration of (1) shows that $P_M(g<g_1)$ depends upon $(g-M_s)$; hence if both g and M_s are changed by the addition of Δ, the probability will remain constant. Furthermore, in this case the carrier-to-noise, R_{db}, is equal to g_{db}. Therefore, if M_s is changed by adding Δ, a relabeling of the abscissa identical to that proposed above (*i.e.*, the addition of $\Delta=(M_{s_{new}}-47)$, will yield the correct curves for the new median signal strength.

Cases 5–7 (Table I)

In these cases both the interference and the signal are fading. However, for all cases the determining parameter is $(R-M_s)$; hence for any given M_I, if M_s and R are both changed by Δ, the curve is unchanged.

To sum up, both Figs. 1 and 2 are immediately applicable for any new signal strength by merely relabeling the abscissa, carrier-to-noise values. The conversion requires the addition of Δ to the existing values where $\Delta=S_{db}-47$ for Fig. 1 and $\Delta=M_s-47$ for Fig. 2.

Use of the Curves and Comparisons With Other Work

Results very similar to several of the cases reported here have been reported recently by Bond and Meyer,[6] who consider situations similar to Cases 1, 4, and 5 except that they do not include the effect of the constant receiver noise. Their results are presented on the basis of the necessary increase in mean-signal to mean-interference ratio with fading, in order that the resultant signal to noise be maintained at the nonfading value.

In the case of the fading signal and the nonfading interference (Case 4) our results (Curve VI, Fig. 2) agree exactly with Curve 3 of Bond and Meyer's Fig. 1, when the 1.6-db difference between the mean and the median of a Rayleigh is accounted for.

To see the effect of including the receiver noise on the Rayleigh signal-Rayleigh interference case, one needs to compare the results of Curve 2, Fig. 1, of Bond and Meyer's paper with those obtained from curves for Case 5 in Fig. 2 of this paper. If one assumes an initial signal-to-noise ratio of 27 db then curves V_{20db}, V_{10db}, and V_{0db} would yield satisfactory operation times of 49.6 per cent, 90.3 per cent and 98.3 per cent for fading median ratio increases of 0 db, 10 db, and 20 db respectively. Bond and Meyer's data yield 50 per cent, 90.9 per cent, and 99.0 per cent satisfactory operation times for the same increases.

[6] F. E. Bond and H. F. Meyer, "The effect of fading on communication circuits subject to interference," Proc. IRE, vol. 45, pp. 636–642; May, 1957.

Presumably the slight reduction in the amount of time that the 27-db signal-to-noise is obtained in our results is the effect of adding in the constant receiver noise.

Another recent paper[7] discusses the cases of the distributions resulting from the linear addition of two Rayleigh variables and from the addition of a Rayleigh and a Gaussian variable. In both cases approximate expressions are obtained which allow the calculation of the probability of the combination failing to exceed some arbitrary level. However, as far as is known the data of Cases 3 and 6 have not been reported previously, and are believed to be new results.

It might appear that since the curves for 6 and 7 are within a decibel of each other, the fast fading of the interference might always be ignored. While there are some physical arguments in favor of this assumption it is not clear from our results that this neglect will necessarily always be justified.

It might also appear that since the curve for V_{0db} lies almost on top of the curves for Cases 6 and 7 and that since Case 5 may be solved analytically, there is no need to undertake the numerical integrations inherent in Cases 6 and 7. The problem here is that while choosing the median of the Rayleigh interference (Case 5) at zero decibels happens to make these three curves coincide, there does not appear to be any obvious method to determine this fact beforehand (*i.e.*, if the median of the Rayleigh had been chosen as some other value, then the coincidence of the curves would vanish).

The curves of this paper might be utilized in various ways to arrive at conclusions concerning the expected per cent of time that any particular link can be expected to yield any desired signal-to-noise ratio. If the link and the interference always have the same fading characteristics, then the curves may be used directly. If the desired and undesired distributions vary with time, then one faces the additional problem of dividing the total operating time into parts, each of which is reasonably closely approximated by one of the cases.

For example, if one assumes that the interference always consists of "over-the-horizon" signals of magnitudes sufficient to cause the median of the short term interference to exceed the receiver noise level for 20 per cent of the time, and in addition assumes that 95 per cent of the time the desired signal for the link in question is nonfading (at a level 47 db above receiver noise) while 5 per cent of the time it has a Rayleigh fade (with median 47 db above receiver noise), then from the curves for Cases 3 and 6 the over-all C/N could be expected to fall below 20 db for

$$(0.95\times 0.011)+(0.05\times 0.38)=0.0295 \text{ per cent of the time.}$$

[7] H. L. McCord, "An estimate of the degradation in signal detection resulting from the addition of the video voltages from two radar receivers," 1957 IRE National Convention Record, pt. 2, pp. 83–89.

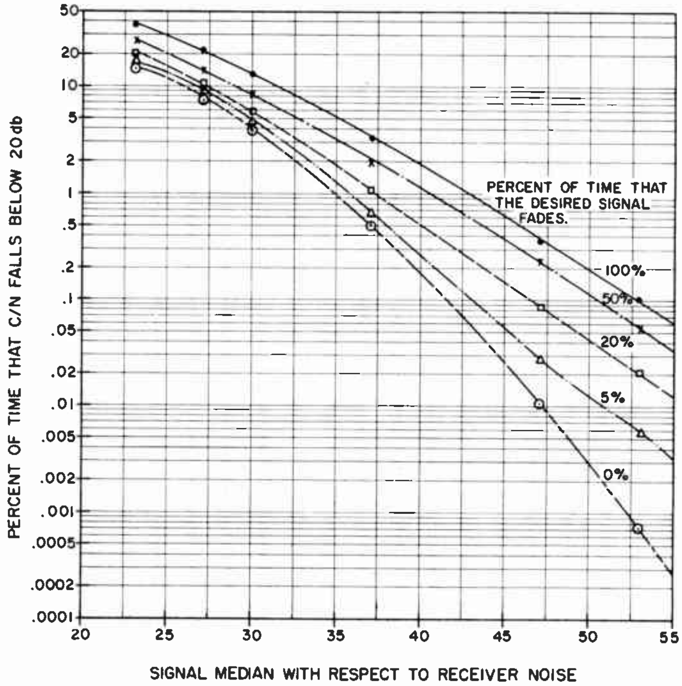


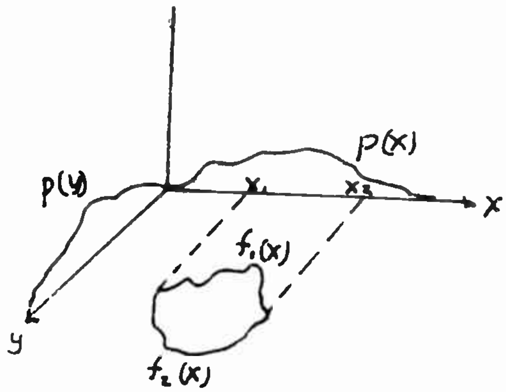
Fig. 4—Failure to exceed threshold vs received signal power with fading time as a parameter. A desired signal which exhibits a fast Rayleigh for the per cent of the total time indicated on each curve is combined with an interference consisting of receiver noise, plus an over-the-horizon interfering signal exhibiting a fast Rayleigh fade whose median follows a slow Gaussian variation so that the median of the Rayleigh exceeds the receiver noise level for 20 per cent of the time (type 3 or 6, Table I). The curves show the per cent of the time that the resultant C/N fails to reach 20 db as a function of the received signal level. The parameter is the percentage of the total time during which signal fading occurs.

The 5 per cent curve of Fig. 4 is generated by repeating this calculation as the signal median is varied from 23 db to 55 db (a previous section outlines the procedure for utilizing Figs. 1 and 2 as the signal median varies from the 47-db case). The zero per cent curve of Fig. 4 comes directly from Curve $\text{III}_{20 \text{ per cent}}$ of Fig. 1 by varying the signal median and the 100 per cent curve of Fig. 4 comes directly from Curve VI of Fig. 2, again by allowing the signal median to vary. The 20 per cent curve and the 50 per cent curve of Fig. 4 are derived by a procedure identical to that employed in the 5 per cent case and cover respectively the case where the desired signal is steady 80 per cent of the time and fades for 20 per cent of the time, and the case where the desired signal fades for 50 per cent of the time.

Appendix I

Given two independent probability densities, $p(x)$ and $p(y)$ together with some relation between x and y that effectively divides the x-y plane into two parts, or given two probability densities $p(x)$ and $p_x(y)$ and some restriction on y that again divides the plane into two portions.

Then the probability of x and y falling within the closed region in the first case or of y falling between the two restrictive values y_1 and y_2 in the second case is equal to the area of the enclosed curve in the $P(x)-P(y)$ or $P(x)-P_x(y)$ planes when x is allowed to vary from one extreme value to the other. This can be seen as follows:



The probability of x and y lying within the area enclosed by $f_1(x)$ and $f_2(x)$ is

$$P(x, y) = \int_{x_1}^{x_2} \int_{f_1(x)}^{f_2(x)} p(x)p(y)dydx \tag{9}$$

but

$$\int_{f_1(x)}^{f_2(x)} p(y)dy = \int_{-\infty}^{f_2(x)} p(y)dy - \int_{-\infty}^{f_1(x)} p(y)dy \tag{10}$$

and by definition

$$P(y) = \int_{-\infty}^{y} p(y)dy. \tag{11}$$

Therefore

$$P(x, y) = \int_{x_1}^{x_2} \{ P_y[f_2(x)] - P_y[f_1(x)\} p(x)dx. \tag{12}$$

However, we may use the definition of a cumulative probability (11) to change the variable from dx to $d(P(x))$. When this is done one obtains

$$P(x, y) = \int_{P(x_1)}^{P(x_2)} \{ P_y[f_2(x)] - P_y[f_1(x)]\} d[P(x)] \tag{13}$$

which is the area within the curve generated in the $P(x)-P(y)$ plane as x varies from x_1 to x_2. By an identical procedure the second case yields a similar result.

This result provides a useful solution in two classes of problems. In the first $P_y(f(x))$ and $P(x)$ are both given and one is expressible in terms of the other so that it is possible to integrate the resultant expression.

In the second case the $P_y(f(x))$ and $P(x)$ are tabulated functions but a simple analytic expression for the integration does not exist. In this case (13) lends itself to simple numerical integration techniques. This approach was used in Cases 3, 6, and 7.

Simultaneous Asynchronous Oscillations in Class-C Oscillators*

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***Summary*—Consideration is given to simultaneous asynchronous oscillations in systems having a moving reference point. The system chosen for study is a Class-C pentode oscillator. It is shown, both analytically and experimentally, that self-starting asynchronous oscillations are attainable in such a system. Transient and steady-state solutions are obtained for the Class-C oscillator containing one and two degrees of freedom.**

The method of analysis employed is that of equivalent linearization as described by Kryloff and Bogoluiboff. The existence of asynchronous oscillations is also explained by the phenomenon of negative discrimination.

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Introduction

A FAIRLY general equivalent circuit of an oscillator having one degree of freedom is shown in Fig. 1. The components L, G, and C represent the external frequency-determining elements and the element denoted by NL represents the nonlinear negative-resistance device necessary to sustain oscillation. This nonlinear element may represent a vacuum tube, transistor, klystron, or any other suitable active element.

For these devices it is usually possible to express the current i as some function of the terminal voltage v. This function will of course depend upon the nature of the nonlinear device.

Let

$$i = f(v). \tag{1}$$

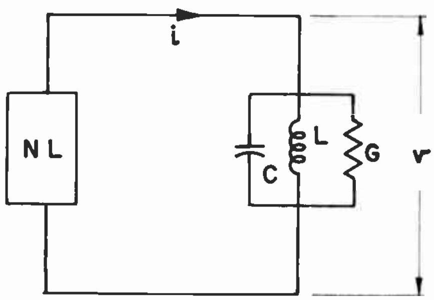

Fig. 1—Equivalent circuit of an oscillator.

Then the system equation may be written as

$$\ddot{v} + \delta\dot{v} + \omega_0^2 + \dot{f}(v)/C = 0 \qquad (2)$$

where $\omega_0^2 = 1/LC$ and $\delta = G/C$.

Assuming quasi-linear elements, it is possible to approximate the solution by

$$v = A \sin(\omega_0 t + \phi). \qquad (3)$$

A and ϕ are slowly varying functions of time as compared to $\sin(\omega_0 t)$. This limitation restricts A and ϕ to vary by only a small fraction of their value during one period of the normal oscillating frequency. A solution of this type determines the envelope of the oscillator output during buildup, thereby making it possible to obtain the length of time necessary for an oscillator to reach a stable condition.

The nonlinear elements may be grouped into three classes depending on their mode of operation.

1) There are those elements whose characteristics depend only on the rms value of the voltage appearing across their terminals. Thermally sensitive devices fall into this category and they may be represented by

$$i = f_1(A) \sin(\omega_0 t + \phi). \qquad (4)$$

This relation holds, provided the thermal time constant is sufficiently large compared to $1/\omega_0$. One example of an oscillator employing such an element is the Wien-bridge oscillator containing a lamp in the cathode lead.

2) Elements such as diode clippers and vacuum tubes respond to the "instantaneous" value of the applied voltage and may be represented by

$$i = f_2[A \sin(\omega_0 t + \phi)]. \qquad (5)$$

3) The third group of elements, and those with which this paper is concerned, have a moving reference point. The most common example is the Class-C amplifier or oscillator, in which the bias is derived by rectification of the rf voltage applied to the grid. In this case the bias depends on the peak value of the signal and the current may then be written as

$$i = f_3[A, A \sin(\omega_0 t + \phi)]. \qquad (6)$$

When the moving reference point is taken into consideration it is shown that an ordinary Class-C oscillator is able to sustain simultaneous oscillations at two unrelated frequencies and that these oscillations will build up from noise. This conclusion is verified experimentally.

The subject of simultaneous oscillations in electronic circuits has already received extensive analysis. One of the first treatments of this subject was given by Van der Pol,[1] in which he considered a triode generator with two degrees of freedom. By assuming that the volt-ampere characteristic of the triode could be represented by a cubic relation, he came to the conclusion that simultaneous oscillations were impossible. Extension of his analysis shows that simultaneous oscillations may exist when the frequency of one of the modes is three times that of the other.

Simultaneous oscillations are usually divided into two classes. When an oscillator is simultaneously operating at two frequencies, f_1 and f_2, and the relation $mf_1 = nf_2$ holds, where m and n are integers, the oscillations are said to be synchronous. If this relation cannot be satisfied, the oscillations are then termed asynchronous. Experimentally, it is difficult to determine if a pair of frequencies are asynchronous because it is impossible to measure any frequency exactly. Thus to any specified precision of frequency measurement it will always be possible to find values of m and n such that the above relationship between f_1 and f_2 will be satisfied.

It will be shown that as the sum of m and n becomes large the tendency toward synchronization becomes so small that the tendency is not enough to overcome the small frequency fluctuations due to noise in the circuit elements. Thus, in practical situations, synchronous oscillations do not occur in situations where both m and n are large.

Skinner[2] has shown that in order for stable asynchronous oscillations to occur, there must be at least a fifth-order term in the power-series expression of the volt-ampere characteristic. Therefore, it must be possible to represent (5) by the series

$$i = f(v) = A_1 v + A_2 v^2 + A_3 v^3 + A_4 v^4 + A_5 v^5. \qquad (7)$$

For stable asynchronous oscillations to occur it is necessary that

$$A_3 < 0$$

$$A_5 > 0$$

and that the antiresonant impedance of the two modes be not too unequal. For this case, however, the oscillations are not self-starting. This limitation is pointed out by Schaffner,[3] who developed a general method for ob-

[1] B. van der Pol, "An oscillation hysteresis in a triode generator with two degrees of freedom," *Phil. Mag.*, vol. 43, pp. 700–719; April, 1922.

[2] L. Skinner, "Criteria for stability in circuits containing nonlinear resistance," Ph.D. dissertation, University of Illinois, Urbana, Ill.; 1948.

[3] J. Schaffner, "Simultaneous oscillations in oscillators," IRE Trans. on Circuit Theory, vol. 1, pp. 2–8; June, 1954.

taining the transient and steady-state solutions for both synchronous and asynchronous oscillations.

In the last part of this paper, the existence of asynchronous simultaneous oscillations will be explained on the basis of the discrimination of nonlinear elements. Discrimination is defined as the ratio of the gains for two coexisting signals of different frequencies which pass through a nonlinear device at the same time.

Equivalent Linearization

A short and nonrigorous derivation of Kryloff's and Bogoluiboff's[4] method of equivalent linearization will now be presented so that the assumptions that are made will be brought to light. This method has proved to be extremely useful in studying circuits containing time-varying as well as nonlinear elements.

Differentiation of (3) gives

$$\dot{v} = \dot{A}\sin(\omega_0 t + \phi) + \omega_0 A\cos(\omega_0 t + \phi) + A\dot{\phi}\cos(\omega_0 t + \phi). \tag{8}$$

If the zeroth order approximation is imposed (*i.e.*, no damping and no nonlinearities in the differential equation) then A and ϕ are constants and the first derivative of v with respect to time is given by

$$\dot{v} = \omega_0 A\cos(\omega_0 t + \phi). \tag{9}$$

Imposing this restriction on (8), the condition is obtained that

$$\dot{A}\sin(\omega_0 t + \phi) + \dot{\phi}A\cos(\omega_0 t + \phi) = 0. \tag{10}$$

Although this condition may seem somewhat arbitrary, Kryloff and Bogoluiboff[4] have shown that this method of analysis yields results correct to the second order in ϵ, where in this instance ϵ measures the departure of the nonlinear differential equation from the corresponding equation with the nonlinear terms absent. Since it was assumed that the elements under consideration were quasi-linear, which means that ϵ is small, this method of solution will, in general, yield results that are an accurate approximation to the exact solution.

Differentiation of (9) with respect to time and using the condition given in (10) results in

$$\ddot{v} = \omega_0\dot{A}/\cos(\omega_0 t + \phi) - \omega_0^2 A\sin(\omega_0 t + \phi). \tag{11}$$

Substitution of (9) and (11) in (2) and integration over a full period leads to

$$\omega\dot{A} + \frac{\omega\delta A}{2} = -\frac{1}{2\pi}\int_0^{2\pi}\frac{1}{C}f(v)\cos(\omega_0 t + \phi)d(\omega_0 t). \tag{12}$$

In carrying out this integration it is assumed that the amplitude of A remains constant during one period, which is consistent with the previous assumption that A is a slowly varying function of time.

[4] N. Kryloff and N. Bogoluiboff, "Introduction to Nonlinear Mechanics," Princeton University Press, Princeton, N. J.; 1943.

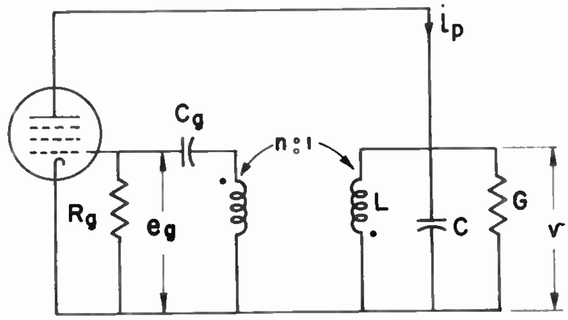

Fig. 2—Class-C oscillator employing a pentode as the active element.

If $f(v)$ is written as a power series and the relation in (3) is substituted in the series, the resulting expression is

$$f(v) = c_0 + c_1\sin(\omega_0 t + \phi) + c_2\sin 2(\omega_0 t + \phi) + \cdots \tag{13}$$

where the c_n's will be functions of A.

Substituting this expression in (12) and carrying out the integration gives

$$\dot{A} = -\frac{A}{2C}\left[G + \frac{c_1}{A}\right]. \tag{14}$$

Defining the effective electronic conductance as $G_{1e} = c_1/A$ results in

$$\dot{A} = -\frac{A}{2C}[G + G_{1e}] \tag{15}$$

which is identical to that obtained by Schaffner.[3]

Buildup in a Class-C Oscillator

Fig. 2 shows a typical Class-C oscillator in which the grid bias is derived from rectification in the grid circuit in the usual manner. In the following analysis it is assumed that R_g is very large so that the grid current is small and grid circuit losses may be neglected. This assumption is justified inasmuch as the oscillations will be building up at a relatively slow rate. Since there is only a small change in grid voltage from one period to the next, grid current will flow for only a small portion of the cycle and the grid circuit losses will be small when compared with the other losses in the oscillator. With the time constant of the grid circuit adjusted so that it is large, compared with a period of the natural frequency of the tank circuit, the bias on the tube will be very nearly equal to the peak value of the grid voltage.

In order to analyze this circuit, it is necessary to know the manner in which the plate current depends on the grid and plate voltages. It has been found that a good approximation of this dependence may be obtained if it is assumed that the plate current varies exponentially with grid voltage. This approximation is extremely good for pentodes, even when the grid is driven somewhat positive. The fact that the plate current is independent of the plate voltage indicates an infinite plate resistance, which is very nearly the case with pentode vacuum tubes.

On the basis of these assumptions the plate current may be written as

$$i_p = K_1 \exp (K_2 e_g) \tag{16}$$

where e_g is the instantaneous grid voltage and K_1 and K_2 are constants depending upon the tube geometry. If the tube characteristic is matched to (16) when the grid voltage is zero, the relation becomes

$$i_p = i_0 \exp (g_m e_g / i_0) \tag{17}$$

where i_0 and g_m are, respectively, the plate current and transconductance, evaluated when the grid voltage is zero.

In the case of a Class-C oscillator e_g is composed of the grid bias plus an alternating component. On the basis of the previous assumptions,

$$e_g = -n[A + A \sin (\omega_0 t + \phi)]. \tag{18}$$

The negative sign indicates the phase reversal which is required for oscillation and which is produced by the transformer.

Combining (17) and (18) and using the power-series expression for the exponential gives

$$i_p = i_0[I_0(kA) - 2I_1(kA) \sin (\omega_0(t + \phi) + 2I_2(kA) \cos 2(\omega_0 t + \phi) + \cdots] \exp (-kA) \tag{19}$$

where I_0, I_1, I_2, $\cdots$ are modified Bessel functions of the first kind and k is equal to ng_m/i_0.

Therefore, the effective electronic conductance is

$$G_{1e} = \frac{-2i_0 I_1(kA) \exp (-kA)}{A}. \tag{20}$$

Substitution of this value in (15) gives

$$\dot{A} = \frac{A}{2C}\left[\frac{2i_0 I_1(kA) \exp (-kA)}{A} - G\right]. \tag{21}$$

The condition $\dot{A} = 0$ is satisfied in either of two ways. One occurs when $A = 0$, and the other when

$$G_{1e} = G = \frac{2i_0 I_1(kA) \exp (-kA)}{A}. \tag{22}$$

The solution corresponding to $A = 0$ is unstable since any disturbance will cause the oscillation to build up. This is in agreement with the generally accepted theories that in order for an oscillator to start there must be some initial disturbance, such as a switching transient or thermal noise.

The other value for A represents the solution corresponding to the desired condition of sustained oscillation. Therefore, by knowing the circuit and tube parameters and by using (22), it is possible to predict the steady-state value of the tank voltage.

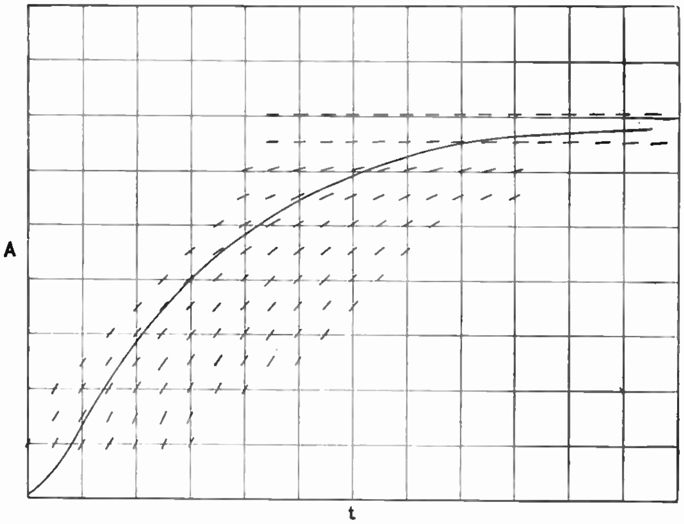

Fig. 3—Buildup of a Class-C oscillator.

The initial rate of rise (A small) is given by

$$\dot{A} = A(ng_m - G)/2C. \tag{23}$$

For oscillations that are to build up rapidly it is seen that it is desirable to make C as small as possible and ng_m as large as might be practical. The steady-state amplitude is given approximately by

$$A = i_0(2/\pi G^2 n g_m)^{1/3} \tag{24}$$

when $ng_m A/i_0 > 2$. As can be seen from (23) and (24), the same factors that lead to a rapid buildup of the oscillation limit the final amplitude of the signal.

It is possible to construct a plot of A as a function of time by using the method of isoclines. This method makes use of the fact that $\dot{A} = g(A)$ and enables one to make a plot such as in Fig. 3. The small line segments in the figure represent the value of $\dot{A}$ at arbitrarily selected values of A. If the value of A at $t = 0$ is known it is then possible to construct the actual build-up curve of the oscillator.

Simultaneous Oscillations

The preceding section has used the method of equivalent linearization to obtain the transient solution of a Class-C oscillator with one degree of freedom. This method of analysis can be extended to treat systems with two degrees of freedom. The method of analysis used in this section is essentially the same as that presented by Schaffner.[3]

The circuit shown in Fig. 4 has two natural frequencies and presents the possibility of simultaneous oscillations. The kind of oscillation which actually occurs depends on the relationship between the frequencies f_1 and f_2. If the ratio of the frequencies can be expressed by $f_1/f_2 = m/n$, where m and n are integers, the resulting oscillations are said to be synchronous. If the condition that m and n are integers cannot be met (*i.e.*, the frequencies are incommensurate), the resulting oscillations are called asynchronous.

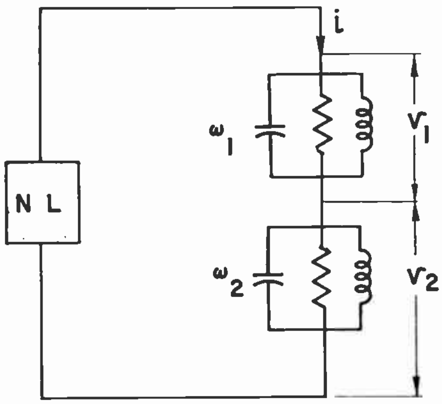

Fig. 4—Equivalent circuit of an oscillator having two degrees of freedom.

Prior to Skinner's[2] work it was generally believed that simultaneous asynchronous oscillations were impossible; however, Skinner was able to produce them but only in a system that was not self-starting. It will now be shown that an ordinary Class-C oscillator, having the characteristics described in the previous section, can give rise to self-starting asynchronous oscillations.

The voltages v_1 and v_2, as shown in Fig. 4, are assumed to be of the form:

$$v_1 = A \sin (\omega_1 t + \phi_1)$$
$$v_2 = B \sin (\omega_2 t + \phi_2) \tag{25}$$

where A, B, ϕ_1, and ϕ_2, are slowly varying functions of time as compared to $\sin (\omega_1 t)$ and $\sin (\omega_2 t)$. The current flowing is then given by

$$i = f(v_1 + v_2) = AG_{1e} \sin (\omega_1 t + \phi_1) + A\omega_1 C_{1e} \cos(\omega_1 t+\phi_1) + BG_{2e} \sin (\omega_2 t + \phi_1) + B\omega_2 C_{2e} \cos (\omega_2 t + \phi_2) \tag{26}$$

+harmonic terms and terms containing combinations of ω_1 and ω_2.

The terms C_{1e} and C_{2e} represent equivalent capacitances associated with the locking of the two frequencies and are only present for synchronous or nearly synchronous oscillations. For asynchronous oscillations both C_{1e} and C_{2e} will be zero. As was previously stated, if the sum of m and n is large, C_{1e} and C_{2e} and the resulting tendency toward synchronization will be small. This is due to the fact that, if $f(v)$ is expanded in a power series, the first term to contribute to C_{1e} or C_{2e} will be the $(n+m-1)$th term.

As was shown by Schaffner,[3] the following results hold for asynchronous oscillations,

$$\dot{A} = -\frac{A}{2C_1}[G_1 + G_{1e}]$$
$$\dot{B} = -\frac{B}{2C_2}[G_2 + G_{2e}]. \tag{27}$$

The assumptions made in determining these results were that the circuits were high-Q and that their resonant frequencies were not too close together.

For the Class-C oscillator having a grid time constant that is large compared to $1/(f_1-f_2)$, the grid bias will be equal to the sum of A and B. The assumption of a large grid time constant is made for two reasons. The first is that it simplifies the analysis, and the second is so that the case described in this paper will not be confused with another situation in which simultaneous oscillations were obtained by adjusting the grid time constant to be equal to the reciprocal of the difference frequency of the two modes.[5] When this is done the grid bias varies at the beat frequency of the two modes, causing a locking effect and making simultaneous oscillations possible.

Then, on the basis of the preceding assumptions,

$$i = i_0 \exp -k[A \sin (\omega_1 t + \phi_1) + B \sin (\omega_2 t + \phi_2) + A + B] \tag{28}$$

where the negative sign is again due to the phase reversal introduced by the transformer. The preceding expression can be written as

$$\begin{aligned} i = i_0[&I_0(kA) - 2I_1(kA) \sin (\omega_1 t + \phi_1) \\ &+ 2I_2(kA) \cos 2(\omega_1 t + \phi_1) + \cdots] \\ &\cdot[I_0(kB) - 2I_1(kB) \sin (\omega_2 t + \phi_2) \\ &+ 2I_2(kB) \cos 2(\omega_2 t + \phi_2) + \cdots] \\ &\cdot\exp - k[A + B]. \end{aligned} \tag{29}$$

Then, by definition and use of (27) one has

$$\dot{A} = \frac{A}{2C_1}\left[\frac{2i_0 I_0(kB)I_1(kA) \exp -k[A+B]}{A} - G_1\right]. \tag{30}$$

$$\dot{B} = \frac{B}{2C_2}\left[\frac{2i_0 I_0(kA)I_1(kB) \exp -k[A+B]}{B} - G_2\right]. \tag{31}$$

The ratio of the two preceding equations is

$$\frac{dA}{dB} = \frac{C_2}{C_1}\left[\frac{2i_0 I_0(kB)I_1(kA) \exp -k[A+B] - AG_1}{2i_0 I_0(kA)I_1(kB) \exp -k[A+B] - BG_1}\right]. \tag{32}$$

A graphical solution of (32) may be obtained by the method of isoclines as explained in the preceding section. In this case, values for both A and B are chosen and the resulting value of dA/dB is found from (32). Knowledge of the slope at a large number of points makes it possible to draw a family of curves corresponding to the solution of (32). The resulting family of curves, commonly called trajectories, is shown in Fig. 5. The arrows indicate the direction that the oscillation will follow with increasing time, the particular path depending on the initial values of A and B.

From Fig. 5 it can be seen that there is only one set

[5] L. Hazeltine, "Oscillating audion circuits," Proc. IRE, vol. 6, pp. 63–98; April, 1918.

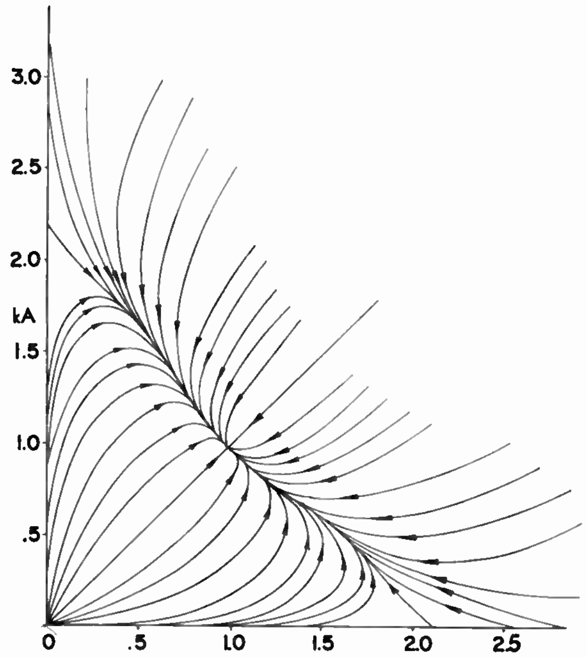

Fig. 5—Trajectories corresponding to the buildup in a Class-C oscillator having two degrees of freedom.

of values for A and B for which the system is at equilibrium and that this condition corresponds to stable asynchronous oscillations. The fact that the oscillations are self-starting can be seen from the following:

Solving (30) and (31) for small values of A and B gives

$$A = A_0 \exp\left(\frac{ng_m - G_1}{2C_1}\right)t$$

$$B = B_0 \exp\left(\frac{ng_m - G_2}{2C_2}\right)t \qquad (33)$$

where A_0 and B_0 are values, due to thermal noise or to transients, at $t=0$. Hence it is evident that both A and B subsequently increase with time and that the oscillations are therefore self-starting.

The steady-state amplitudes are given by

$$A = i_0[1/\pi n g_m]^{1/2}[G_2/G_1^3]^{1/4}$$

$$B = i_0[1/\pi n g_m]^{1/2}[G_1/G_2^3]^{1/4}. \qquad (34)$$

Discrimination

When two signals, having unrelated frequencies, pass through an amplitude limiter the relative transmission of the larger of the signals is usually greater than that of the smaller signal. This property of nonlinear elements has been termed discrimination[6,7] (*i.e.*, the larger signal discriminates against the smaller). The limiting stages in fm receivers are an excellent example of this phenomenon. In these stages the unwanted noise signal is reduced as compared to the desired signal.

[6] K. Amo, "Signal to Noise Discrimination in Amplitude Limiters," Stanford University, Stanford, Calif., Electronics Res. Lab., Tech. Rep. No. 17 (Nonr 22510); August 2, 1954.

[7] W. Edson, "Frequency memory in multimode oscillators," IRE Trans. on Circuit Theory, vol. CT-2, pp. 58–66; March, 1955.

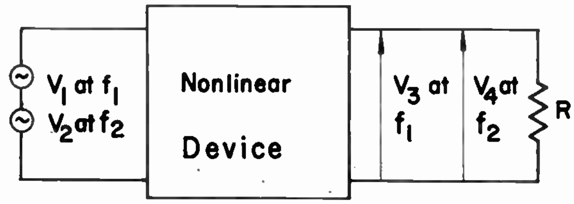

Fig. 6—Circuit illustrating the quantities used in the definition of discrimination.

A precise definition of discrimination can be made with the aid of the quantities shown in Fig. 6. The relative transmission at each frequency is defined by

$$\gamma_1 = V_3/V_1$$

$$\gamma_2 = V_4/V_2. \qquad (35)$$

It is convenient to define discrimination as the logarithm of the ratio of the relative transmissions.

$$\text{Discrimination} \equiv \underset{V_1>V_2}{D} \equiv 20 \log_{10} (\gamma_1/\gamma_2). \qquad (36)$$

The restriction that $V_1 > V_2$ is chosen so that the resulting discrimination will be positive when the gain of the larger signal is greater than the gain of the smaller one. This is the most common situation for nonlinear devices.

Fig. 7 shows a simple diode clipper circuit and the resulting discrimination characteristic, which was calculated subject to the condition that $V_2 \ll V_1$. The discrimination is zero when $|V_1| < |V_0|$, since the diodes have not yet begun to conduct and the voltage waveform of V_3 is identical in character to that of V_1. When V_1 is increased, the diodes conduct during the portion of the cycle when the magnitude of V_1 is greater than the magnitude of V_0, which results in the output voltage, V_3, being a clipped sine wave. This produces a positive discrimination which increases as V_1 is increased and approaches a maximum of 6 db.

In order to see what part discrimination plays in oscillators, the equivalent circuit, shown in Fig. 8, is constructed. In this circuit the functions normally performed by the active element are divided into two parts. One part is linear amplification and the other is the nonlinear function of amplitude limiting.

If the frequency-determining elements comprise a network having two zeros of transmission at two unrelated frequencies, the usual result is that only one of these frequencies will be present in the steady state. When the oscillations begin, the amplitude of one of the modes will be slightly larger than the other. Upon passing through the usual type of nonlinear element the relative transmission of the larger signal will be greater, and it will therefore have a larger rate of growth than the

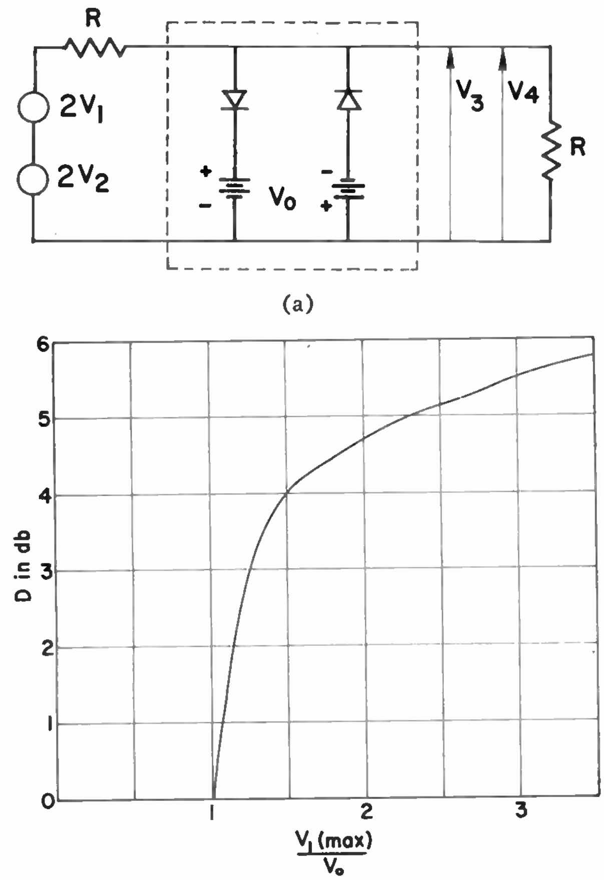

Fig. 7—(a) Diode clipper circuit. (b) Resulting discrimination characteristic.

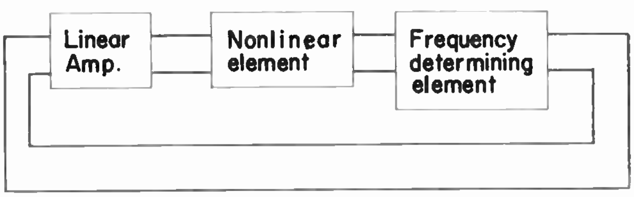

Fig. 8—Equivalent circuit of an oscillator.

smaller signal. Therefore, once one mode obtains the initial advantage it will dominate throughout the entire build-up process, continually discriminating against the smaller signal until, at equilibrium, it is the only signal present.

It is enlightening to examine the opposite situation where the relative transmission to the smaller signal is greater than that to the larger one. For this case the discrimination of the device producing the phenomenon will be negative. One such device is a Class-C amplifier where it is well known that the index of amplitude modulation of a modulated carrier is increased by passage through the amplifier.[8]

If a device having a negative discrimination characteristic is used for the nonlinear element in the feedback loop in Fig. 8, the loop gain of the smaller signal will be greater than that of the larger one. Since there is more gain to the smaller signal, its rate of growth will be greater and it will increase in amplitude at the expense of the larger signal. As the amplitude of the small signal increases, its rate of growth will decrease until it is building up at the same rate as the other mode. Now both signals will build up simultaneously until their equilibrium positions are reached.

The difficulty in achieving a self-starting asynchronous oscillator is in obtaining a nonlinear device which limits and yet exhibits the phenomenon of negative discrimination for all values of the applied signal. These conditions must be met in order that the only stable state of operation will be that when both signals are present. That these requirements are fulfilled by the Class-C oscillator, previously described, will now be shown.

Making the same assumptions as before, the plate current is given by (28). The relative transmissions are then:

$$\gamma_A = RG_{1e}$$
$$\gamma_B = RG_{2e} \tag{37}$$

where R is the plate load resistance.

The discrimination then becomes

$$\underset{A>B}{D} = 20 \log_{10}\left[\frac{B}{A}\,\frac{I_0(kB)}{I_1(kB)}\,\frac{I_1(kA)}{I_0(kA)}\right]. \tag{38}$$

At low levels the discrimination is zero. Therefore, the relative transmissions will be equal, allowing both signals to build up simultaneously. At large signals

$$\underset{A>B}{D} = 20 \log_{10} (B/A). \tag{39}$$

From the preceding relation it can be seen that, at high levels, the transmission to the smaller signal is greater than that to the larger one. This is the necessary condition of negative discrimination.

Fig. 9 shows a plot of A vs B, where the contours are constant values of discrimination. As can be seen, the discrimination is less than or equal to zero for all values of A and B.

The concept of discrimination can be extended to include two-terminal devices. In this case the discrimination turns out to be

$$\underset{V_1>V_2}{D} = 20 \log_{10} (G_{1e}/G_{2e}) \tag{40}$$

where G_{1e} and G_{2e} are the equivalent electronic conductances previously defined.

Experimental Results

The circuit shown in Fig. 10 was constructed in order to verify the predictions of the foregoing analysis. A 6AK5 pentode, V_1, was chosen for the Class-C oscilla-

[8] B. S. Glasgow, "Principles of Radio Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., 1st ed., p. 289; 1936.

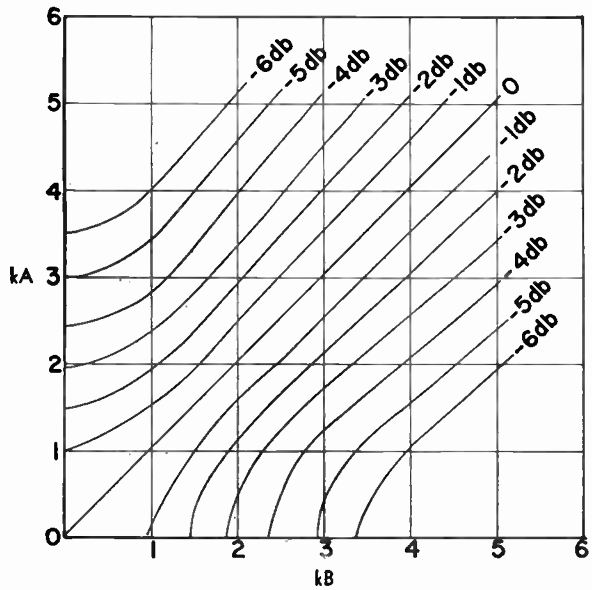

Fig. 9—Contours of constant discrimination for the Class-C amplifier.

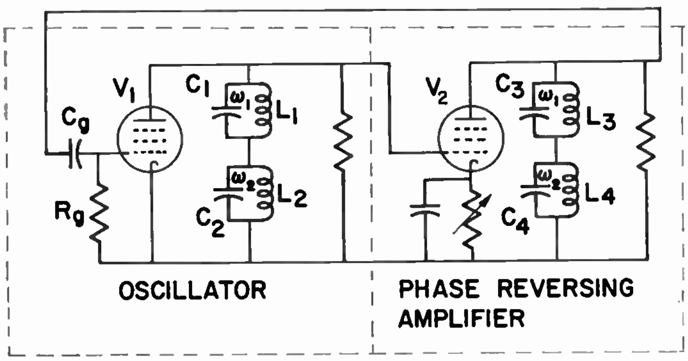

Fig. 10—Experimental circuit used to test for simultaneous asynchronous oscillations.

tor, because of its high transconductance and stability. A second tube of the same type was used to supply the required phase reversal and necessary gain.

The elements shown in Fig. 10 were adjusted so that the loop gains to both frequencies were approximately equal. The two tuned circuits in the phase-reversing amplifier were heavily loaded and therefore did not contribute to determining the frequencies of oscillation. The frequencies at which the oscillations took place were determined solely by the elements L_1, C_1, L_2, and C_2.

In adapting the preceding analysis to this circuit it was assumed that the only nonlinearity present was due to the first tube and that the second tube was essentially a linear amplifier. However, the phase reversing amplifier does exhibit some degree of nonlinearity, as is always the case, and had some effect on the over-all operation of the oscillator. In its normal mode of operation (*i.e.*, constant bias and small signal) the nonlinear effects result in a small amount of positive discrimination. This will tend to reduce the effectiveness of the Class-C stage to some extent. However, since the nonlinearities involved in the second tube, when properly adjusted, were small, they did not appreciably affect the over-all operation of the oscillator.

The variable resistor in the cathode lead of V_2 was used to vary the loop gain of the oscillator. When the cathode bias on V_2 was zero, oscillations existed at only one frequency. This was due to the fact that the loop gain was so high that positive-grid clipping occurred in V_2. Such peak-clipping introduces positive discrimination which tends to cancel the negative discrimination produced by V_1. When the net discrimination around the loop is positive, oscillations can exist at only one frequency.

When the loop gain was decreased, by increasing the cathode bias on V_2, simultaneous oscillations occurred. These oscillations were stable and had frequencies of 869.5 kc and 946.5 kc. Varying the grid time constant over the range from 0.5 μsec to 50 μsec had no appreciable effect on the frequencies and there was not any apparent locking effect when the grid time constant was equal to the reciprocal of the difference frequency.

It should be noted that the ratio of the two frequencies is very close to the ratio of 12:11. It was difficult to prove, experimentally, that they were asynchronous. However, calculation of the equivalent capacitances, C_{1e} and C_{2e}, show that they are extremely small compared with the capacitances in the tuned circuit. The values of the equivalent capacitances would be given by

$$C_{1e} = \frac{4I_{12}(kB)I_{11}(kA)}{\omega_1}$$

$$C_{2e} = \frac{4I_{12}(kA)I_{11}(kB)}{\omega_2}$$

if the oscillations were synchronous. Since both I_{11} and I_{12} are of the order of 10^{-4} for $kA \approx 4$, these capacitances will be extremely small compared with capacitances of the order of 1000 $\mu\mu$f which were in the tuned circuits. Therefore, it is improbable that any synchronization was present. Due to these results, it is felt that the oscillations present were truly asynchronous.

A relevant article by Wolfgang Feist[9] has just come to the authors' attention. This article describes some attempts at building a stable low-frequency oscillator by detecting the beat frequency of two high-frequency oscillations. In order that the frequencies of the two high-frequency oscillations should vary in the same manner with changes in various parameters (*e.g.*, temperature, supply voltages, etc.), the negative resistance sources for both frequencies should be derived from the same active element. These requirements make it evident that simultaneous asynchronous oscillations are necessary.

[9] W. Feist, "Über die gleichzeitige Erzeugung zwier Schwengungen in einem Oszillator und die Konstanz der Differenz-frequenz," *Nachrtech. Z.*, vol. 5, pp. 215–222; May, 1957.

The circuit arrangement used by Feist was almost identical with that shown in Fig. 10; the only differences were that he used tetrodes instead of pentodes and series resonant frequency determining circuits instead of parallel ones. It is interesting to note that in all of the tests he used large grid-leak resistors and arranged the grid time constant so that it was always greater than the reciprocal of the difference frequency. The foregoing analysis applies directly to his circuit since the dependence of the plate current on the grid voltage is approximately the same for tetrodes as it is for pentodes.

When using lumped elements for the frequency determining circuits he obtained many different sets of simultaneous oscillations. One of these sets had frequencies of 104 kc and 87 kc. In order for this set to have been synchronous it would have been necessary for m and n to have had values of 51 and 61. These large values for m and n lead to the conclusion that the oscillations were asynchronous.

More conclusive proof that the oscillations were asynchronous was obtained when the lumped element circuits were replaced by quartz crystals. In this case both frequencies were approximately equal to 130 kc, the difference between them being only 67 cycles. In this instance the corresponding values of m and n would have to be so large that synchronization between the two frequencies would be impossible. It is felt that the theory presented here adequately explains the experimental results obtained by Feist.

Conclusion

It has been shown that self-starting asynchronous oscillations are possible in Class-C oscillators using pentode vacuum tubes. The occurrence of these oscillations is predicted by the use of the method of equivalent linearization. They are also qualitatively explained by the phenomenon of negative discrimination, which seems to be an extremely useful concept for predicting the behavior of nonlinear devices. While a two-tube circuit was used for the experimental test, the conditions were such that one may confidently predict that the same sort of behavior will occur in the usual one-tube circuit.

The reason that simultaneous asynchronous oscillations have been observed in contrast to ordinary experience is attributed to the fact that a pentode rather than a triode was used and that the grid leak resistance was sufficiently high so that the grid current was negligible. The fact that the grid current was extremely small insures that the limiting was due to a shift in the operating point rather than to conductive loading on the positive peaks of the voltage wave. Clipping of the positive peak of the voltage wave introduces some positive discrimination, and if this positive discrimination is enough to overcome the negative discrimination produced by the characteristics of the vacuum tube, it will be impossible for simultaneous asynchronous oscillations to exist.

Theoretical Diversity Improvement in Frequency-Shift Keying*

JOHN N. PIERCE†

Summary—**Methods of using diversity to improve frequency-shift keyed receptions in the presence of Rayleigh fading are analyzed. In the absence of prior information about signal amplitude and phase, square-law combination is optimum; the error rate for this combination method has been found. If signal amplitude and phase are exactly known prior to reception of the signal, coherent combination and detection are optimum; at low error probability this yields only a 3-db improvement. Nonoptimum switch diversity yields only slightly less diversity gain than square-law combination. For dual diversity, correlation of the fading on the separate antennas does not give a large loss if the correlation coefficient is moderate. Correlated noise yields a similar small loss.**

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Introduction

The advantages of diversity reception in increasing the reliability of transmission in the presence of randomly fading signals have been known for some time. Several investigators[1–4] have demonstrated the gain realizable from diversity in the reception of

[1] Z. Jelonek and E. Fitch, "Diversity reception," *Wireless Eng.*, vol. 24, pp. 54–62; January, 1947.

[2] J. L. Glaser and S. H. Van Wambeck, "Experimental evaluation of diversity receiving systems," PROC. IRE, vol. 39, pp. 252–255; March, 1951.

[3] S. H. Van Wambeck and S. H. Ross, "Performance of diversity receiving systems," PROC. IRE, vol. 39, pp. 256–264; March, 1951.

[4] R. E. Lacy, M. Acker, and J. L. Glaser, "Performance of space and frequency diversity receiving systems," 1955 IRE CONVENTION RECORD, pt. 2, pp. 148–152.

continuous signals by commonly used combining techniques. Kahn[5] and Brennan[6] derived optimum combination of signals; Staras[7] determined the gain realizable with independently fading Rayleigh distributed signals using this optimum combination method.

In the field of binary transmissions by frequency-shift keying, Montgomery[8] demonstrated the realizable diversity gain using several methods of signal combination, both with two-filter and limiter-discriminator detection. More recently, Law[9] did valuable work in determining the optimum method of coherent combination of signals for detection of fsk transmissions, together with error rates when a memory of the signal strength is available.

It is the purpose here to determine optimum methods of diversity combination when memory of the signal strength is not available. We will then compare the error probability realized with this system to that obtained with the optimum coherent system, and to that obtained by using the familiar nonoptimum "switch diversity." Finally, the perturbing effects of correlated signal or noise on a system designed to be optimum for independent signals and noises will be analyzed. Ideal cross-correlation filtering shall be considered throughout rather than limiter-discriminator detection.

Basic Assumptions

The object here is to determine optimum combination methods and error rates for diversity reception of frequency-shift keyed transmissions in the presence of Rayleigh fading and white noise. Several basic assumptions are made: the signal amplitude and phase are constant during each received baud; the noise in each receiver is independent and has the same rms value; the noise in mark and space channels is also independent. In the latter part of the discussion the perturbing effects of correlated noise or signal will be investigated. Cross-correlation detection is presumed—frequently referred to as baud-synchronous or predicted wave detection. The samples from the mark and space channels are used in the following analysis.

Reiger[10] gives four results that will be pertinent: given a signal power S, baud length T, and receiver noise power of n_0 watts in a 1-cycle band, the probability density functions for the envelope samples in the mark and space channels when a mark is transmitted are

$$\begin{aligned} p(w)dw &= \frac{wdw}{n_0}\exp\left(-\frac{w^2+2ST}{2n_0}\right)I_0\left(\frac{w\sqrt{2ST}}{n_0}\right) \\ p(z)dz &= \frac{zdz}{n_0}\exp\left(-\frac{z^2}{2n_0}\right), \end{aligned} \tag{1}$$

where w is the mark filter sample and z, the space filter sample. The probability density functions for the inphase samples from the two channels when a mark is transmitted are

$$\begin{aligned} p(u)du &= \frac{du}{\sqrt{2\pi n_0}}\exp\left[-\frac{(u-\sqrt{2ST})^2}{2n_0}\right] \\ p(x)dx &= \frac{dx}{\sqrt{2\pi n_0}}\exp\left[-\frac{x^2}{2n_0}\right], \end{aligned} \tag{2}$$

with u and x the mark and space filter samples. The probability of error in a baud with envelope detection is

$$p(e) = \frac{1}{2}\exp\left(-\frac{ST}{2n_0}\right), \tag{3}$$

and the probability of error in a baud with coherent detection is

$$p(e) = \frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/2n_0}}^{\infty}\exp(-t^2)dt. \tag{4}$$

Decision Process

Given a certain amount of prior statistical information, it is a straightforward job to determine the decision rule for deciding whether mark or space was transmitted. We first form the joint probability density function (pdf) of the samples from the mark and space channels of the total number of receivers under the assumptions of transmitted mark and transmitted space. If there are M receivers, these will be $2M$-dimensional functions. Calling the set of sample values A, we form the conditional density functions, $p(A \mid \text{mark})$, $p(A \mid \text{space})$. If mark and space are equally likely to have been transmitted, the average pdf of the set A is

$$p(A) = (1/2)p(A \mid \text{mark}) + (1/2)p(A \mid \text{space}).$$

Then the probability that a mark or space was transmitted, given the sample set A, is

$$\begin{aligned} p(\text{mark} \mid A) &= (1/2)p(A \mid \text{mark})/p(A), \\ p(\text{space} \mid A) &= (1/2)p(A \mid \text{space})/p(A). \end{aligned}$$

We will decide that a mark is transmitted if $p(\text{mark}\mid A)$ is larger than $p(\text{space}\mid A)$ or, more simply, extracting the common factors, if $p(A\mid\text{mark})$ is larger than $p(A\mid\text{space})$. The decision rule will be an arithmetic rule for comparing the two probabilities given the set

[5] L. Kahn, "Ratio squarer," Proc. IRE, vol. 42, p. 1704; November, 1954.

[6] D. G. Brennan, "On the maximum signal-to-noise ratio realizable from several noisy signals," Proc. IRE, vol. 43, p. 1530; October, 1955.

[7] H. Staras, "The statistics of combiner diversity," Proc. IRE, vol. 44, pp. 1057–1058; August, 1956.

[8] G. F. Montgomery, "Message error in diversity frequency-shift reception," Proc. IRE, vol. 42, pp. 1184–1187; July, 1954.

[9] J. W. Allnatt, E. D. J. Jones, and H. B. Law, "Frequency diversity in the reception of selectively fading binary frequency-modulated signals," *Proc. IEE*, pt. B, vol. 104, pp. 98–110; March, 1957.

H. B. Law, "The signal/noise performance rating of receivers for long-distance synchronous radiotelegraph systems using frequency modulation," *Proc. IEE*, pt. B, vol. 104, pp. 124–129; March, 1957.

——, "The detectability of fading radiotelegraph signals in noise," *Proc. IEE*, pt. B, vol. 104, pp. 130–140; March, 1957.

[10] S. Reiger, "Error probabilities of binary data transmission systems in the presence of random noise," 1953 IRE Convention Record, pt. 9, pp. 72–79.

of samples A. Having found the optimum decision rule, the error probability is found by routine statistical methods. Since the fsk channel is symmetric, it is sufficient to determine the error rate when mark is transmitted, mark and space error probabilities being identical.

We consider here two extreme cases of available *a priori* information. If no prior information is assumed about amplitude and phase of the several received signals, we work only with the statistical distribution of received powers which corresponds to Rayleigh fading. If signal amplitude and phase in both mark and space channels are known prior to making a decision, we may use a form of synchronous or in-phase combination and detection, a process which will yield a different decision rule, combination method, and lower error rate. For a particular system where the available *a priori* information lies between the two extremes it would be possible to find an optimum decision rule which would yield an error rate intermediate between the two cases to be considered. For simplicity the two extremes will be referred to as noncoherent and coherent diversity combination.

Noncoherent Diversity

Decision Rule

We assume no information other than the density function of the received power which, for Rayleigh distribution of amplitudes, is given by $p(S)dS = \exp(-S/S_0)dS/S_0$. S_0 is the long-time average received power. The mark channel envelope distribution (1), which is a function of the instantaneous power S, must be averaged over all possible values:

$$p(w) = \int_S p(w \mid S)p(S)dS$$
$$= \int_0^\infty \frac{w}{n_0} \exp\left(-\frac{w^2 + 2ST}{2n_0}\right) I_0\left(\frac{w\sqrt{2ST}}{n_0}\right) \cdot \exp\left(-\frac{S}{S_0}\right)\frac{dS}{S_0}$$
$$= \frac{w}{n_0 + S_0T} \exp\left(-\frac{w^2}{2n_0 + 2S_0T}\right). \tag{5}$$

The space sample distribution remains the same as that given by (1) since it is independent of S.

If we now denote the samples from the mth receiver by a subscript m on the samples (there being M receivers altogether) the joint pdf of the samples is

$$p(W, Z \mid \text{mark}) = \prod_{m=1}^{M} \frac{w_m}{n_0 + S_0T} \cdot \exp\left(-\frac{w_m^2}{2n_0 + 2S_0T}\right) \cdot \frac{z_m}{n_0} \exp\left(-\frac{z_m^2}{2n_0}\right)$$

$$p(W, Z \mid \text{space}) = \prod_{m=1}^{M} \frac{z_m}{n_0 + S_0T} \cdot \exp\left(-\frac{z_m^2}{2n_0 + 2S_0T}\right) \cdot \frac{w_m}{n_0} \exp\left(-\frac{w_m^2}{2n_0}\right), \tag{6}$$

where (W, Z) denotes the set of samples $(w_1, \cdots, w_M, z_1, \cdots, z_M)$. By straightforward algebraic manipulation and extraction of all common factors from the two pdf's, we find

$$p(W, Z \mid \text{mark}) > p(W, Z \mid \text{space}) \text{ if } \sum_{m=1}^{M} w_m^2 > \sum_{m=1}^{M} z_m^2. \tag{7}$$

Thus the optimum decision rule, as given by (7), indicates a square-law combination of the mark samples and space samples, the two sums being compared to determine the transmitted symbol.

Probability of Error

It is now a simple task to compute the probability of error. Let $v = w^2/2n_0$, $y = z^2/2n_0$. The normalizing factor $1/2n_0$ has been inserted for convenience. The new variables have density functions

$$p(v)dv = \frac{dv}{1 + S_0T/n_0} \exp\left(-\frac{v}{1 + S_0T/n_0}\right)$$
$$p(y)dy = dy \exp(-y).$$

Let $V = \Sigma v_m$, $Y = \Sigma y_m$. The density functions of V and Y are found by taking the Mfold convolution over the identical component distributions. The result is well known:

$$p(V)dV = \frac{V^{M-1}dV}{(1 + S_0T/n_0)^M(M-1)!} \exp\left(-\frac{V}{1 + S_0T/n_0}\right)$$
$$p(Y)dY = \frac{Y^{M-1}dY}{(M-1)!} \exp(-Y). \tag{8}$$

Since V and Y are proportional to the sums of the squares, an error occurs (when a mark is transmitted) every time Y is larger than V. The error probability is thus

$$p(e) = \int_{Y>V}\int p(V)p(Y)dVdY$$
$$= \int_0^\infty dV \frac{V^{M-1}}{(1 + S_0T/n_0)^M(M-1)!} \cdot \exp\left(-\frac{V}{1 + S_0T/n_0}\right) \int_V^\infty dY \frac{Y^{M-1}}{(M-1)!} \exp(-Y)$$
$$= \int_0^\infty dV \frac{V^{M-1}}{(1 + S_0T/n_0)^M(M-1)!} \cdot \exp\left(-\frac{V}{1 + S_0T/n_0}\right) \sum_{m=0}^{M-1} \frac{V^m}{m!} \exp(-V)$$

$$= \sum_{m=0}^{M-1} \frac{1}{m!(M-1)!(1+S_0T/n_0)^M} \int_0^\infty dV\, V^{m+M-1}$$

$$\cdot \exp\left(-\frac{2+S_0T/n_0}{1+S_0T/n_0} V\right)$$

$$= \frac{1}{(2+S_0T/n_0)^M} \sum_{m=0}^{M-1} \frac{(m+M-1)!}{m!(M-1)!} \left(\frac{1+S_0T/n_0}{2+S_0T/n_0}\right)^m.$$

If we write $(1+S_0T/n_0)/(2+S_0T/n_0) = 1-1/(2+S_0T/n_0)$ and make a binomial expansion of each term in the sum, we get the more convenient form of the result by rearranging the terms

$$p(e) = \sum_{m=0}^{M-1} \frac{(2M-1)!(-1)^m}{(M-1)!(M-1-m)!\,m!(M+m)} (2+S_0T/n_0)^{-M-m}. \quad (9)$$

At large signal to noise ratios the leading term predominates and we have, to a good approximation,

$$p(e) \cong \frac{(2M-1)!}{(M-1)!M!} (S_0T/n_0)^{-M}, \quad S_0T \gg n_0. \quad (10)$$

In Fig. 1, the error rate from (9) is plotted as a function of S_0T/n_0 with the number of receivers in the diversity system as a parameter.

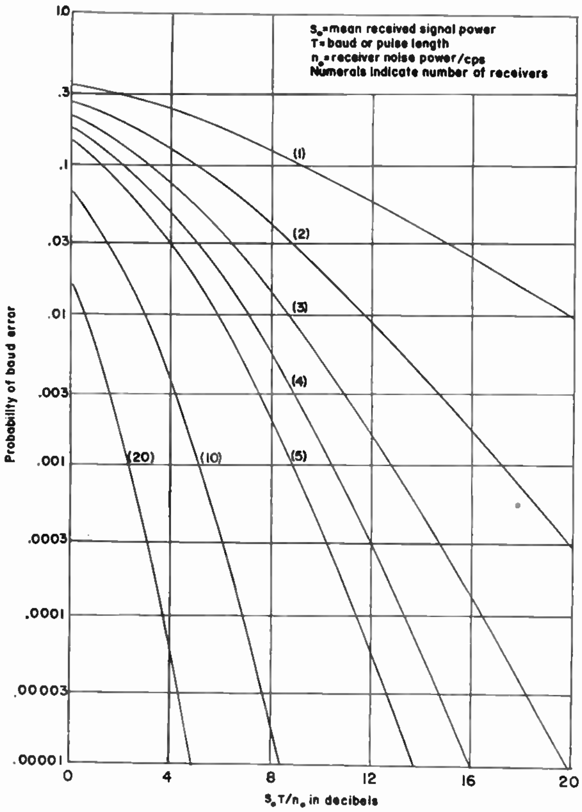


Fig. 1—Square-law combination diversity.

Coherent Diversity

Decision Rule

The phase and instantaneous signal amplitude are presumed to be known exactly before reception of a baud. It will be assumed that the instantaneous signal power is the same in the mark and space channels. Since phase is known, we work with the probability density functions for the in-phase samples given by (2). Subscripts are again added to denote the output from a particular receiver. The instantaneous power at each receiver terminal is also different and will be denoted by S_m. (We presume here that the power received on either mark or space frequencies is identical at each receiver terminal.) We can write the joint pdf's by inspection. Letting (U, X) be the set of samples $(u_1 \cdots, u_M, x_1 \cdots x_M)$,

$$p(U, X \mid \text{mark}) = \prod_{m=1}^{M} \frac{1}{2\pi n_0} \exp\left[-\frac{(u_m - \sqrt{2S_mT})^2 + x_m^2}{2n_0}\right]$$

$$p(U, X \mid \text{space}) = \prod_{m=1}^{M} \frac{1}{2\pi n_0} \exp\left[-\frac{(x_m - \sqrt{2S_mT})^2 + u_m^2}{2n_0}\right]. \quad (11)$$

By simple operations we extract the decision rule

$$p(U, X \mid \text{mark}) > p(U, X \mid \text{space})$$

$$\text{if } \sum_{m=1}^{M} \sqrt{S_m}u_m > \sum_{m=1}^{M} \sqrt{S_m}x_m. \quad (12)$$

In words, the samples from each receiver are weighted by the expected signal amplitude and the sums of the weighted mark samples and space samples compared to determine the transmitted symbol.

Probability of Error

Let $\sigma = \Sigma S_m$, a constant for any particular decision but having a statistical distribution as a function of time, and let $v_m = (S_m/\sigma)^{1/2}u_m$, $y_m = (S_m/\sigma)^{1/2}x_m$. Then

$$p(v_m)dv_m = \frac{dv_m}{\sqrt{2\pi n_0 S_m/\sigma}} \exp\left[-\frac{(v_m - S_m\sqrt{2T/\sigma})^2}{2n_0S_m/\sigma}\right]$$

$$p(y_m)dy_m = \frac{dy_m}{\sqrt{2\pi n_0 S_m/\sigma}} \exp\left[-\frac{y_m^2}{2n_0S_m/\sigma}\right].$$

Now let $V = \Sigma v_m$, $Y = \Sigma y_m$. Since the v_m, y_m are normally distributed variables, their sums will also have a normal distribution with means and variances equal to the sums of the means and variances of the component distributions, so that

$$p(V)dV = \frac{dV}{\sqrt{2\pi n_0}} \exp\left[-\frac{(V - \sqrt{2\sigma T})^2}{2n_0}\right]$$

$$p(Y)dY = \frac{dY}{\sqrt{2\pi n_0}} \exp\left[-\frac{Y^2}{2n_0}\right]. \quad (13)$$

Thus the optimum decision rule with diversity gives distributions which are a function of the sum of the instantaneous powers and have the same form as (2).

(This is, as we would expect, identical to the result for combination of continuous signals as given by Kahn[5] and Brennan.[6]) We may therefore use the error probability as given by (4), replacing the power at one receiver by the sum of the powers. This is then averaged over the distribution of the sum of the powers to get the average error rate. The sum of these exponentially distributed powers yields a density function similar to (8) and identical to that given by Staras:[7]

$$p(\sigma)d\sigma = \frac{\sigma^{M-1}d\sigma}{S_0^M(M-1)!}\exp\left(-\frac{\sigma}{S_0}\right),$$

so that the error rate is given by

$$p(e) = \int_0^\infty p(\sigma)p(e\mid\sigma)d\sigma$$

$$= \int_0^\infty \frac{\sigma^{M-1}d\sigma}{S_0^M(M-1)!}\exp\left(-\frac{\sigma}{S_0}\right)\cdot\frac{1}{\sqrt{\pi}}\int_{\sqrt{\sigma T/2n_0}}^\infty \cdot\exp(-t^2)dt. \qquad (14)$$

The double integral in (14) can be readily evaluated but yields unwieldy expressions for the error rate. A more meaningful comparison with the noncoherent case can be obtained by finding the asymptotic expression for the error probability at large signal to noise ratios. By expanding the integrand in a Maclaurin series in $(1/S_0)$, the factor $[S_0^{-M}\exp(-\sigma/S_0)]$ becomes $[S_0^{-M}+\text{terms}$ of the order of $S_0^{-M-1}]$ so that for $S_0T\gg n_0$ we may use the leading term only and write

$$p(e) \sim \int_0^\infty \frac{\sigma^{M-1}d\sigma}{S_0^M(M-1)!}\cdot\frac{1}{\sqrt{\pi}}\int_{\sqrt{\sigma T/2n_0}}^\infty \exp(-t^2)dt$$

$$= \frac{1}{\sqrt{\pi}}\int_0^\infty \exp(-t^2)dt\int_0^{2n_0t^2/T}\frac{\sigma^{M-1}d\sigma}{S_0^M(M-1)!}$$

$$= \frac{1}{\sqrt{\pi}}\int_0^\infty \exp(-t^2)dt\cdot\frac{(2n_0t^2/S_0T)^M}{M!},$$

or

$$p(e) \sim \frac{(2M-1)!}{M!(M-1)!}(2S_0T/n_0)^{-M}, \quad S_0T/n_0 \gg 1. \qquad (15)$$

A comparison with (10) shows that at small probability of error the coherent combination and detection is equivalent to doubling the effective input power for the noncoherent combination.

As was pointed out earlier, systems using somewhat less of the prior information will yield an error rate intermediate between those of the coherent and noncoherent combination diversity. For instance, if received power but not phase is exactly known, an optimum decision rule could be found which would yield somewhat less than the 3-db gain of coherent over noncoherent combination and detection. It is worthwhile to note that errors in the prior information, which was assumed to be correct, can cause a rather large increase in the probability of error. As an extreme example, if coherent combination and detection of the several signals are employed, rapid fading which causes the phase and amplitude of the received signals to vary randomly from one baud to the next will result in an error probability of one half; the noncoherent diversity combination under the same conditions will still yield the theoretical diversity improvement.

Nonoptimum Combining

It is informative to evaluate the loss sustained when using some nonoptimum form of diversity combination or decision rule. A common type is the so-called "switch diversity" where only the receiver having the largest instantaneous signal input is used at any given instant. To determine the error probability we find the probability density function for the maximum of the M signal powers and average the error rate, (3), over this density function.

The probability that the signal power at a particular receiver is smaller than a given level S_L is

$$p(S_m < S_L) = \int_0^{S_L} p(S_m)dS_m = \int_0^{S_L}\exp(-S_m/S_0)dS_m/S_0$$

$$= 1 - \exp(-S_L/S_0).$$

The probability that all M signal powers are smaller than S_L is the Mth power of this:

$$p(S_1, S_2\cdots, S_m < S_L) = [1-\exp(-S_L/S_0)]^M.$$

The probability that all signals are smaller than the level is also the probability that the largest of them is smaller than the given level. This last expression is then the cumulative distribution function for the largest signal power. The corresponding density function is found by differentiating with respect to S_L and evaluating at S which we now use to denote the *largest* signal power. This gives for the pdf

$$p(S)dS = M[1-\exp(-S/S_0)]^{M-1}\exp(-S/S_0)dS/S_0. \qquad (16)$$

The error probability is then

$$p(e) = \int_0^\infty p(e\mid S)p(S)dS$$

$$= \int_0^\infty \frac{1}{2}\exp\left(-\frac{ST}{2n_0}\right)M\left[1-\exp\left(-\frac{S}{S_0}\right)\right]^{M-1}\cdot\exp\left(-\frac{S}{S_0}\right)\frac{dS}{S_0}.$$

If we let $t=1-\exp(-S/S_0)$, the integral becomes

$$p(e) = \int_0^1 \frac{M}{2}t^{M-1}(1-t)^{S_0T/2n_0}dt$$

$$= \frac{M}{2}\frac{\Gamma(M)\Gamma(S_0T/2n_0+1)}{\Gamma(M+S_0T/2n_0+1)}$$

$$= \frac{1}{2}\prod_{m=1}^{M}\frac{m}{m+(S_0T/2n_0)}, \qquad (17)$$

which at low probability of error is

$$p(e) \sim 2^{M-1}M!(S_0T/n_0)^{-M}. \tag{18}$$

The method used here for deriving (17) is identical to that used by Montgomery,[8] and the result is similar in form, although not identical, to his result for limiter-discriminator detection. The effective power loss at low probability of error is just the Mth root of the ratio of the two coefficients of $(S_0T/n_0)^{-M}$ given by the asymptotic expressions in (10) and (18). Table I gives this loss for various orders of diversity. For small numbers of receivers the loss is not great and considerations of ease of implementation, switching transients, or some other factor are probably more significant in determining whether or not to use the optimum square-law combination.

TABLE I

LOSS OF SWITCH DIVERSITY FROM SQUARE-LAW COMBINATION

M	2	3	4	5	6	8	10
db loss	0.62	1.27	1.85	2.36	2.87	3.63	4.20

NON-INDEPENDENT FADING

Complete statistical independence of the fading on separated antennas can rarely be achieved. It is therefore worthwhile to derive expressions for the error rate when correlated fading occurs. In the case of dual diversity an analytic expression may be obtained for square-law combination of the signals.

It is first necessary to find the error probability as a function of the two signal powers prior to averaging. In the equation pair (1) let $v_m = w_m^2/2n_0$, $y_m = z_m^2/2n_0$; then

$$p(v_m)dv_m = dv_m \exp(-v_m - S_mT/n_0)I_0(2\sqrt{v_mS_mT/n_0}),$$

$$p(y_m)dy_m = dy_m \exp(-y_m).$$

If we now let $V = v_1+v_2$, $Y = y_1+y_2$, the density function of Y is given by (8); the density function of V may be found by using Laplace transform techniques. We will use r as the variable of the transformation.

$$\begin{aligned}\mathcal{L}[p(v_m)] &= \int_0^\infty \exp(-rv_m)p(v_m)dv_m \\ &= \frac{1}{r+1}\exp\left[-\frac{S_mTr}{n_0(r+1)}\right] \\ \mathcal{L}[p(V)] &= \mathcal{L}[p(v_1)]\mathcal{L}[p(v_2)] \\ &= \frac{1}{(r+1)^2}\exp\left[-\frac{\sigma Tr}{n_0(r+1)}\right]\end{aligned}$$

where $\sigma = S_1+S_2$. The error probability is

$$\begin{aligned}p(e\,|\,\sigma) &= \int_{Y>V}\int p(V)p(Y)dVdY \\ &= \int_0^\infty p(V)dV\cdot\int_V^\infty Y\exp(-Y)dY \\ &= \int_0^\infty \mathcal{L}_V^{-1}\left\{\frac{1}{(r+1)^2}\exp\left[-\frac{\sigma Tr}{n_0(r+1)}\right]\right\} \\ &\quad\cdot(1+V)\exp(-V)dV \\ &= \left\{\left(1-\frac{d}{dr}\right)\frac{1}{(r+1)^2}\exp\left[-\frac{\sigma Tr}{n_0(r+1)}\right]\right\}_{r=1} \\ &= \left(\frac{1}{2}+\frac{1}{16}\frac{\sigma T}{n_0}\right)\exp\left(-\frac{\sigma T}{2n_0}\right), \text{ or}\end{aligned}$$

$$p(e\,|\,S_1,S_2) = \left(\frac{1}{2}+\frac{1}{16}\frac{S_1T+S_2T}{n_0}\right)\cdot\exp\left(-\frac{S_1T+S_2T}{2n_0}\right). \tag{19}$$

Eq. (19) must be averaged over the joint pdf of S_1, S_2. The joint Rayleigh distribution of amplitudes is well known; expressed in terms of power rather than amplitude, we have

$$p(S_1,S_2) = \frac{1}{S_0^2(1-\mu)}\cdot\exp\left[-\frac{S_1+S_2}{S_0(1-\mu)}\right]I_0\left[2\sqrt{\frac{\mu S_1S_2}{S_0^2(1-\mu)^2}}\right]. \tag{20}$$

The coefficient μ is the normalized power correlation coefficient. (The power correlation differs by at most 2.7 per cent from the envelope correlation.) We have then

$$p(e) = \iint p(e\,|\,S_1,S_2)p(S_1,S_2)dS_1dS_2.$$

Before substituting (19) and (26) in the integral we make the changes of variable $s_1 = S_1/S_0$, $s_2 = S_2/S_0$, $a = S_0T/n_0$ to simplify notation:

$$p(e) = \int_0^\infty\int\left(\frac{1}{2}+\frac{as_1}{16}+\frac{as_2}{16}\right)\cdot\exp\left[-\left(\frac{a}{2}+\frac{1}{1-\mu}\right)(s_1+s_2)\right]I_0\left[2\sqrt{\frac{\mu s_1s_2}{(1-\mu)^2}}\right]\frac{ds_1ds_2}{1-\mu}.$$

The symmetry in s_1, s_2 permits writing the sum of the three double integrals implicit in this equation as

$$p(e) = \int_0^\infty\left(\frac{1}{2}+\frac{as_1}{8}\right)\exp\left[-\left(\frac{a}{2}+\frac{1}{1-\mu}\right)s_1\right]ds_1\cdot\int_0^\infty\exp\left[-\left(\frac{a}{2}+\frac{1}{1-\mu}\right)s_2\right]\cdot I_0\left[2\sqrt{\frac{\mu s_1}{(1-\mu)^2}s_2}\right]\frac{ds_2}{1-\mu}$$

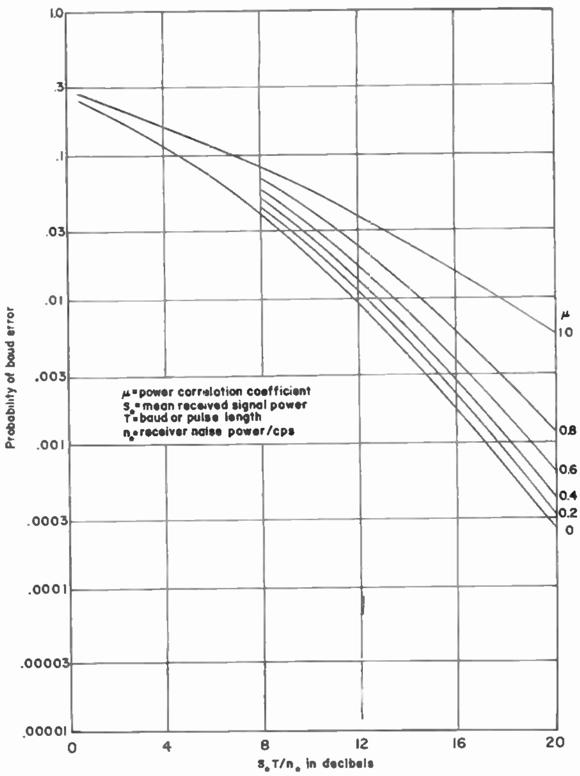


Fig. 2—Dual diversity with correlated fading.

$$= \int_0^\infty \left(\frac{1}{2} + \frac{as_1}{8}\right) \exp\left[-\left(\frac{a}{2} + \frac{1}{1-\mu}\right)s_1\right] ds_1$$

$$\cdot \frac{2}{a(1-\mu)+2} \exp\left[\frac{2\mu}{a(1-\mu)^2 + 2(1-\mu)} s_1\right]$$

$$= \frac{3(1-\mu)a^2 + 10a + 8}{[(1-\mu)a^2 + 4a + 4]^2},$$

so that in terms of the original notation

$$p(e) = \frac{3(1-\mu)(S_0T/n_0)^2 + 10(S_0T/n_0) + 8}{[(1-\mu)(S_0T/n_0)^2 + 4(S_0T/n_0) + 4]^2}. \quad (21)$$

The error rate is plotted in Fig. 2 with μ as a parameter. The curve for $\mu = 0$ is just the curve of Fig. 1 for two receivers. The curve for $\mu = 1$ shows a slight improvement over the curve of Fig. 1 for one receiver. This is due to the assumed lack of correlation of noises at the two receiver inputs. The results here are comparable to those given by Staras[11] for correlated continuous signals with almost the full diversity gain being realized even for relatively large values of the normalized correlation coefficient.

Correlated Noise

Frequently it may be the case that part of the noise in the several receivers comes from a common source and thus the total noise will be highly correlated. To obtain an estimate of the effect of this on square-law combination, we consider the extreme case of all the noise derived from a common source; the signal fading will be presumed independent at the several received inputs.

Let the instantaneous noise energy of all mark channel outputs be n_A and of the space channel outputs n_B. The two (uncorrelated) variables have identical distributions:

$$p(n_A) = (1/n_0) \exp(-n_A/n_0),$$
$$p(n_B) = (1/n_0) \exp(-n_B/n_0).$$

All space channel outputs have identical output samples, $z = \sqrt{2n_B}$; the mark channel sample density function can be found by recognizing that the assumption of Rayleigh fading of the signal implies normal distributions of in-phase and quadrature components of the signal. Thus we effectively interchange the role of signal and noise in (1) so that

$$p(w \mid n_A) = \frac{w}{S_0T} \exp\left(-\frac{w^2 + 2n_A}{2S_0T}\right) I_0\left(\frac{w\sqrt{2n_A}}{S_0T}\right) \quad (22)$$

is the required density function for all the mark channel samples. To perform the square-law combination we let $v = w^2/2n_0$, $y = z^2/2n_0$, and $a = S_0T/n_0$. Then at each receiver $y = n_B/n_0$, and v has the conditional density function

$$p(v \mid n_A) = \frac{1}{a} \exp\left(-\frac{v + n_A/n_0}{a}\right) I_0\left(2\sqrt{\frac{n_A/n_0}{a^2} v}\right).$$

Now let $V = \Sigma v_m$, $Y = \Sigma y_m$; if there are M receivers, $Y = Mn_B/n_0$; the density function of V can be found using Laplace transforms:

$$\mathcal{L}[p(v \mid n_A)] = \int_0^\infty \exp(-rv) p(v \mid n_A) dv_A$$

$$= \frac{1}{ar+1} \exp\left[-\frac{n_A r}{n_0(ar+1)}\right].$$

$$\mathcal{L}[p(V \mid n_A)] = \{\mathcal{L}[p(v \mid n_A)]\}^M$$

$$= \frac{1}{(ar+1)^M} \exp\left[-\frac{Mn_A r}{n_0(ar+1)}\right].$$

It is now permissible to integrate out the n_A and n_B dependencies.

$$\mathcal{L}[p(V)] = \int_0^\infty p(n_A) \mathcal{L}[p(V \mid n_A)] dn_A$$

$$= \frac{1}{(ar+1)^M} \int_0^\infty \frac{dn_A}{n_0} \exp\left(-\frac{n_A}{n_0}\right)$$

$$\cdot \exp\left[-\frac{Mr}{(ar+1)} \frac{n_A}{n_0}\right]$$

$$= (ar+1)^{1-M}(Mr + ar + 1)^{-1}.$$

[11] H. Staras, "Diversity reception with correlated signals," *J. Appl. Phys.*, vol. 27, pp. 93–94; January, 1956.

The probability density of Y may be written by inspection as

$$p(Y) = \frac{1}{M} \exp\left(-\frac{Y}{M}\right).$$

The error rate is given by

$$\begin{aligned} p(e) &= \int_0^\infty p(V)dV \int_V^\infty p(Y)dY \\ &= \int_0^\infty \mathcal{L}_V^{-1}[(ar+1)^{1-M}(Mr+ar+1)^{-1}]dV \\ &\quad \cdot \exp(-V/M) \\ &= [(ar+1)^{1-M}(Mr+ar+1)^{-1}]_{r=1/M} \\ &= [(a/M+1)^{1-M}(a/M+2)^{-1}] \end{aligned}$$

or, in terms of the original notation

$$p(e) = (S_0T/Mn_0 + 1)^{1-M}(S_0T/Mn_0 + 2)^{-1} \quad (23)$$

which at low probability of error (large signal to noise ratio) is approximately

$$p(e) \sim M^M(S_0T/n_0)^{-M}, \quad S_0T \gg Mn_0. \quad (24)$$

The decibel loss at low error rate caused by the noise being completely correlated is given in Table II. The loss in decibels at a fixed error rate, when only a part of the noise arises from a common source, will be somewhat less than the tabulated values. In this connection it may be pointed out that switch diversity is not subject to perturbing effects due to correlated noise since only one receiver is operative at any instant.

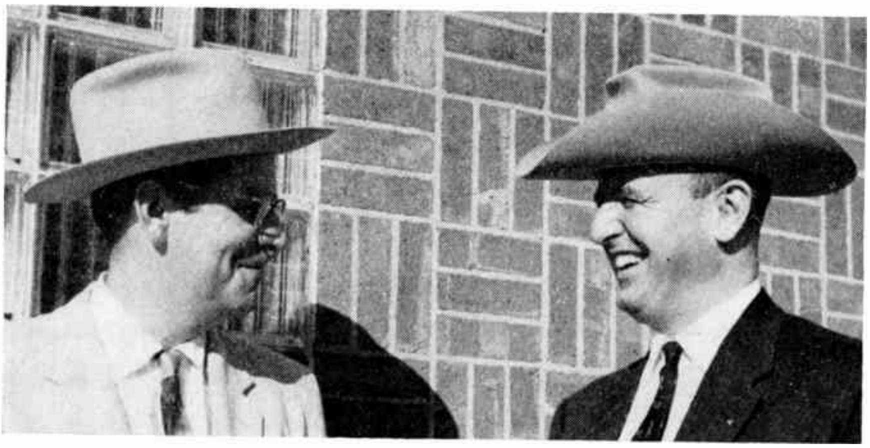
TABLE II
LOSS CAUSED BY COMPLETELY CORRELATED NOISE

M	2	3	4	5	6	8	10
db loss	0.62	1.44	2.16	2.79	3.34	4.27	5.03

ACKNOWLEDGMENT

Mrs. Margaret D. Hill performed part of the hand calculation and prepared the two figures.

Correspondence



The President Smiles*

We, the IRE members of Texas, found the January, 1958 PROCEEDINGS to be of poor taste and offensive in one respect. This is, the picture of our respected President for 1958 on page 2.

We have taken steps to correct this error. Enclosed is a picture of Mr. Fink which will meet with the approval of all Texans. No man can be appreciated in Texas unless he wears a "ten-gallon" hat or owns a few hundred oil wells.

In the picture, Floyd Crum, (left), Chairman of the Beaumont-Port Arthur Section, has just presented Mr. Fink with a "ten-gallon Texas hat," thus protecting the dignity due Mr. Fink by all Texans.

The smiling face of Mr. Fink, in contrast with the glum and despondent picture that was presented in the January issue of PROCEEDINGS, attests to the service which the Beaumont-Port Arthur Section has rendered the IRE.

WES ECKLES
Fellowship Chairman
Beaumont-Port Arthur Section

* Received by the IRE, January 22, 1958.

WWV Standard Frequency Transmissions*

Since October 9, 1957, the National Bureau of Standards radio stations WWV and WWVH have been maintained as constant as possible with respect to atomic frequency standards maintained and operated by the Boulder Laboratories, National Bureau of Standards. On October 9, 1957, the USA Frequency Standard was 1.4 parts in 10^9 high with respect to the frequency derived from the UT 2 second (provisional value) as determined by the U. S. Naval Observatory. The atomic frequency standards remain constant and are known to be constant to 1 part in 10^9 or better. The broadcast frequency can be further corrected with respect to the USA Frequency Standard as indicated in the table opposite. This correction is *not* with respect to the current value of frequency based on UT 2. A minus sign indicates that the broadcast frequency was low.

The WWV and WWVH time signals are synchronized; however, they may gradually depart from UT 2 (mean solar time corrected for polar variation and annual fluctuation in the rotation of the earth). Corrections are determined and published by the U. S. Naval Observatory.

* Received by the IRE, March 13, 1958.

WWV and WWVH time signals are maintained in close agreement with UT 2 by making step adjustments in time of precisely plus or minus 20 msec on Wednesdays at 1900 UT when necessary; one minus step adjustment was made at WWV and WWVH on December 11, 1957, January 15, 1958, and February 5 and 19, 1958.

WWV Frequency†

	Parts in 10^9		
Day	Dec., 1957 1500 UT	Jan., 1958 1500 UT	Feb., 1958 1500 UT
1	−4.8	−3.4	−2.6
2	−4.9	−3.3	−2.5
3	−4.9	−3.2	−2.5
4	−5.0	−3.1	−2.5
5	−5.0	−3.0	−2.5
6	−5.0	−2.9	−2.4
7	−4.9	−2.9	−2.3
8	−4.9	−2.8	−2.3
9	−4.9	−2.8	−2.2
10	−4.8	−2.8	−2.2
11	−4.7	−2.7	−2.2
12	−4.6	−2.6	−2.2‡
13	−4.4	−2.6	−2.1
14	−4.3	−2.6	−2.1
15	−4.2	−2.6	−2.1
16	−4.1	−2.7	−2.1
17	−4.1	−2.7	−2.1
18	−4.0	−2.7	−2.1
19	−4.0	−2.7	−2.2‡
20	−3.9	−2.7	−2.2
21	−3.8	−2.8	−2.3
22	−3.8	−2.8	−2.3
23	−3.8	−2.9	−2.3
24	−3.7	−2.9	−2.3
25	−3.7	−2.9	−2.4
26	−3.7	−2.9	−2.4‡
27	−3.7	−2.9	−2.5
28	−3.7	−2.9	−2.5
29	−3.6	−2.8	
30	−3.6	−2.8	
31	−3.6	−2.7	

† WWVH frequency is synchronized with that of WWV.
‡ Decrease in frequency of 0.5×10^{-9} at 1900 UT at WWV.

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A New Type of Low-Noise Electron Gun for Microwave Tubes*

During an investigation of noise reduction in backward-wave amplifiers which was conducted at the Hughes Research Laboratories, a series of experiments was undertaken to study the effects on beam noisiness of radical changes in the dc potential distribution in the immediate vicinity of the cathode. In particular, the diode region of an electron gun was designed to allow manipulation of electron emission and flow at exceedingly low potentials and thus to permit examination of the underlying assumptions and validity of existing one-dimensional noise theories.

These experiments, performed in 1956 and early 1957, resulted in a cascade backward-wave amplifier with a noise figure lower than 4 db at S band, as described at the Fifteenth Annual Electron Tube Research Conference.[1] Since that time the results have been reproduced on a number of tubes; tube noise figures in the vicinity of 3.5 db have been measured in several instances. Here the purpose is to formally announce these results and to briefly comment on the type of low-noise electron gun which has evolved from this work.

The basic features of the gun are 1) emission which originates predominantly from the edge and side of the cathode, and 2) a potential profile for these edge electrons which, in the cathode region, departs drastically from the usual (approximately Fry-Langmuir) potential distribution. We are led to the concept of *linear edge current* (as measured in terms of current per unit length of available cathode edge) as being a key design parameter for these low-noise guns, as distinguished from uniform current *density* which, on the basis of one-dimensional noise theories, is regarded as a controlling factor in the performance of conventional low-noise guns.

It is evident that the basic noise quantities in the beam (velocity and current fluctuations and their correlation) were being manipulated in the very low potential region of the gun near the cathode. Indeed, the unusual potential profile permits a possible drifting action at very low mean velocities or, more probably, a gradual acceleration of the beam through an extended region where the thermal velocity spread is large. It should be noted that Saito[2] has recently measured some correlation in typical beams and Siegman, Watkins, and Hsieh[3] have shown that such a correlation can develop on a one-dimensional beam as it is accelerated along the Fry-Langmuir potential profile from the potential minimum to a potential of several tenths of a volt. Thus one might expect to find an enhancement of this correlation and consequently a further reduction in the theoretically predicted minimum noise figure as a result of the unique potential profile characteristic of this gun.

Another possible explanation lies in a reduction of initial kinetic noise voltage by virtue of the edge emission mechanism, as proposed at the Tube Research Conference.[1] The particular combination of geometry and electric fields which produces the edge emission also gives rise to the distorted potential configuration, and vice versa. Therefore it has not yet been possible to experimentally determine which of these features is really the essential one.

Current research is being devoted to the frequency-scaling characteristics of the noise reduction mechanism. It is planned to describe in detail the operation of the gun and its relation to existing noise theories in a future paper.

Malcolm R. Currie
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* Received by the IRE, February 17, 1958.

[1] M. R. Currie and D. C. Forster, "New results on noise in electron beams," presented at the Fifteenth Annual Electron Tube Research Conference, Berkeley, Calif.; June, 1957.

[2] S. Saito, "Determination of the Correlation Ratio π/S of the Noise in an Electron Beam," M.I.T., Res. Lab. of Electronics, Cambridge, Mass., Quarterly Progress Rept., pp. 29, 30; October 15, 1957.

[3] A. E. Siegman, D. A. Watkins, and H. Hsieh, "Density-function calculations of noise propagation on an accelerated multivelocity electron beam," *J. Appl. Phys.*, vol. 28, pp. 1138–1148; October, 1957.

S-Band Traveling-Wave Tube with Noise Figure Below 4 DB*

In the preceding letter Currie reports noise figures below 4 db obtained at S band with a backward-wave amplifier. His tube uses an annular cathode with a resulting hollow electron beam. Experiments have been undertaken at Bell Telephone Laboratories to see whether comparable noise performance could be obtained in a conventional, low-noise traveling-wave tube employing the usual type of cathode, *i.e.*, one having a plane circular emitting surface. The electrode configuration and voltage profile, however, were similar to those used by Currie. The main observation to be reported is the achievement of noise figures as low as 3.5 db.

The test vehicle employed was an S-band amplifier in which reproducible noise figures of 4.8–5.0 db had been observed previously. Initial experiments were confined to changes in the gun configuration. They gave no overall improvement in noise figure but indicated best performance at lower currents. In order to provide more gain at the lower current, the helix pitch was decreased to give a γa of 1.7. This change, together with the more important changes in the cathode region, yielded tubes in which the 3.5-db noise figure was observed. The changes in the cathode region included operation of the beam forming electrode at a positive potential giving rise to enhanced electron emission from the cathode edge and thereby to a somewhat hollow beam.[1] The voltage configuration of the other electrodes was similar to that used by Currie in that it produced a low velocity drift region near the cathode.

Operating parameters of the original and modified versions of the tube are given in Table I.

TABLE I

Parameter	Original Version of S-Band Tube	Modified Version
I_k (microamps)	500	60
V_{helix} (volts)	560	200
γa	1.05	1.7
Cathode diameter (inches)	0.025	0.025
Helix TPI	100	160
Helix ID (inches)	0.051	0.051
Gain (db)	22	20
F_{opt} (db) at 600 gauss	4.8	4.3
F_{opt} (db) at 1600 gauss	4.8	3.5

It should be noted that the lowest value of noise figure was obtained with a magnetic field strength considerably higher than that required for focusing alone. This effect has also been observed by Currie. To date one tube with a 3.5-db noise figure and two tubes with 3.8-db noise figures have been made. In all cases the noise figure improved with increasing magnetic field strength, the observed slope up to 1600 gauss being approximately 0.1 db/100 gauss with little or no improvement beyond. The noise figure

* Received by the IRE, February 17, 1958.

[1] The hollow nature of the beam was confirmed experimentally in a demountable system by K. J. Harker using a pin-hole probe in conjunction with a 2:1 enlarged version of the gun.

has been measured over a 500-mc band and found to be relatively frequency-insensitive, being constant within 0.2 db with all voltages fixed.

The present study is being continued in an effort to obtain a better understanding of the noise reduction mechanism involved.

M. Caulton
G. E. St. John
Bell Telephone Labs., Inc.
Murray Hill, N. J.

Design Considerations for Circulator Maser Systems*

In its present form the low-noise maser amplifier[1] is a single-port device. That is, the low-level signal enters and the amplified signal leaves the maser by the same port. By connecting a circulator to the single-port maser, maximum gain-bandwidth is obtained, and the maser is isolated from the noise radiated by the receiver.[2] A circulator maser system is shown in Fig. 1. The signal entering through the antenna is directed by the circulator into the maser, and the amplified signal from the maser is directed by the circulator to the receiver.[3]

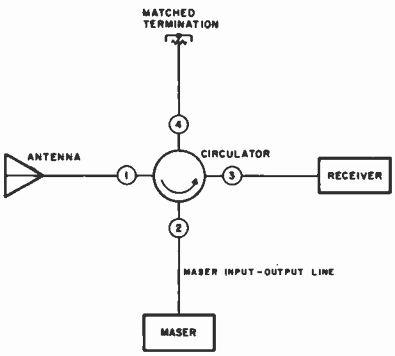

Fig. 1—Circulator maser system.

The effect of various characteristics on system noise temperature and maser stability will be considered. Examination of Fig. 1 indicates the following primary noise contributions:

T_M = noise temperature of maser proper.

T_R = noise temperature[4] of receiver. (Because the receiver is preceded by a maser amplifier, the noise contribution due to T_R becomes T_R/G_M, where G_M is the available maser gain expressed as a power ratio.)

T_{LA} = noise temperature at the antenna terminals due to noise emitted by the matched load which is reflected at the antenna.

T_{LM} = noise temperature at the maser input terminals due to noise emitted by the matched load which is transmitted directly through the circulator because of finite isolation between ports 4 and 2 (L_{42}). For the case of the matched load near room temperature, $T_{LM} = 290/L_{42}$.

T_{RM} = noise temperature at the maser input terminals due to noise emitted by the receiver[5] which is transmitted directly through the circulator because of finite isolation between ports 3 and 2 (L_{32}).

L = all dissipative losses between the antenna and the maser terminals; *i.e.*, the loss in the antenna feed line, the circulator forward loss between ports 1 and 2 (L_{12}), and the loss in the maser input-output transmission line (expressed as available signal power input divided by available signal power output. A loss of 0.1 db yields $L = 1.0233$).

The noise contribution due to L is calculated as follows.[6] The noise factor[7] F_1 of a passive matched network of loss L at a physical temperature T_1 can be shown to be

$$F_1 = 1 + (L - 1)\frac{T_1}{290}. \qquad (1)$$

The noise factor F_{12} of two networks in cascade is

$$F_{12} = F_1 + L(F_2 - 1). \qquad (2)$$

Combining (1) and (2) with the definition for effective noise temperature, $T = (F-1)290$, the effective noise temperature T_{12} of two cascaded networks is

$$T_{12} = (L - 1)T_1 + LT_2, \qquad (3)$$

where T_2 is the effective noise temperature of network 2.

The general expression for effective maser system noise temperature T_e can now be calculated. The path of the received signal is traced from the receiver back toward the antenna terminals. The noise temperatures are added at the points in the circuit at which they occur. Transformations from point to point through lossy components are calculated using (3).

A simplified form of the resultant expression[8] for effective maser system temperature is written below by assuming that the loss in the maser input-output transmission line is less than 0.4 db.

$$T_e = T_{LA} + (L - 1)T_0 + L\left[T_{RM} + T_{LM} + T_M + \frac{T_R}{G_M}\right], \qquad (4)$$

where all noise temperatures are in ° Kelvin, and T_0 is the physical temperature of the lossy components contributing to L. The fractional error due to the above simplification is under 10 per cent. T_R here is taken to represent the effective receiver noise temperature at the maser output terminals. That is, the losses between the maser output terminals and the receiver are included in T_R by the use of (3).

If L_{32} and L_{42} are approximately 25 db, and if the matched load is at room temperature (290°K), T_{LM} and T_{RM} are each approximately 1°K. If, in addition, the antenna swr is 1.1 or less, T_{LA} is less than 1°K. However, if the antenna swr is higher, the matched termination may have to be refrigerated in order to reduce its noise contribution.

It is, therefore, concluded that for the case where L_{32} and $L_{42} \approx 25$ db and an antenna swr ≈ 1.1, the contributions of T_{LA}, T_{LM}, and T_{RM} become small; letting T_0 equal 290°K, (4) thus becomes

$$T_e = (L - 1)290 + L\left[T_M + \frac{(F_R - 1)290}{G_M}\right]. \qquad (5)$$

The bracketed term in (5) is the effective noise temperature at the maser input terminals. For high-gain masers, this term will be considerably less than 290°K. Under this condition, and if $L < 1.2$ (0.75 db), (5) can be approximated by

$$T_e = 66.7l + [T_M + (F_R - 1)\ 290/G_M], \qquad (6)$$

where l is the loss L expressed in decibels. As can be seen, T_e increases by approximately 7°K per 0.1 db of loss for $l < 0.75$ db. For the case where the bracketed term is 6°K, T_e is doubled for $l \approx 0.1$ db.

It is thus seen that the noise due to the dissipative losses in the antenna feed line, the circulator, and the maser input-output line, can be significantly greater than the noise contributed by a high-gain maser.[9] The circulator dissipative loss is likely to be a significant fraction of the losses, particularly at frequencies below 3000 mc.[10]

In addition to the noise considerations above, the question of maser stability must be considered.

Single-port masers will experience variations in gain and may even oscillate, depending upon load conditions, due to the reflec-

* Received by the IRE, January 21, 1958. This work was supported by the Department of Defense.

[1] A. L. McWhorter and J. W. Meyer, "A solid-state maser amplifier," *Phys. Rev.*, vol. 109, pp. 312–318; January, 15, 1958.
H. E. D. Scovil, G. Feher, and H. Seidel, "Operation of a solid state maser," *Phys. Rev.*, vol. 105, pp. 762–763; January 1, 1957.

[2] A. E. Siegman, "Gain bandwidth and noise in maser amplifiers," Proc. IRE, vol. 45, pp. 1737–1738; December, 1957.

[3] In some applications it may be desirable to cascade two single-port maser amplifiers. This may be done by cascading two circulator maser systems, or more simply by modifying the circuit of Fig. 1 by replacing the termination at circulator port 4 with the receiver, and connecting a second maser at port 3. This arrangement would be useful, for example, when a requirement for large bandwidth has resulted in a lower maser gain, in which case the noise contribution due to the receiver becomes substantial.

[4] This receiver noise temperature (T_R) is the effective receiver input noise temperature defined as $(F_R - 1)290$ where F_R is the receiver noise factor expressed as a power ratio. See also, J. Greene, "Noise factor and noise temperature," Proc. IRE, vol. 46, p. 2A; January, 1958.

[5] The noise temperature corresponding to the noise power available *from* the input terminals of a receiver is, in general, not equal to the effective input noise temperature T_R. For example, the input impedance of a crystal mixer acts as a resistive noise generator at room temperature (about 290°K). For a particular RCA 6861 low-noise traveling-wave tube, it was measured to be 470°K at 2700 mc.

[6] It is assumed that the noise power available from the ferrite circulator is equal to that available from a matched resistive network having the same dissipative loss in the direction of propagation that is of interest.

[7] H. T. Friis, "Noise figures of radio receivers," Proc. IRE, vol. 32, pp. 419–422; July, 1944.

[8] Additional noise sources that must be considered, but that are not included in the present discussion, are the dissipative losses in the antenna itself, and the noise temperature seen by the antenna. Since the latter contribution can be substantial, and may be the limiting factor in system sensitivity, extreme refinement in the design of the circulator maser system may be unnecessary in some applications.

[9] A. L. McWhorter and F. R. Arams, "System-noise measurement of a solid-state maser," this issue, p. 913.

[10] C. L. Hogan, "The low-frequency problem in the design of microwave gyrators and associated elements," IRE Trans. on Antennas and Propagation, vol. 4, pp. 495–501; July, 1956.

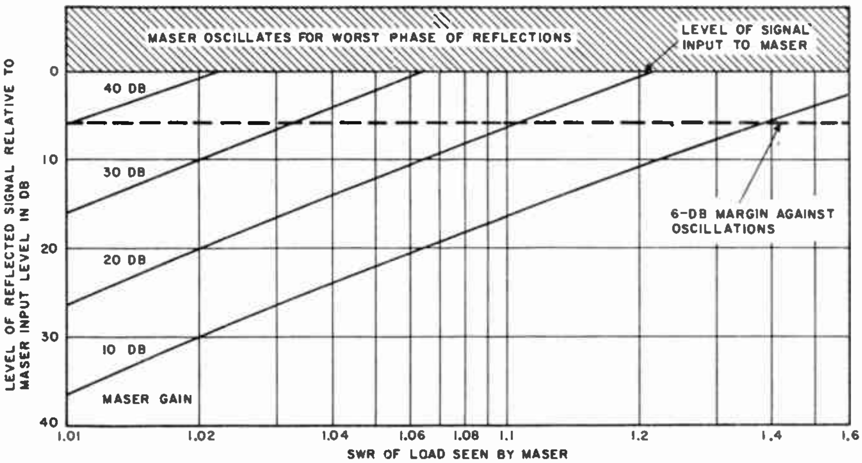

Fig. 2—Effect of load swr and maser gain on maser stability.

tion of some of the amplified signal back into the maser. Fig. 2 is a calculated plot of the level of the reflected amplified signal relative to the input signal as a function of load impedance, expressed in terms of swr. The parameter is maser gain in db. The shaded area represents the conditions of load swr and maser gain for which oscillations may occur. Thus, for a maser gain of 20 db, and for a reflection of arbitrary phase, a 6-db minimum margin to prevent oscillations is obtained by holding the swr to 1.1. The permissible value of swr can be increased by controlling the phase of the reflection or decreasing the oscillation margin.

Reflections from all components of the system will contribute to the swr at the maser terminals. Contributions due to antenna and receiver mismatch may be significant, even though isolated by L_{21} and L_{32}, respectively. For example, it is required that $L_{32}=25$ db in order to reduce a receiver swr of 2 to a value of 1.04 as seen by the maser.

From the above considerations, it can be concluded that the noise and stability requirements for a low-noise circulator maser system are as follows:

Noise

1) The circulator must have a low dissipative loss (a few tenths of a decibel) and an isolation of 25 db or more.
2) The antenna swr should be about 1.1, unless the matched termination of circulator port 4 is refrigerated.
3) The dissipative loss in the maser input-output line and in the antenna feedline must be low.
4) The maser gain should exceed 20 db to reduce the receiver noise contribution to a low value.

Stability

The swr seen by the maser must be low (unless the phases of the reflections are controlled). This in turn requires

1) A low circulator swr.
2) Reasonable antenna and receiver swr in combination with high circulator isolation.

The noise and stability considerations discussed above result in stringent requirements on circulator performance, which become increasingly difficult to realize at the lower microwave frequencies.[10] Fig. 3 shows

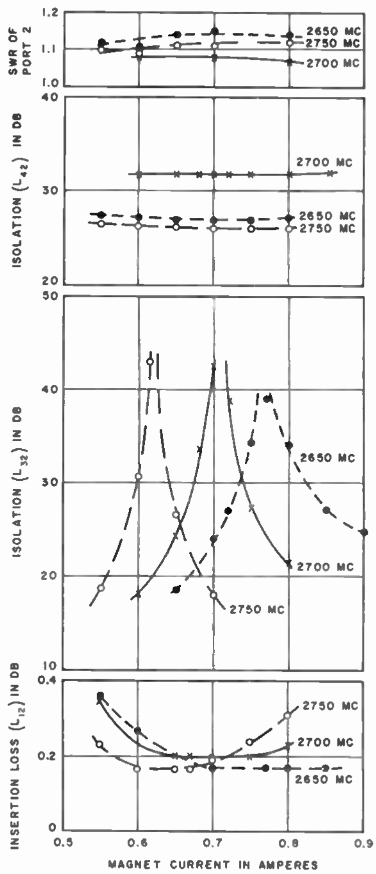

Fig. 3—*S*-band circulator performance.

the measured characteristics (near 2700 mc) of a tunable circulator that was developed for *S*-band maser application. The circulator insertion loss was less than 0.2 db near 2700 mc, and remained below 0.4 db from 2300 to 3150 mc (30 per cent bandwidth). The increase in loss is believed to be due to bandwidth limitations in the hybrid junctions rather than to losses in the ferrite.

Helpful discussions with J. Greene, B. Salzberg, and E. G. Fubini are gratefully acknowledged.

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System-Noise Measurement of a Solid-State Maser*

The effective input noise temperature[1] of a 2800-mc, three-level[2] solid-state maser of the reflection cavity type has previously been reported[3] to be less than 20°K, where this temperature refers to the maser cavity only, and not to the complete amplifier system. These first measurements were made with a room temperature isolator on the output side of the maser, which, because of the reciprocal nature of reflection-cavity masers, introduced about 300°K noise into the system. As a result, it was difficult to make a precise measurement of the relatively small amount of noise contributed by the maser itself. By combining a recently designed low-loss *S*-band circulator[4] with the maser, it has now been possible to form a complete amplifier system with an effective input noise temperature under 25°K.

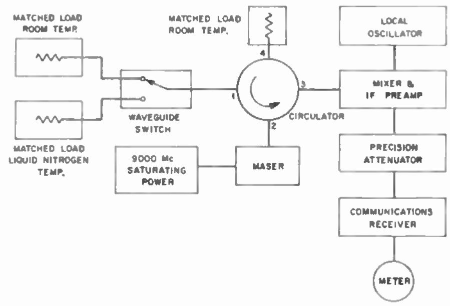

Fig. 1—Block diagram of instrumentation for noise measurement.

Fig. 1 shows the experimental arrangement used for the measurement. Port 1 of the circulator could be connected through the waveguide switch either to the matched load at room temperature or to the matched load at liquid nitrogen temperature, the loads having a vswr of 1.04 and 1.07, respectively, looking back through the switch. The maser, which was connected to port 2, has a voltage gain-bandwidth of about 1.8×10^6 $\sec^{-1.5}$ $K_3Co_{0.995}Cr_{0.005}(CN)_6$ is used as the paramagnetic salt and the operating temperature is 1.25°K. With 15 mw of 9000-mc saturating power, the external coupling at the amplifying frequency was adjusted to give a gain of the order of 30 db with 50-kc bandwidth. The receiver system connected to port 3 had an over-all noise figure of 8.5 db, the communications receiver being used to provide a bandwidth small in comparison

* Received by the IRE, January 20, 1958. The research reported in this document was supported jointly by the Army, Navy, and Air Force under contract with the Mass. Inst. Tech., Cambridge, Mass., and by the Dept. of Defense.

[1] The effective input noise temperature T_e is related to the noise figure F by the expression

$$T_e = 290(F-1).$$

[2] N. Bloembergen, "Proposal for a new type solid state maser," *Phys. Rev.*, vol. 104, pp. 324–327; October 15, 1956.

[3] A. L. McWhorter, J. W. Meyer, and P. D. Strum, "Noise temperature measurement on a solid-state maser," *Phys. Rev.*, vol. 108, pp. 1642–1644; December 15, 1957.

[4] F. R. Arams and G. Krayer, "Design considerations for circulator maser systems," this issue, p. 912.

[5] A. L. McWhorter and J. W. Meyer, "A solid state maser amplifier," *Phys. Rev.*, vol. 109, pp. 312–318; January 15, 1958. The microwave cavity design of this experimental maser is far from ideal. A substantial improvement in bandwidth may be expected.

with that of the maser. Port 4 was terminated by a matched load at room temperature with a vswr of 1.05.

By switching back and forth between the room temperature load and the liquid nitrogen temperature load at port 1, and measuring the ratio of the noise outputs with the precision attenuator, the noise temperature of the entire amplifier system was found to be $20 \pm 5°K$.

The sources of noise in the system other than the maser itself are the circulator, the maser input-output line, and the receiver. The circulator noise arises from:

1) Attenuation in transmission from port 1 to port 2,
2) Noise from the matched load at port 4 being reflected from port 1,
3) Noise from the matched load at port 4 being directly transmitted to port 2,
4) Noise from the receiver at port 3 being transmitted to port 2.

Since the isolation between ports 4 and 2 was greater than 35 db, and between ports 3 and 2 greater than 27 db, and because of the low vswr of the loads at port 1, the noise from sources 2), 3), and 4) combined is under 1°K. The insertion loss between ports 1 and 2 is 0.2 ± 0.1 in db, only a small part of which is reflective. Hence the noise arising from 1) is between 7 and 21°K.

The attenuation between port 2 of the circulator and the maser cavity is 0.4 ± 0.05 db, of which about 0.15 db is at room temperature and the rest occurs in the coaxial line leading down through the helium dewar to the cavity. This line, which is stainless steel for thermal insulation, was resilvered following the original maser noise measurements, but is still lossy. Depending on the assumptions made for the thermal distribution along the coaxial line,[3] the total noise generated between port 2 and the maser is between 7 and 23°K. (With special dewars to permit the use of *S*-band waveguide, almost all of this noise could be eliminated.) The third source of noise, the receiver at port 3, contributes less than 2°K because of the high gain of the maser.

From the above discussion it appears that the noise temperature of the maser proper must be very close to the theoretical value, which is about 2°K under the operating conditions used. In addition, the dissipative loss of the circulator is probably closer to the lower estimate of 0.1 db.

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Re-Invention by Young Engineers*

In the September, 1957 issue, the IRE Student Quarterly published some comments on "Creativity," addressed to our Student Members. The present remarks are related thereto, but addressed to our mature members who work with, or influence younger men. A point of great importance to the student or young graduate, is the judgment we pass on the re-creation or re-invention of something already old. If a young man performs this feat, *with no knowledge of the prior work*, it can be as genuine a creation as when it was first done, and it may well be an example of innate ability. All too often, our potential inventors are abashed, and actually set back, when they discover their ideas have been recorded in prior art. An inhibition arises which may even affect the next creative act. We older (and wiser?) engineers must do our part to counteract this; we must recognize the re-creation or re-invention as a sign of merit. The young student or engineer should be made to feel encouraged and confident. At the same time, he will resolve to learn more of his field, so that his future creations will be extensions well beyond present knowledge, rather than repetitions of the past.

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* Received by the IRE, February 6, 1958.

Leakage in Foil Solenoids*

Worcester *et al.*[1] state that "holes or slots can be machined through the wall of the (foil) solenoid . . . " from which one might assume, in the absence of information on subsequent treatment, that the formation of burrs between layers is unimportant. The importance of burrs leaking current radially can be estimated by using the formula which applies to the case in which an imperfect insulant is used between the foil. This can be shown to be

$$\frac{Z}{Z_0} = c\left(\frac{R-1}{R+1}\right)\cdot \log_e \left\{\frac{c(R-1)^2 + R^2}{c(R-1)^2 + 1}\right\} \quad (1)$$

where Z/Z_0 = the ratio of coil resistance with leakage to that without leakage, $R = r_2/r_1$, the ratio of the outer to inner radius of the coil, and c = a leakage parameter defined by

$$c = \frac{s(1-s)}{4\pi^2} \cdot \frac{\rho_2}{\rho_1} \cdot \frac{1}{N^2}. \quad (2)$$

In this, s = the space factor, N = the number of layers, and ρ_2, ρ_1 = the resistivities of the insulant and foil respectively. Eq. (1) is shown for practical radius ratios in the three significant decades of c in Fig. 1. It can be seen that leakage is serious if c falls below ten. For a typical coil with $N = 10^3$, $s = 0.75$, (2) then shows it is necessary to maintain

$$\rho_2/\rho_1 > 2.10^9. \quad (3)$$

With a typical polyester film insulant this ratio can be made 10^{18} or more, so insulant leakage is negligible.

Now let us replace the insulant resistivity in (2) with one giving a leakage equivalent to nr/r_i burrs per turn, where each burr is taken to be a cube whose side is equal to the insulant thickness.

(The r/r_i factor allows the number of burrs to increase with the circumference of the turns which is the case when the end faces of the coil are machined.)

Then ρ_2 becomes

$$\rho_1 \cdot \frac{2\pi r_i L}{t^2(1-s)^2} \cdot \frac{1}{n} \quad (4)$$

where L = the coil length and t = the foil thickness, and inserting into (2)

$$n = \frac{1}{2\pi c} \cdot \left(\frac{1}{1-s}\right)\left(\frac{r_i}{t}\right)\left(\frac{L}{t}\right) \cdot \frac{1}{N^2}. \quad (5)$$

If c is to be greater than ten, and taking typical values of

$$s = \tfrac{3}{4}$$

$$N = 10^3$$

$$L/t = 3000$$

$$r_i/t = 1000$$

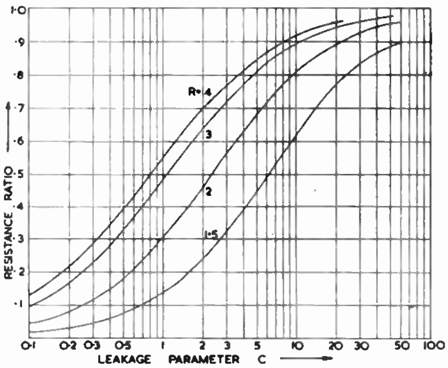

Fig. 1—Ratio of coil resistance with leakage to that with no leakage vs leakage parameter. *R* is the coil radius ratio.

we have

$$n < 0.2. \quad (6)$$

Thus, any more than one such "cubic burr" in every five turns is serious.

Our experience of coils using 0.001-inch aluminium foil and $s = \frac{3}{4}$ is that careful machining of the end faces does not give the very high freedom from burrs required, and that the equivalent of one cubic burr per turn is typical. If no holes or slots are required, winding with the insulant protruding is a solution, though this tends to produce a thicker barrier for heat flow from the coil than can otherwise be obtained. If holes and slots are required, then it seems necessary to remove at least 0.005 inch of metal from the exposed surfaces of a coil using 0.001-inch foil by chemical treatment.

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* Received by the IRE, December 24 1957.

[1] W. G. Worcester, A. L. Weitzmann, and R. J. Townley, "Lightweight aluminium foil solenoids for traveling-wave tubes," IRE Trans. on Electron Devices, vol. 3, pp. 70–74; January, 1956.

Lightning Enhancement of a VHF Tropospheric Scatter Signal*

The correspondence article by Bauer and Flood[1] was quite interesting in that it spoke of forward scatter from lightning strokes at 915 mc over a 400-mile path. This appears to supplement the findings of the writer at vhf.

In this instance, lightning enhancement of a tv video tropospheric scatter signal occurs each time a flash of lightning is seen visually. Channel 2 in Houston, 200 miles, and Midland, 300 miles from San Antonio, Texas, have been received in this manner. While listening to the video carrier, the BFO is turned on. When a lightning flash occurs nearby, the scatter signal increases greatly in intensity for the duration of the flash. Reflections off individual lightning bolts have been heard and observed visually when the storm is in the near vicinity. As the storm area moves off, the individual enhancements merge into a greater number but are weaker in intensity. Reflections lasting for many seconds have been observed, but the majority occur only for the split-second duration of the visual flash.

Back-scatter has been noted with the antenna pointed away from the station but in the direction of the storm area. Generally, the direction of the lightning from the antenna heading is not important because of the poor directivity pattern caused by using a 50-mc four-element Yagi at 55.25 mc.

The increase in "sferics" or noise spikes at vhf is considerably less than that at hf or lf so that the end result is a noticeable enhancement of the scatter signal. This has also been tried at the audio carrier frequency of 59.75 mc with the result that occasionally a complete word can be heard.

Of perhaps greater significance would be the detection of lightning occurring above the troposphere, possibly in the lower regions of the ionosphere, in this manner.

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* Received by the IRE, December 27, 1957.

[1] L. H. Bauer and W. A. Flood, "UHF forward scatter from lightning strokes," Proc. IRE, vol. 45, p. 1743; December, 1957.

Analysis of Sampled-Data Systems Containing Nonlinear Element*

Most of the literature in the field of sampled-data control has been concerned with linear systems. Nevertheless, many practical systems inevitably contain nonlinearity in the components to certain extent, and techniques for treating such systems are needed. In the following, a method for analyzing sampled-data feedback control systems containing nonlinear element is introduced.

* Received by the IRE, December 9, 1957.

A basic nonlinear sampled-data feedback control system is shown in Fig. 1. It is assumed that the nonlinear controlled system $G(s, e)$ may be described by a nonlinear element $N(e)$ connected in series with a linear element $G(s)$, as shown in Fig. 2. Due to the presence of an input $r(t)$, the actuating error would be $e(t)$ which is sampled by ideal sampler S to yield pulsed data $e^*(t)$. The discrete-data function $e^*(t)$ consists of a train of very narrow pulses which may be treated as equivalent impulses. The strength of each impulse is equal to the amplitude of the corresponding pulse of the signal. Thus, the discrete-data function can be expressed as

$$e^*(t) = a_0\delta(t) + a_1\delta(t-T) + a_2\delta(t-2T) + \cdots + a_k\delta(t-kT) + \cdots \quad (1)$$

where the coefficients $a_0, a_1, a_2, \cdots, a_k, \cdots$, represent the values of the actuating error at the respective sampling instants.

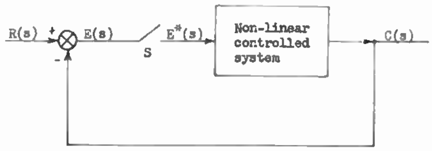

Fig. 1—A basic nonlinear sampled-data feedback control system.

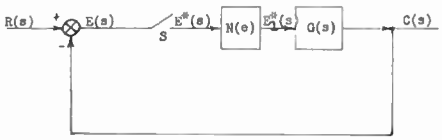

Fig. 2—Equivalent block diagram.

Since the nonlinear element $N(e)$ is time-invariant, the output signal $e_1^*(t)$ from $N(e)$ is also a discrete-data function the coefficients of which are related to the coefficients of $e^*(t)$ by the function, $e_1 = N(e)$, characterizing the nonlinear element. In symbols, $e_1^*(t)$ may be written as

$$e_1^*(t) = N(a_0)\delta(t) + N(a_1)\delta(t-T) + N(a_2)\delta(t-2T) + \cdots + N(a_k)\delta(t-kT) + \cdots. \quad (2)$$

Then, the z transforms of $e^*(t)$ and $e_1^*(t)$ are given by

$$E(z) = a_0 + a_1z^{-1} + a_2z^{-2} + \cdots + a_kz^{-k} + \cdots \quad (3)$$

and

$$E_1(z) = N(a_0) + N(a_1)z^{-1} + N(a_2)z^{-2} + \cdots + N(a_k)z^{-k} + \cdots \quad (4)$$

respectively. For well-behaved systems, these two series converge rapidly.

From Fig. 2, it is obtained that

$$C(s) = G(s)E_1^*(s) \quad (5)$$

$$C(z) = G(z)E_1(z) \quad (6)$$

$$C(z, m) = G(z, m)E_1(z) \quad (7)$$

and

$$E(z) = R(z) - C(z) = R(z) - G(z)E_1(z) \quad (8)$$

where the functions of z are the z transforms of the corresponding time functions; $C(z, m)$ and $G(z, m)$ are the modified z transforms associated with $C(s)$ and $G(s)$, respectively. In general, both $R(z)$ and $G(z)$ are ratios of two polynomials in z and can be expanded into power series in z^{-1}:

$$R(z) = r_0 + r_1z^{-1} + r_2z^{-2} + \cdots + r_kz^{-k} + \cdots \quad (9)$$

$$G(z) = g_0 + g_1z^{-1} + g_2z^{-2} + \cdots + g_kz^{-k} + \cdots \quad (10)$$

in which r_k's and g_k's depend upon the input signal $r(t)$ and the controlled system $G(s)$, respectively, and they are known constants for a given system subjected to a specified input.

Upon substituting (3), (4), (9), and (10) into (8), there results

$$\begin{aligned} a_0 + a_1z^{-1} + a_2z^{-2} + \cdots + a_kz^{-k} + \cdots \\ = r_0 + r_1z^{-1} + r_2z^{-2} + \cdots + r_kz^{-k} + \cdots \\ -(g_0 + g_1z^{-1} + g_2z^{-2} + \cdots + g_kz^{-k} + \cdots) \\ \cdot [N(a_0) + N(a_1)z^{-1} + N(a_2)z^{-2} + \cdots \\ + N(a_k)z^{-k} + \cdots]. \end{aligned} \quad (11)$$

Equating the coefficients of like terms, one obtains

$$a_0 = r_0 - g_0N(a_0) \quad (12)$$

$$a_1 = r_1 - [g_1N(a_0) + g_0N(a_1)] \quad (13)$$

$$a_2 = r_2 - [g_2N(a_0) + g_1N(a_1) + g_0N(a_2)] \quad (14)$$

and

$$a_k = r_k - \sum g_mN(a_n) \quad (15)$$

where $m+n=k$. From these equations, the coefficients a_k's which define the functions of $e^*(t)$, $e_1^*(t)$, $E(z)$, and $E_1(z)$ are readily evaluated either analytically or graphically. Consequently, by substituting the function of $E_1(z)$ into (7), the modified z transform of the system output is determined, from which the system response to input $r(t)$ can be found through inverse transformation. The method described above may be applied to the design of nonlinear compensators for sampled-data control systems.

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Doppler Equation for Earth Satellite Measurements*

In two Correspondence items,[1] techniques of radio tracking earth satellites were described.

It seems appropriate to offer further background and derivation data for the equation defining minimum slant range (Ro) of the satellite.

Incidentally, in Peterson's paper, a typographical error was apparently made inasmuch as the (velocity) "v" term in the

* Received by the IRE, December 26, 1957.

[1] R. R. Brown, P. E. Green, Jr., B. Howland, R. M. Lerner, R. Manasse, and G. Pettengill, "Radio observations of the Russian earth satellite," Proc. IRE, vol. 45, pp. 1552–1553; November, 1957.

A. M. Peterson, "Radio and radar tracking of the Russian earth satellite," Proc. IRE, vol. 45, pp. 1553–1555; November, 1957.

numerator of the equation for Ro, should be v^2.

For readers who are curious as to the derivation of the equation

$$Ro = \frac{fov^2}{c\dot{f}_{\max}}$$

per Brown, *et al.*,[1] we offer the following.

The basic Doppler equation applicable to the earth satellite problem is:

$$D_f = \frac{vfo}{c}\cos\gamma \quad \text{(See Fig. 1)}$$

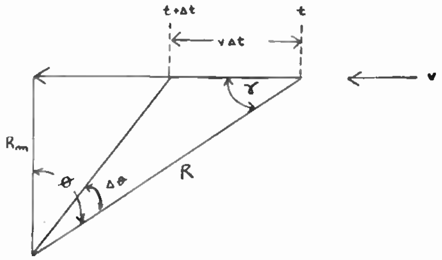

Fig. 1—Geometry of earth satellite problem.

where

D_f = frequency difference between transmitted and received carrier frequencies.

v = speed of satellite.

fo = transmitted carrier frequency.

γ = angle between satellite velocity vector and slant range vector.

R = slant range between receiving antenna and satellite.

c = velocity of propagation (186,000 miles per second).

R_m = minimum slant range.

$\theta = 90° - \gamma$.

$$D_f = \frac{vfo}{c}\cos\gamma = \frac{vfo}{c}\sin\theta \qquad (1)$$

$$\frac{d(D_f)}{dt} = \frac{vfo}{c}\cos\theta\frac{d\theta}{dt} \qquad (2)$$

but

$$\sin\Delta\theta \approx \frac{v\Delta t}{R} \approx \Delta\theta$$

for small angles.

$$\therefore \frac{\Delta\theta}{\Delta t} \approx \frac{v}{R} \approx \frac{d\theta}{dt} \qquad (3)$$

$$\frac{d(D_f)}{dt} = \frac{vfo}{c}\cos\theta\frac{v}{R} = \frac{fov^2}{cR}\cos\theta \qquad (4)$$

since

$$\frac{d(D_f)}{dt} = \text{maximum}$$

when $D_f = 0$ at which time

$$\theta = 0° \quad \text{and} \quad R = R_m$$

$$\frac{d(D_f)}{dt}\max = \frac{fov^2}{cRm} \qquad (5)$$

$$Rm = \frac{fov^2}{c\left(\frac{d(D_f)}{dt}\right)\max}. \qquad (6)$$

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A UHF Solid-State Maser*

The three-level solid-state maser as proposed by Bloembergen[1] has been operated successfully in the microwave range[2,3] verifying the predicted gain bandwidth[3] and noise properties.[4] In the usual configuration the paramagnetic crystal is placed in a cavity supporting both the saturating and amplifying frequency modes. This construction presents two difficulties. First, independent tuning of the two resonant frequencies is severely restricted and second, for high ratios of pumping to amplifying frequency, the crystal must be much smaller than the cavity, otherwise nodes of the saturating frequency magnetic field will occur in the sample, thus preventing complete saturation of the pumping levels. Both these difficulties may be overcome by using a foreshortened mode, such as a reentrant structure, at the lower frequency; however, coupling and tuning problems are still present. These considerations are especially critical at uhf where it is essential to operate at the highest possible pumping frequency and "filling factor" to overcome the loss in efficiency due essentially to the much lower quantum energy per transition.

To avoid these difficulties a maser has been constructed using a cavity mode at the pumping frequency and a lumped resonant circuit at the amplifying frequency. The paramagnetic salt is potassium cobalticyanide with 0.5 per cent chromicyanide impurity, as described in detail by McWhorter and Meyer.[3] At zero magnetic field this crystal has two doublets spaced at 5120 mc. Upon application of magnetic field, each doublet develops a separation of the order of a few hundred mc at fields of 50 to 100 oersteds. The individual separations are dependent upon the angle of the dc field with respect to the crystalline axes. By pumping between either of the two upper levels and the lowest level, it has been possible to obtain maser action between the bottom pair at 300 mc. The pumping frequency is 5300 mc, the magnetic field about 60 oersteds, and the saturating power, the order of 10 mw. The cavity is cylindrical and operates in the TE_{112} mode, with a wire loop lying in the plane of the rf magnetic field at the center of the cavity. The loop is resonated by placing dielectric between the loop leads which project out of the cavity through a small hole in the sidewall. The cavity mode is loop-coupled while the uhf circuit is coupled externally by means of a capacitative probe placed near the ungrounded lead from the loop. This arrangement seems to have little effect on the Q of the cavity mode and allows completely independent tuning of the two resonances. The losses in the lumped uhf circuit are conveniently overcome by plating the loop with lead, thus yielding superconductivity at the liquid helium temperatures used and a Q of the order of 10,000. By using a cylindrical crystal which completely fills the loop, the "filling factor" is greater than 50 per cent; that is, practically all the uhf magnetic field occurs in the sample. Since the sample is also much smaller than the wavelength of the cavity mode, it is saturated uniformly throughout its volume.

Preliminary measurements at 1.6°K give a typical gain of 10 db and bandwidth of 100 kc for the reflected power from the system. Correspondingly greater gains are obtained as the bandwidth is lowered. As an oscillator, the output power is of the order of 0.1 μw. Work is now in progress to develop nonreciprocal coupling arrangements so that the ultimate low noise figure of the order of 1°K may be obtained in an operational amplifier. This would be of special interest not only in communication and radar systems but also in radio astronomy research where discovery and possible study of the 327-mc deuterium line is of considerable interest.

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* Received by the IRE, December 30, 1957. The research reported in this document was supported jointly by the Army, Navy, and Air Force under contract with Mass. Inst. Tech., Cambridge, Mass.

[1] N. Bloembergen, "Proposal for a new type solid state maser," *Phys. Rev.*, vol. 104, pp. 324–327; October 15, 1956.

[2] H. E. D. Scovil, G. Feher, and H. Seidel, "Operation of a solid state maser," *Phys. Rev.*, vol. 105, pp. 762–763; January 15, 1957.

[3] A. L. McWhorter and J. W. Meyer, "A solid state maser amplifier," *Phys. Rev.*, (to be published).

[4] A. L. McWhorter, J. W. Meyer, and P. D. Strum, "Noise temperature measurement on a solid state maser," *Phys. Rev.*, (to be published).

A Property of Ladder Networks*

The object of this note is to generalize a property of driving point impedances so that it is applicable to the transfer functions of ladder networks. A driving point impedance $Z(s)$ is a positive real function and satisfies the condition[1] that

$$|\arg Z(s)| \le |\arg s| \text{ for } |\arg s| \le \frac{\pi}{2}. \qquad (1)$$

Now consider the following ladder network consisting of fixed, linear, and passive elements and having no mutual coupling between any of the series and shunt arms. The impedance $Z_i(s)$ is the driving point impedance of the entire ladder network to the right of the point at which the impedance is indicated in Fig. 1. Now the transfer admittance which relates the kth series current I_k to the input voltage E_0 may be written as the following product, where the integer k is odd.

$$\frac{I_k}{E_0} = \frac{I_1}{E_0}\cdot\frac{E_2}{I_1}\cdot\frac{I_3}{E_2}\cdot\frac{E_4}{I_3}\cdots\frac{E_{k-1}}{I_{k-2}}\cdot\frac{I_k}{E_{k-1}} \qquad (2)$$

$$= \frac{Z_2Z_4\cdots Z_{k-1}}{Z_1Z_3\cdots Z_k}. \qquad (3)$$

It should be noted that this method of expressing a transfer function may also be found in Guillemin's book on network synthesis.[2] But since each Z_i satisfies condition

* Received by the IRE, December 4, 1957.

[1] O. Brune, "Synthesis of a finite two-terminal network when the driving point impedance is a prescribed function of frequency," *J. Math. Phys.*, vol. 10, pp. 191–236; October 1931.

[2] E. A. Guillemin, "Synthesis of Passive Networks," John Wiley and Sons, Inc., New York, p. 531; 1957.

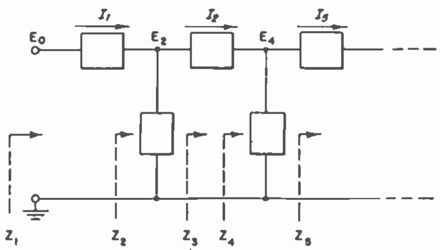

Fig. 1.

(1), the transfer admittance I_k/E_0 must satisfy the following generalization of (1).

$$\left|\arg \frac{I_k}{E_0}\right| \leq k\,|\arg s| \quad \text{for} \quad |\arg s| \leq \frac{\pi}{2}\cdot \tag{4}$$

In a similar fashion, restrictions on the arguments of E_p/E_0, I_k/I_1, and E_p/E_0 may be obtained.

$$\left|\arg \frac{E_p}{E_0}\right| \leq p\,|\arg s|,$$

$$\left|\arg \frac{I_k}{I_1}\right| \leq (k-1)\,|\arg s|,$$

and

$$\left|\arg \frac{E_p}{I_1}\right| \leq (p-1)\,|\arg s|$$

for

$$|\arg s| \leq \frac{\pi}{2}\cdot \tag{5}$$

In these expressions, p is an even integer.

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Translation of Foreign Articles*

Recent information on the importance attached by Soviet scientific organizations on the translation and dissemination of scientific and technical literature should cause an American engineer or scientist to wonder at the lack of such support in this country. It appears that several thousand people in the U.S.S.R. are engaged on a full-time basis translating and directly reprinting the scientific and technical literature of the world and then distributing it to extremely long lists of technical workers. This is done extremely rapidly, in some cases in a matter of days by "express" routing. Whole issues of magazines from cover to cover, including advertising, are reprinted in the original language since most Soviet scientific personnel read at least one other language than their own.

In this country nothing remotely equivalent exists. Several technical magazines including the PROCEEDINGS publish abstracts of such literature (many months later) but it is either very difficult or impossible to obtain the original article, and even when it is obtainable, it requires translation. If the Federal government is really interested in promoting science and technology, then why not have an equivalent service sponsored by the Department of Commerce or the State Department with the originals and translations being sold by the Government Printing Office at nominal cost? This certainly would be well worth the cost and more important it would provide access to foreign technical periodicals which are presently unavailable to the large majority of American engineers and scientists but readily available to their Russian counterparts.

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Upon receipt of the above, the Editors solicited the letter printed below to find out the nature and scope of translating activity in the U. S.[1]

About three years ago, the National Science Foundation embarked on a program of support for the cover-to-cover translation of selected important Russian scientific journals. The program has operated by making grants to qualified scientific organizations for the partial support of this work with the remainder of the income being derived from the sale of subscriptions. At present the Foundation is supporting the translation of fifteen journals by grants to the following organizations: Acta Metallurgica, American Geophysical Union, American Institute of Biological Sciences, American Institute of Physics, American Mathematical Society, and Massachusetts Institute of Technology (Radio Engineering and Electronics). In addition to the journals supported by the Foundation, the National Institutes of Health supports the translation of eight journals and three private translating firms are translating an additional fifteen journals in physics, chemistry, ceramics, atomic energy, electricity, and biochemistry. The subscription costs of these thirty-eight journals range from \$9.80 to \$200 per annum depending on the size and the amount of support. In addition to these the Foundation also supports the translation of monograph books and abstracts.

Realizing the importance of scientific accuracy in the translations, the Foundation has depended upon the various scientific institutions for advice and oversight in carrying through this program.

Plans have been developed over the past year for enlarging the translation program for basic science as well as for technology. The Foundation, the Office of Technical Services at the Department of Commerce, and the National Institutes of Health are coordinating their efforts in planning this expanded program. If funds for support become available, then the enlarged program will be started and suitable announcements will be made.

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Mixer Crystal Noise*

In Appendix II of the paper by Messenger and McCoy,[1] there is a derivation of crystal mixer noise temperature. This seems unsound, mainly because it is based on the assumption of thermal equilibrium, as exemplified by the following quotation.

"Obviously, if a network and its generator are in thermal equilibrium, they may be separated without affecting the noise temperature measured at the network output. Thus, the noise temperature of the crystal with no input, t_{av}, is the same as that of the crystal connected to an input termination whose temperature is also t_{av}."

The definition of noise factor is based on thermal noise at a standard temperature T_0, and hence it is natural that the expressions for noise performance of a diode mixer should be framed to use the parameter noise temperature ($\div T_0$). However, a diode is not in thermal equilibrium and, although it is well established that an equivalent network may be used to calculate the input-output power and impedance relations, there is no basis for assuming that the noise characteristics may be calculated by assigning equivalent noise temperature values to the elements of the equivalent network.

For thermal noise in a resistance, at temperature T, a unique relation exists between conductance and mean square noise current $\overline{i^2}$ so that the available power $\overline{i^2}/4g$ is given by $kTdf$. For most devices exhibiting conductance which is a function of an accelerating field, $\overline{i^2}/4g$ is not constant but, for given conditions, an ***equivalent*** noise temperature $\overline{i^2}/4kgdf$ may be used. In dealing with nonlinear devices, the impedance is defined to be consistent with standard rules, and therefore it appears to be most sound to start from the average value of $\overline{i^2}$ in calculating noise power.

A representation of the mixer "box" including noise, is shown in Fig. 1 for a crystal with negligible spreading resistance. The noise current generators are fixed for a particular crystal by the amplitude of the beating oscillator voltage, and if low-frequency excess noise is small at the intermediate frequency concerned, $\overline{i_r^2}$ and $\overline{i_0^2}$ will be equal.

The noise temperature ratio (ntr) is given by:

$$t = \frac{1}{L_s'} + \frac{1}{L_i'} + \frac{\overline{i_r^2}}{4kT_0}\left\{\frac{1}{g_sL_s'} + \frac{1}{g_iL_{in'}}\right\} + \frac{\overline{i_0^2}}{4kT_0g_0} \tag{1}$$

where $L_s'(L_i')$ = conversion loss in available powers from signal (image) to output.

With a nominally image-matched mixer the variation in output admittance as the source admittances are varied is complex; but if we confine our interest to the effect of pure conductance mismatches equal at signal and image, or to the effect of mismatches caused by the tolerances of the crystal alone, the variation in output admittance is purely conductive. If the mismatch at the input is

* Received by the IRE, November 19, 1957.

[1] Received by the IRE, December 9, 1957.

* Received by the IRE, December 2, 1957.

[1] G. C. Messenger and C. T. McCoy, "Theory and operation of crystal diodes as mixers," PROC. IRE, vol. 45, pp. 1269–1283; September, 1957.

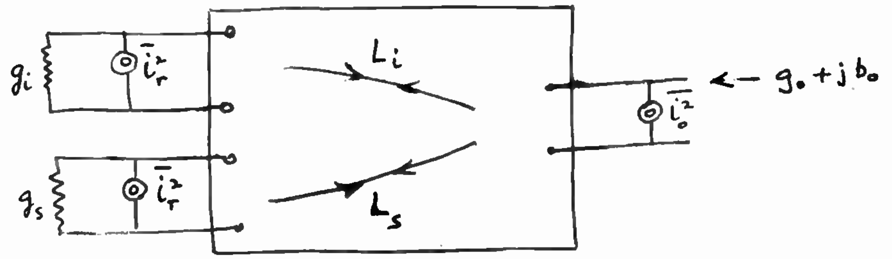

Fig. 1.

small, then $L_s' = L_i' \approx L$, the conversion loss for matched conditions.

Hence,

$$t \approx \frac{2}{L} + \frac{\overline{i_r^2}}{4kT_0} \frac{2}{gL} + \frac{\overline{i_0^2}}{4kT_0 g_0}. \qquad (2)$$

Now g_0 varies in sympathy with g and therefore the last two terms are at a maximum or minimum together, and the expression may be put in the form of directly measurable parameters, thus:

$$t = \frac{2}{L} + \left(t_m - \frac{2}{L}\right)\frac{g_m}{g_0} \qquad (3)$$

where t_m = ntr for matched conditions and g_m = output conductance for matched conditions.

Although this variation is important in practice, for mismatch alone may cause a large change in equivalent noise temperature, the point to be brought out here is that the derivation based on thermal noise concepts does not show any significant effects with mismatch and does not distinguish between mismatches of opposite phase. When the input terminals are short-circuited, the output conductance is theoretically the same as the average over the beating oscillator cycle,[2] and we then get $t = t_{av}$ as used by Messenger and McCoy. The practical value of t obtained in this condition is commonly less than unity, the value with matched source greater than unity, and the value with open circuit inputs may be large. Such results cannot be explained by thermal noise concepts but are consistent with the foregoing analysis.

It is possible to treat the image-reflected mixer in a similar way. For a strict analysis it is necessary to allow for the changed conversion loss, but it is sufficient to point out here that the short-circuit image mixer gives smallest ntr, and the open-circuit image mixer gives largest ntr. The spreading resistance cannot usually be neglected, and apart from contributing noise at all three pairs of terminals, it prohibits the practical achievement of very high or very low source impedances at the intrinsic mixer terminals. Furthermore, a strict analysis should include the effects of higher order conversion terms.

The fact that noise temperatures are lower for crystals of better conversion efficiency is surely for reasons more subtle than a simple swamping of the crystal noise. One reason mentioned by Nicoll[3] is that a good crystal has a high value of the exponent α in the equation $i = A\{\exp(\alpha V - iR_s) - 1\}$ and this is correlated with a reduction in low-frequency excess noise. A high value of this exponent confers the additional advantage that the mixer does not require such a large driving voltage so that the noise generated by reverse current is less. With silicon crystals, open circuit image gives minimum conversion loss but maximum noise temperature, thus again contradicting the thermal noise treatment of a mixer.[4]

In conclusion may I offer a plea for the abandoning of the term noise temperature as an abbreviation of noise temperature ratio. It seems unfortunate that the word which is commonly dropped is that which is essential with either of the other two.

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[2] L. C. Peterson and F. B. Llewellyn, "The performance and measurement of mixers in terms of linear-network theory," PROC. IRE, vol. 33, pp. 458–476; July, 1945.

[3] G. R. Nicoll, "Noise in silicon microwave diodes," *Proc. IEE*, pt. 3, vol. 101, pp. 317–324; September, 1954.

[4] Messenger and McCoy, *op. cit.*, see (30).

Diode Space and Space Charge*

The small-signal unidimensional theory for a diode space of unspecified potential profile leads to the following set of equations:

$$E = U^* \exp j\omega(t - T) \qquad (1)$$

$$\rho = \frac{1}{4\pi}\left[U^{**} - j\frac{\omega}{V_0}U^*\right]\exp j\omega(t - T) \qquad (2)$$

$$i = -j\frac{\omega}{4\pi}U^* \exp j\omega(t - T) \qquad (3)$$

$$v = \eta \frac{U}{V_0}\exp j\omega(t - T) \qquad (4)$$

in which the indices are for order of differentiation in the unidimension (z), E, ρ, i, v are the ac electric field, space charge density, current density, and electron velocity, respectively; t, the time, T, the dc transit time.

$$T = \int_0^z \frac{dz}{V_0}. \qquad (5)$$

V_0 is the dc velocity, ω, the radian frequency, η, the electron charge to mass ratio, and U, a potential defined by

$$V_0^3 U^{**} + 4\pi\eta i_0 U = 0 \qquad (6)$$

with i_0 the dc current density. The equations are written in Gaussian units. Eq. (6) can also take the form:

$$U^{II}V_0 - U^I V_0^I + 4\pi\eta i_0 U = 0 \qquad (7)$$

with the latin numerals for order of differentiation in the dc transit time T.

It is interesting to note that (7) can be solved easily for the three cases of major interest: drift space, space charge free, and space charge diode.

For the drift space V_0 is a constant and U is periodic radian plasma frequency:

$$\omega_e = \left(\frac{4\pi\eta i_0}{V_0}\right)^{1/2}. \qquad (8)$$

For the space charge free diode:

$$V_0 = \alpha T + V_0(0) \qquad (9)$$

with α constant and $V_0(0)$ the entrance dc velocity. It yields

$$U = (\alpha T + V_0(0))Z_2(u) \qquad (10)$$

with

$$u = \frac{2}{\alpha}\left[4\pi\eta i_0(\alpha T + V_0(0))\right]^{1/2} \qquad (11)$$

and Z_2 the general solution of the Bessel equation of order 2. For the space charge diode, because

$$V_0^{II} = 4\pi\eta i_0, \qquad (12)$$

U is solution of

$$U^{III} = 0. \qquad (13)$$

The space charge diode, of course, is the only rigorous solution. It is also the simplest one. For the two others one must determine validity conditions.

The correct solution being parabolic, the two approximations can only be valid within less than one of their respective periods. For the drift space the error is of the order of 10 per cent when $\omega_e T$ reaches one radiant or 16 per cent of one space charge wave period. For the space charge free diode one should keep in mind the condition

$$(\omega_e T)^2 \ll \left|\frac{V_0}{\alpha T}\right|$$

in which the plasma frequency and the dc velocity correspond to the value of the transit time, and the bars are for absolute value.

All this is well and good but space charge waves have been found along cylindrical beams in drift space. Thus, a self-consistent solution based on certain symmetry requirements (no variations through the cross section) may not lead to the best answer. The influence of the finite cross section and/or of added focusing devices (ions, transverse fields) seems to be to extend considerably the range of the validity of the approximations. Other self-consistent solutions like the Brillouin flow (radial variations, but no axial variations) lead also to periodic solutions. It is within the symmetry requirements that the difficulty lies of course. When is a beam large enough to be unidimensional? When has it small enough axial variations to yield

* Received by the IRE, January 3, 1958.

Brillouin or analogous flows? Are mixed solutions not only possible, but likely? For instance, is it not possible for a large beam to carry waves at the surface and to follow the unidimensional space charge solution on its axis? A more thorough analysis of dc conditions in beams, an analysis which would not from the start assume any symmetry requirement but possibly rotational symmetry, may yield the answer to this paradox.

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Error Rates in Data Transmission*

It is interesting to compare 1) the error rate reduction of a data transmission link which may be obtained by the use of error-correcting codes and binary signaling, and 2) the gain obtainable by choosing signaling alphabets other than binary. Roughly speaking, the decoding process in the first case amounts to digital correlation and in the latter case analog correlation, or integration.

Suppose sequences of k binary information digits are encoded into sequences of n digits where each sequence contains $(n-k)$ redundant digits.[1] Sidestepping the problem of the existence or precise structure of the "best" of such codes, one can compute fairly easily the error probability of the sequence, provided the digit error probability p is known and is independent for each digit, and provided also that the best code exists. 2^k vectors out of a total of 2^n vectors are chosen for transmission. If errors occur during transmission, a maximum of $(2^{n-k}-1)$ combinations of errors can be corrected. The best code will correct all single, double, etc., up to s-tuple errors plus a number j_{s+1} of $(s+1)$-tuple errors, where

$$\sum_{i=0}^{s} \binom{n}{i} + j_{s+1} = 2^{n-k}. \quad (1)$$

The total probability of receiving either the correct vector or any one of the correctable error vectors is

$$\sum_{i=0}^{s} \binom{n}{i} p^i q^{n-i} + j_{s+1} p^{s+1} q^{n-s-1}$$
$$\text{with } q = 1 - p. \quad (2)$$

Thus, the probability of receiving a sequence incorrectly is

$$P(n, k) = 1 - \Big[\sum_{i=0}^{s} \binom{n}{i} p^i q^{n-i} + j_{s+1} p^{s+1} q^{n-s-1} \Big] \quad (3)$$

with s and j_{s+1} determined by (1). For uncoded, straight binary transmission $n=k$ and the sequence error rate is

$$P(k) = 1 - (1-p)^k. \quad (4)$$

As an example, $P(n, k)$ and $P(k)$ are plotted in Fig. 1 as functions of p for $k=5$ (this is the case of teletype) and $n=16$. The sequence error rates are $P(16, 5) = 1-(q^{16} + 16pq^{15} + 120p^2q^{14} + 560p^3q^{13} + 1351p^4q^{12})$ and $P(5) = 1-q^5$. For small p, the asymptotic expressions are $P(16, 5) = 469p^4$ and $P(5) = 5p$.

The significance of these relations is as follows: suppose a 60-wpm teletype channel is available which prints an incorrect character in the average once every 20 lines. A slow-down by a factor 3.2 to about 18 wpm combined with the use of the best (16, 5) code would result in a wrong character only once every hundred thousand years. If the channel originally produced one wrong character every line, roughly once every minute, the same procedure would still result in an average period of about seven months between wrong letters. It is assumed here that the transmission rate of the binary symbols remains constant. This results in a slow-down of information transmission by a factor 3.2 because the (16, 5) code requires 16 binary symbols for the transmission of one character compared to five symbols for standard teletype.

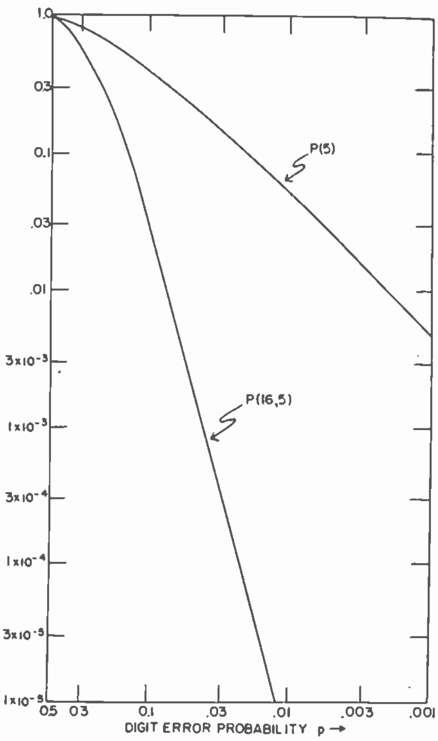

Fig. 1.

In most practical cases, a fixed rate of information transmission is desired, which in turn requires a speed-up of the binary signaling rate if an error-correcting code is to be used.

In general, this speed-up will increase the error probability p per binary symbol and no further assessment of the value of error-correcting codes is possible unless p is known as a function of the properties of the transmission channel.

The second way of encoding binary sequences, to be considered here, is the assignment of $M=2^k$ different signals, each corresponding to one of the 2^k binary sequences. The duration of each of these signals is k times the length of a binary pulse for fixed rate of data transmission. In many cases, a choice of symbols requiring amplitude discrimination at the receiver is not advantageous. Quantized frequency modulation, sometimes also called multiple frequency shift keying, may then be a good choice. Ordinary fsk is a special case with $M=2$. This type of modulation has been extensively investigated and applied to teletype by Jordan and others.[2] However, they did not use the maximum likelihood detector, whose operation is described below.

In quantized frequency modulation, the signal is a pulse of energy E transmitted on one out of M frequencies. Assume additive white channel noise of power density n_0 per cps. Assume further that coherent detection cannot be used, a condition frequently encountered in practice. Filters, matched to the pulse shape, are located at each of the M frequencies. The filter outputs are fed to envelope detectors and all M envelope samples are measured at the end of each pulse period. The signal corresponding to the largest sample is accepted. Suppose the pulse was transmitted on frequency f_i. The snr at the peak of the filter output would be $2\beta = 2E/n_0$.[3] Therefore, the probability distribution of the normalized envelope sample belonging to frequency f_i is[4]

$$p(y_i) = y_i \exp\left(-\frac{y_i^2 + 2\beta}{2}\right) I_0(y_i\sqrt{2\beta}). \quad (5)$$

The outputs of the other $(M-1)$ detectors are generated by noise alone and their distribution is

$$p(y) = y \exp\left(-\frac{y^2}{2}\right). \quad (6)$$

The receiver will select the correct signal if none of the noise samples exceeds y_i. The probability for this is

$$\left[\int_0^{y_i} y \exp\left(-\frac{y^2}{2}\right) dy\right]^{M-1}$$
$$= \sum_{r=0}^{M-1} (-1)^r \binom{M-1}{r} \exp\left(-\frac{r y_i^2}{2}\right). \quad (7)$$

The total probability for a correct decision is obtained by integration over y_i:

$$1 - p_M = \sum_{r=0}^{M-1} (-1)^r \binom{M-1}{r} \exp(-\beta) \int_0^\infty y_i \cdot \exp\left(-\frac{r+1}{2} y_i^2\right) I_0(y_i\sqrt{2\beta}) dy_i \quad (8)$$

and one obtains finally

* Received by the IRE, January 2, 1958.

[1] D. Slepian, "A class of binary signaling alphabets," *Bell Sys. Tech. J.*, vol. 35, pp. 203–234; January, 1956.

[2] D. B. Jordan, H. Greenberg, E. E. Eldredge, and W. Serniuk, "Multiple frequency shift teletype systems," PROC. IRE, vol. 43, pp. 1647–1665; November, 1955.

[3] S. Reiger, "Error probabilities of binary data transmission systems in the presence of random noise," 1953 IRE NATIONAL CONVENTION RECORD, pt. 8, pp. 72–79.

[4] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. J.*, vol. 24, pp. 46–156; January, 1945. See section 3.10.

$$p_M = \frac{\exp(-\beta)}{M}\sum_{r=2}^{M}(-1)^r\binom{M}{r}\exp\left(\frac{\beta}{r}\right). \quad (9)$$

In case $M=2$ (binary fsk) this reduces to $p_2=\frac{1}{2}\exp(-\beta/2)$. For large β, p_M approaches

$$p_M = \left(\frac{M-1}{2}\right)\exp(-\beta/2) = (M-1)p_2.$$

Assuming constant average transmitter power, fixed rate of data transmission, and quantized frequency modulation and detection as described above, let us now compare the error probabilities for 1) uncoded, straight binary transmission, 2) "best" (n, k) error-correcting code and binary signaling, and 3) signaling by means of $M=2^k$ different signals. The condition of fixed rate of data transmission implies $\beta_2(k)=n\beta_2(n, k)/k = \beta_M/k$ and, therefore,

$$p_2(n, k) = \tfrac{1}{2}(2p_2)^{k/n}. \quad (10)$$

Eq. (10) is the relation between the digit error probabilities in cases 1) and 2). The sequence error rates $P(k)$ and $P(n, k)$ are obtained from (3) and (4). Eq. (9) is the sequence error rate in case 3).

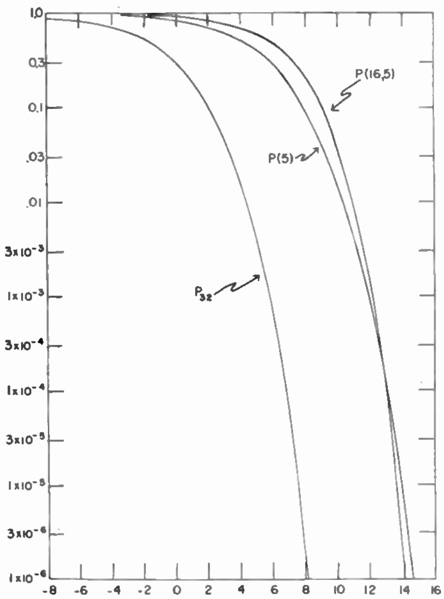

Fig. 2.

In Fig. 2, the sequence error rates are shown as functions of β_2 with the parameters of the previous example, $k=5$, $n=16$, $M=2^k=32$. The bandwidth required in the error-correcting case is 3.2 times greater than for uncoded binary transmission. The same increased bandwidth is needed for the 32-frequency transmission because 32 frequencies require 16 times the bandwidth of 2 frequencies, but the lengthening of the symbols by a factor 5 permits 5 times closer frequency spacing. It is seen that the error-correcting code gives a lower error rate than no code, provided β_2 is larger than 13 db, but much larger gains may be obtained by lengthening the symbols and selecting a large number of them.

Finally, if coherent detection is possible, a somewhat different choice of symbols is appropriate but the results are similar, although not quite as simple numerically.

The writer is indebted to R. Urbano of the Air Force Cambridge Research Center for the numerical calculations of (9).

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Common Emitter Transistor Amplifiers*

Design engineers working with transistors usually think of the common collector (emitter follower) configuration when a high input impedance circuit and power gain is desired. Such might be the case, for instance, if one were asked to design a transistor audio amplifier input stage driven by a crystal or ceramic pickup. It is surprising to find that the same input impedance and power gain can be obtained with the ordinary common emitter circuit with its input impedance increased by the addition of a series resistor. To see that this is really so, consider the following two circuits shown in Fig. 1.

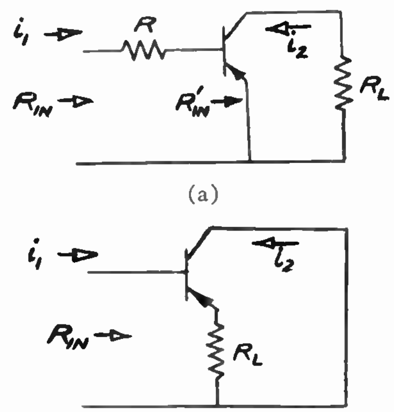

Fig. 1—(a) Common emitter amplifier. (Input resistance increased by addition of R.) (b) Emitter follower amplifier

Assumptions:

1) Each circuit has the same operating point and battery supply.
2) Each circuit is driven by a constant sinusoidal current of i_1 amperes (small signal).
3) $R_L \ll$ collector junction resistance, so that current gain is approximately the short circuit current gain α_{fe}.
4) $R_{in}' \ll R_{in}$.
5) Resistance R is adjusted until the input impedances of each circuit are equal.
6) Voltage drop across emitter base junction is neglected.

* Received by the IRE, January 13, 1958.

Under these assumptions the power gains of each circuit are as follows.

For Common Emitter:

$$R_{in} = R + R_{in}'$$
$$P_{in} = i_1^2 R_{in} = i_1^2(R + R_{in}')$$
$$i_2 = \alpha_{fe} i_1$$
$$P_{out} = i_2^2 R_L = \alpha_{fe}^2 i_1^2 R_L$$
$$\therefore\ G_1 = \frac{P_{out}}{P_{in}} = \frac{\alpha_{fe}^2 i_1^2 R_L}{i_1^2(R + R_{in}')} = \frac{\alpha_{fe}^2 R_L}{R + R_{in}'}.$$

For Emitter Follower:

$$i_2 \cong \alpha_{fe} i_1$$
$$i_{load} = i_1 + i_2 = i_1 + \alpha_{fe} i_1 = i_1(1 + \alpha_{fe})$$
$$\cong \alpha_{fe} i_1 \quad \text{for} \quad \alpha_{fe} >> 1$$
$$R_{in} = \frac{i_2 R_L}{i_1} = \frac{\alpha_{fe} i_1 R_L}{i_1} = \alpha_{fe} R_L$$
$$P_{in} = i_1^2 R_{in} = i_1^2 \alpha_{fe} R_L$$
$$P_{out} = i_{load}^2 R_L = \alpha_{fe}^2 i_1^2 R_L$$
$$\therefore\ G_2 = \frac{P_{out}}{P_{in}} = \frac{\alpha_{fe}^2 i_1^2 R_L}{i_1^2 \alpha_{fe} R_L} = \alpha_{fe}.$$

Now if R is varied so that $R+R_{in}'=\alpha_{fe}R_L$, then

$$G_1 = \frac{\alpha_{fe}^2 R_L}{\alpha_{fe} R_L} = \alpha_{fe} = G_2$$

and the input impedances and power gains of each circuit are equal.

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Electromagnetic Analogs for the Gravitational Fields in the Vicinity of a Satellite*

With the advent of Sputniks, there has been a reawakened interest in space flight and its attendant problems. One of the more important of these is the phenomenon of weightlessness or the freedom from gravitational fields. Strictly speaking, the gravitational field in the vicinity of a satellite is not exactly zero but is balanced against centrifugal force only at one specific radius (for a given velocity of travel). The remaining field, together with the rotation, gives rise to some peculiar effects that would not be expected from our everyday experience. Understanding these effects gives a clearer insight into the relationship between electromagnetic and gravitational forces.

Consider, for example, what happens when a satellite is separated from the rocket stage that gave it its final velocity. Assuming the satellite is pushed forward by a spring or similar mechanism, the satellite will have a slightly higher specific energy than the rocket stage. As a result, the major axis of

* Received by the IRE, January 24, 1958.

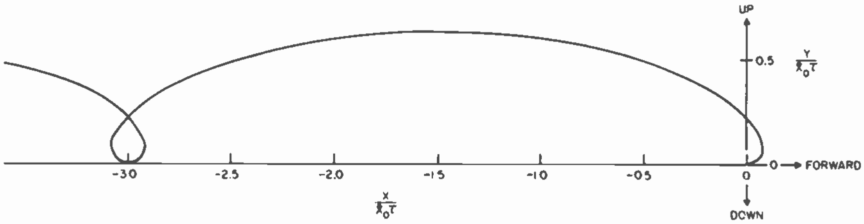

Fig. 1—Path of satellite relative to rocket stage.

the satellite orbit will be slightly greater than that of the rocket and, if we refer to Kepler's laws, we find that the period is slightly longer. The net result is that, though the satellite had a velocity greater than that of the rocket at the moment of separation, it travels through most of its orbit with a velocity slower than that of the rocket and, when one revolution has been completed, finds itself a considerable distance to the rear. To an observer located on the rocket with his feet always directed toward the earth, the path of the satellite would appear as shown in Fig. 1. The maximum rearward relative velocity is seven times that of the initial forward relative velocity.

Examining this problem more closely, if we consider a satellite traveling on a circular orbit with a reference frame centered on it and rotating with it so that the negative Y axis always points to the center of the earth and the positive X axis points in the direction of the satellite's motion, then, as shown in the Appendix, we find that the equations of motion for objects in the vicinity of the satellite may be written (neglecting second-order effects)

$$\begin{aligned} \ddot{X} &= -2\omega\dot{Y} \\ \ddot{Y} &= 2\omega\dot{X} + 3\omega^2 Y \\ Z &= -\omega^2 Z \end{aligned} \qquad (1)$$

where $\omega = 2\pi/\tau$ = the angular velocity of the radius vector connecting the satellite with the center of the earth.

It can be seen that the apparent acceleration can be divided into two components. One of these is a function of position only and the other is a function of velocity only. We may think of these accelerations as being due to the presence of a field composed of "gravostatic" and "gravomagnetic" or Coriolis components. The motion is the same as that which would be experienced by an electron moving in a combination of electrostatic and magnetic fields. The magnetic field is uniform and parallel to the Z axis, and a cross section showing the shape of the electrostatic equipotential lines in the YZ plane is shown in Fig. 2.

We do not normally think of gravitational forces as depending on velocity and, when a similar problem arises in connection with high-speed flight across the surface of the earth, we do not ascribe it to gravitation but instead call it the Coriolis acceleration. According to Einstein, however, acceleration and gravitation are fully equivalent. In the case of the electron, the magnetic-field component would normally be produced by other electrons rotating in the wires of a solenoid coil. Is it not then correct to consider our Coriolis field as being produced by the bal-

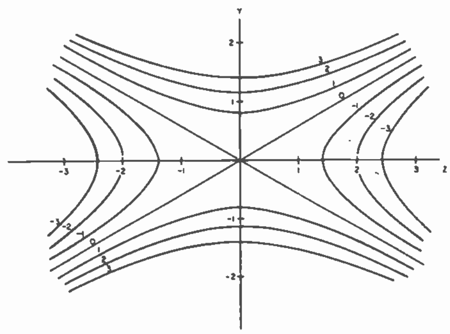

Fig. 2—Shape of gravostatic potential surfaces.

ance of the universe rotating about our satellite? Further reflection indicates that a moving mass must in fact induce something like a Coriolis field with the resultant generation of gravity waves very much like the electromagnetic waves induced by a moving charge. If this were not so, the change in gravitational field due to the mass in motion would be propagated instantly, which is contrary to the principles of relativity.

Newton's law of universal gravitation may be considered to be a first approximation to the relativistic theory of gravitation. The approximation is valid for nearly all the conditions we meet in our normal experience, and it is only when we consider phenomena such as the motion of the perihelion of Mercury that a discrepancy is noted (43 seconds per century). A closer approximation results, however, if we consider the laws of gravitation and Coriolis to parallel the laws of electricity and magnetism. We note, for example, that the law of universal gravitation has the same form as Coulomb's law of electric force except for the sign of the force. (Masses always attract but like charges repel.) It is as though charge were a manifestation of an imaginary mass. In terms of the forces produced, a kilogram of mass is equivalent to 0.86×10^{-10} coulombs of charge. If we now carry this concept further, a pipe carrying a fluid flow of mass generates a Coriolis field that encircles it just as a wire carrying current generates a magnetic field around it. Also, a gyro rotor has a Coriolis dipole moment just as a current carrying loop of wire has a magnetic dipole moment. The Coriolis dipole moment is, of course, just another name for angular momentum; but if we compute the torque developed by a gyro in a Coriolis field, we use the same equation as that for the torque of a current carrying loop in a magnetic field.

It would appear that the only reason we do not normally think in these terms is that the induced Coriolis fields are much too small to be detected in any sensible laboratory experiment. For example, the Coriolis field induced by a gyro rotor, having an angular momentum of 10^7 gm cm^2/second at a distance of 10 cm, is equivalent to a physical rotation of about $10^{-19°}$ per hour. It is possible that these effects may become significant when considered on the grand scale of celestial mechanics as in the study of the relative motions of galaxies. In any event, it may be that the concepts here expressed may have some pedagogical value in explaining the behavior of gyros and similar devices to those familiar with electromagnetic fields.

Appendix

Derivation of Equations of Motion

We consider a system of Cartesian coordinates with the center at the center of the earth and stationary with respect to inertial space. We neglect the effects of the sun, moon, and other nearby objects. In this system, the only force acting on a free body is the gravitational attraction, which varies inversely with the square of the distance from the center of the earth. At the surface of the earth this attraction produces an acceleration of g_0(32.2 feet/second^2), so that at any other radius R the total acceleration will be

$$A = g_0 \frac{R_0^2}{R^2} \qquad (2)$$

where R_0 is the radius at the surface of the earth. Since this acceleration is always directed toward the earth's center, we may resolve the acceleration into its components, writing

$$\left.\begin{aligned} \ddot{X}_1 &= -\frac{X_1}{R} g_0 \frac{R_0^2}{R^2} = -gR_0^2 \frac{X_1}{R^3} \\ \ddot{Y}_1 &= -\frac{Y_1}{R} g_0 \frac{R_0^2}{R^2} = -gR_0^2 \frac{Y_1}{R^3} \\ \ddot{Z}_1 &= -\frac{Z_1}{R} g_0 \frac{R_0^2}{R^2} = -gR_0^2 \frac{Z_1}{R^3} \end{aligned}\right\} \qquad (3)$$

where X_1, Y_1, and Z_1 are the coordinates in our inertial reference system and

$$R = \sqrt{X_1^2 + Y_1^2 + Z_1^2}.$$

These are the equations of motion in the inertial coordinate system which, when solved, will yield Kepler's laws. We now consider a second system in which the Z axis coincides with the Z axis of the first system but which is rotating about its Z axis with an angular velocity ω. The transformation equations may be written

$$\left.\begin{aligned} X_1 &= X_2 \cos \omega t + Y_2 \sin \omega t \\ Y_1 &= -X_2 \sin \omega t + Y_2 \cos \omega t \\ Z_1 &= Z_2 \end{aligned}\right\}. \qquad (4)$$

We note that

$$X_1^2 + Y_1^2 + Z_1^2 = R^2 = X_2^2 + Y_2^2 + Z_2^2.$$

If we now substitute these values in the right-hand side of (3) and their second derivatives in the left-hand side of (3), we obtain a set of equations from which t may be eliminated, yielding the resulting acceleration components in the rotating system. The solution we obtain is

$$\left.\begin{aligned}\ddot{X}_2 &= -2\omega\dot{Y}_2 + X_2\left[\omega^2 - \frac{g_0R_0^2}{R^3}\right]\\ \ddot{Y}_2 &= 2\omega\dot{X}_2 + Y_2\left[\omega^2 - \frac{g_0R_0^2}{R^3}\right]\\ \ddot{Z}_2 &= -Z_2\frac{g_0R_0^2}{R^3}\end{aligned}\right\}. \quad (5)$$

We note that one solution for these equations is

$$\left.\begin{aligned}X_2 &= 0\\ Y_2 &= R_c\\ Z_2 &= 0\\ \dot{X}_2 &= 0\\ \dot{Y}_2 &= 0\\ \dot{Z}_2 &= 0\end{aligned}\right\} \quad (6)$$

where

$$R_c = \left(\frac{g_0R_0^2}{\omega^2}\right)^{1/3}$$

or, in other words, a satellite on a circular orbit with angular velocity ω remains stationary in our rotating coordinate system. We now limit ourselves to the region in the immediate vicinity of the satellite. We adopt a third system of coordinates centered on the satellite with the transformation equations

$$\left.\begin{aligned}X_2 &= X_3\\ Y_2 &= Y_3 + R_c\\ Z_2 &= Z_3\end{aligned}\right\}. \quad (7)$$

Assuming that all the coordinates in this new system are very small relative to R_c, we may write the approximations

$$R \approx R_c + Y_3 \quad (8)$$

$$\frac{g_0R_0^2}{R^3} \approx \omega^2\left(1 - 3\frac{Y_3}{R_c}\right). \quad (9)$$

Making the appropriate substitutions and neglecting the second-order quantities, we obtain

$$\left.\begin{aligned}\ddot{X}_3 &= -2\omega\dot{Y}_3\\ \ddot{Y}_3 &= 2\omega\dot{X}_3 + 3\omega^2Y_3\\ \ddot{Z}_3 &= -\omega^2Z_3\end{aligned}\right\} \quad (10)$$

When we drop the subscripts, these are (1), previously stated.

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Off-Path Propagation at VHF*

The ionospheric forward scatter mode of vhf radio propagation has been successfully employed in communications circuits for several years. Early theoretical investigation[1] of this mode of propagation considered only the mechanism of isotropic scattering in a turbulent solar-controlled ionosphere.

* Received by the IRE, January 30, 1958.

[1] F. Villars and V. F. Weisskopf, "The scattering of electromagnetic waves by turbulent atmospheric fluctuation," *Phys. Rev.*, vol. 94, pp. 232–240; April 15, 1954.

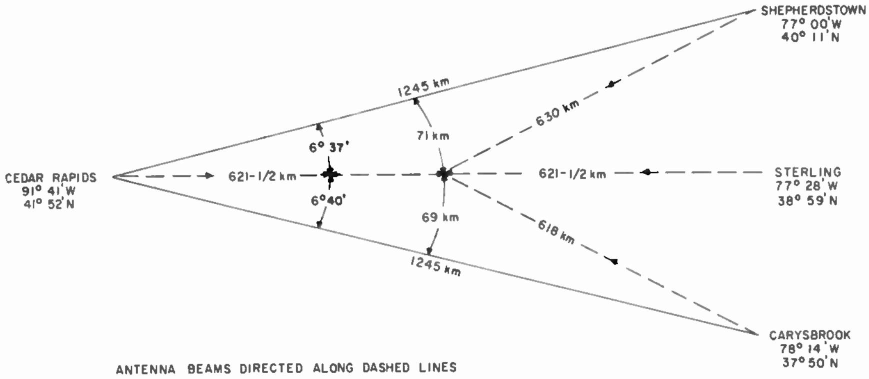

Fig. 1—Geometry for the off-path experiment.

Experimental results reported by Bailey, Bateman, *et al.*[2] were in generally good agreement with the turbulent-scatter theory during the daytime hours, but were evidently influenced by some other mechanism during the nighttime and early morning hours. This other mechanism was tentatively attributed to meteoric ionization.

Eshleman and Manning[3] have developed a theoretical model for radio propagation by scattering from meteor trails which shows that the areas of maximum density for useful meteor trails lie to one side or the other of the great-circle path between the transmitter and the receiver.

In January, 1956 the Central Radio Propagation Laboratory conducted an experiment which was designed to elucidate the relative contributions to ionospheric scatter propagation from these two mechanisms. Transmissions from Cedar Rapids, Iowa, at 49.8 mc beamed towards Sterling, Va. were recorded simultaneously at Sterling and at two off-path receiving sites; one north of Sterling at Shepherdstown, Pa., and the other south of Sterling at Carysbrook, Va. Identical rhombic antennas beamed towards the midpoint of the Cedar Rapids to Sterling path were used at each receiving site. Signal intensities at Carysbrook were greater than at either Sterling or Shepherdstown from 0400 hours to 1115 hours. Shepherdstown signal intensities were greater than at either Sterling or Carysbrook from 1715 hours to 0400 hours. Only at the midday hours, from 1115 to 1715, the signal received on the direct great-circle path to Sterling was greater than that received at either of the two off-path receivers. Signal intensities at all three sites reached maximum values during the midday hours.

These results, therefore, appear to indicate that there are two principal modes involved in ionospheric scatter propagation, one due to solar influence, the other due to meteoric ionization. It is worth emphasizing that the results are representative of wintertime conditions and an east-west path; results were reported earlier[4] from a summertime experiment, using only one "off-path" recording station (Carysbrook), with the same geometry as in the present experiment. The results obtained in July showed the off-path transmission to be dominant for fewer hours of the day.

The geometry of the experiment is shown in Fig. 1. The results of the simultaneous three-station signal intensity observations are presented graphically in Fig. 2.

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[2] D. K. Bailey, R. Bateman, *et al.*, "A new kind of radio propagation at very high frequencies observable over long distances," *Phys. Rev.*, vol. 86, pp. 141–145; April 15, 1952.

[3] V. R. Eshleman and L. A. Manning, "Radio communication by scattering from meteoric ionization," PROC. IRE, vol. 42, pp. 530–536; March, 1954.

[4] D. K. Bailey, R. Bateman, and R. C. Kirby, "Radio transmission at vhf by scattering and other processes in the lower ionosphere," PROC. IRE, vol. 43, pp. 1181–1230; October, 1955.

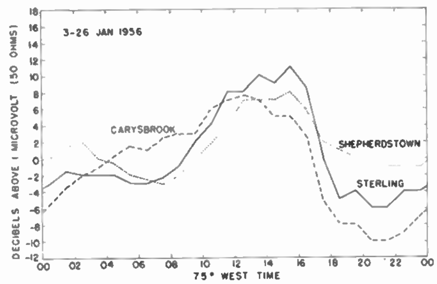

Fig. 2—Diurnal variation of hourly median signal intensities observed simultaneously at Shepherdstown, Sterling, and Carysbrook at 49.8 mc.

Experimental Check of Formulas for Capacitance of Shielded Balanced-Pair Transmission Line*

Gent[1] has developed a new formula for the capacitance of a shielded balanced-pair transmission line, and has compared it with three previously published formulas. He concludes that his formula is valid over a wider range of variables than the others. Experimental evidence in a few particular cases shows that the errors in the other formulas are not as great as estimated, and that the superiority of Gent's formula may be illusory.

* Received by the IRE, December 26, 1957.

[1] A. W. Gent, "Capacitance of shielded balanced-pair transmission line," *Elec. Commun.*, vol. 33. pp. 234–240; September, 1956.

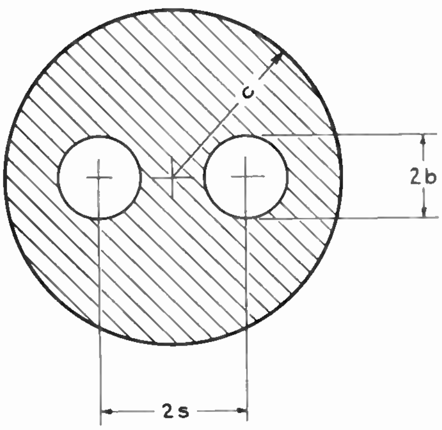

Fig. 1—Configuration of two wires symmetrically placed in a circular conducting shield.

TABLE I

$c = 5$ INCHES, $b = 0.9255$ INCH, $K = 8.8553$ $\mu\mu$F/METER

			C in $\mu\mu$f				
s	b/s	s/c	Measured (King)	(7)	(8)	(12)	(13)
1.500 inches	0.6170	0.3000	29.44	30.19*	29.95*	(29.31)	(27.94)*
1.875 inches	0.4936	0.3750	25.30	25.62	25.57	(25.21)	(24.93)
2.300 inches	0.4024	0.4600	23.80†	23.82	23.83	23.71	23.70
2.875 inches	0.3219	0.5750	24.60	(24.15)*	(24.23)	24.11*	24.41
3.500 inches	0.2644	0.7000	30.60†	(28.42)*	(28.73)*	(28.30)*	(29.24)*

The configuration under consideration is shown in Fig. 1. Two circular wires each of radius b are separated by a distance $2s$ between their centers. The wires are symmetrically placed within a conducting cylinder of inner radius c, and the enclosed space is filled with a dielectric of permittivity K_2.

The formulas compared by Gent are as follows:

(Gent)[2]

$$\frac{\pi K_2}{C} = \cosh^{-1}\frac{s}{b} - 2\tanh^{-1}\frac{s(s^2-b^2)^{1/2}}{c^2}$$
$$= \ln\left\{\frac{s+(s^2-b^2)^{1/2}}{b}\,\frac{c^2-s(s^2-b^2)^{1/2}}{c^2+s(s^2-b^2)^{1/2}}\right\}. \quad (7)$$

(Green, Curtis, and Mead)[3]

$$\frac{\pi K_2}{C} = \ln\left\{\frac{2s}{b}\,\frac{1-s^2/c^2}{1+s^2/c^2}\right\} - \frac{1+4(s/b)^2}{16(s/b)^4}(1-4s^2/c^2). \quad (8)$$

[2] The formula numbers are those used by Gent.
[3] E. I. Green, H. E. Curtis, and S. P. Mead, U. S. Patent No. 2,034,032; March 17, 1936.

(Craggs and Tranter, second-order approximation)[4]

$$\frac{\pi K_2}{C} = \ln\frac{2s}{b} - \ln\frac{c^2+s^2}{c^2-s^2} - \frac{\frac{1}{4}\left(1-\frac{4c^2s^2}{c^4-s^4}\right)^2\left(\frac{b}{s}\right)^2}{1-\left\{\frac{1}{4}+2\left(\frac{s}{c}\right)^2\frac{c^4+s^4}{c^4-s^4}\right\}\left(\frac{b}{s}\right)^2}. \quad (12)$$

(Meinke, or Craggs and Tranter first-order approximation)[4,5]

$$\frac{\pi K_2}{C} = \ln\frac{2s}{b} - \ln\frac{c^2+s^2}{c^2-s^2}. \quad (13)$$

King has made measurements of C using an electrolytic tank.[6] Although these were purely incidental to the evaluation of the tank as an experimental tool, they fall within the range considered by Gent. In Table I, these experimental measurements are compared to the four formulas quoted by Gent.

Those entries for which Gent expects an error greater than 1 per cent are enclosed in parentheses ( ). Those entries for which the divergence from experimental values is greater than 1.5 per cent are identified with an asterisk *. From independent tests, we have found the error in measurements to be less than 0.5 per cent. Two measurements, marked with a dagger †, were compared with an unpublished infinite series of Green, Curtis, and Mead from which (8) was derived, and found to agree within 0.1 per cent. We should therefore expect an asterisk * only where the entry is enclosed in parentheses ( ).

[4] J. W. Craggs and C. J. Tranter, "Capacity of two-dimensional systems of conductors and dielectrics with circular boundaries," *Quart. J. Math. (Oxford)*, vol. 17, pp. 138–144; 1946.
[5] H. H. Meinke, "Du Doppelleitung und der Sternvierer im zylindrisch begrenzten Dieliktrikum," *Eliktrische Nachrichten-Technik*, vol. 17, pp. 108–115; May, 1940.
[6] B. G. King, "Discussion on 'An investigation into some fundamental properties of strip transmission lines with the aid of an electrolytic tank'," *Proc. IEE*, vol. 104, pt. B, p. 72; January, 1957.

Over this range, none of these formulas has a maximum error greater than 8 per cent or less than 5 per cent. The formula with the smallest maximum error is (13), which happens also to be the simplest. The second-order approximation (12) of Craggs and Tranter is much better than the first-order approximation (13) for small values of s, but significantly worse for large values.

These small bits of experimental evidence do not support Gent's conclusion that (7) is valid over a wider range of the variables. On the contrary, they suggest the use of (13), which is simpler than and, over this range, just as accurate as any of the others.

We are indebted to Miss Alberta Strimaitis for the calculations whose results are given in the last four columns of Table I.

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T. C. Chen

Dr. Chen was engaged in development work on microwave relay systems at the RCA Victor Division, Radio Corporation of America, Camden, N. J., in 1945. He was then a research associate at the Moore School of Electrical Engineering of the University of Pennsylvania in 1946, concerned with research and development work on the EDVAC electronic computer. Since 1949, he has been associated with the Burroughs Corporation engaged in research and development of magnetic devices and electronic computing systems. Dr. Chen is currently in charge of the Engineering Techniques Department at the Burroughs Research Center, Paoli, Pa.

❖

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During 1949–1950, Mr. Clarke served as senior research fellow at the Microwave Research Institute of the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. Since 1950, he has been a member of the electrical engineering faculties of the Madras Institute of Technology, the University of Ceylon, Clarkson College of Technology, and the Polytechnic Institute of Brooklyn, where he is currently employed.

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In 1953–1954 Mr. Clarke served as a consultant to C.F.T. Engineering, Ltd., Colombo, Ceylon, on problems relating to the design and production of a low cost radio receiver. In 1957 he served as a consultant to Motorola, Inc., Chicago, Ill., on transistor circuitry and various pulse-code modulation and interference problems.

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Dr. Cohn is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

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Gregor Čremošnik was born in July, 1923 in Sarajevo, Yugoslavia. He received the M.Sc. degree in electrical engineering from the University of Ljubljana, Yugoslavia, in 1950. From 1950 to 1954, he was an assistant in the Electrical Engineering Department of the University of Ljubljana. He held a research fellowship in the Department of Advanced Electrical Engineering at the Swiss Federal Institute of Technology, Zürich, from 1954 to 1957. He received the Doctor of Tech.Sc. degree in 1957 from the Swiss Federal Institute of Technology.

G. ČREMOŠNIK

Dr. Čremošnik has done research on resistance networks as analog computers and has written several papers on analog computer methods. At the present time, he is a research engineer in the Department of Applied Physics at the Swiss Federal Institute of Technology, Section of Industrial Research (AfiF).

He is a member of the Physical Society, Zurich.

❖

Murray I. Disman (S'54) was born in Brooklyn, N. Y. on August 18, 1931. He received the B.E.E. degree from New York University in 1953. From 1953 to 1955, he worked as a research assistant at Stanford University on the phenomenon of frequency memory. Mr. Disman received the M.S. degree in electrical engineering in 1955 and is now completing the requirements for the Ph.D. At present, he is a research associate at the Electronics Research Laboratory at Stanford University, where he is doing research in the field of high-power traveling-wave tubes, which will be his dissertation topic.

M. I. DISMAN

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

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W. A. Edson (M'41–SM'43) was born in Burchard, Neb., on October 30, 1912. He studied electrical engineering at the University of Kansas, receiving the B.S. and M.S. degrees in 1934 and 1935, respectively. The following two years were spent at Harvard University on a fellowship. Upon receiving the D.Sc. degree in communication engineering in 1937, he joined the systems development department of Bell Telephone Laboratories.

W. A. EDSON

In 1941 Dr. Edson joined Illinois Institute of Technology as assistant professor of electrical engineering. He became professor of physics at the Georgia Institute of Technology in 1945, and professor of electrical engineering in 1946. From 1951 to 1952, he was director of the School of Electrical Engineering.

In 1952 he joined Stanford University as a research associate in the Electronics Laboratory and acting professor of electrical engineering. In 1954 he joined the newly-founded General Electric Microwave Laboratory at Stanford, where he now acts as consulting engineer.

He is a member of the American Physical Society, the American Association for the Advancement of Science, and the American Society for Engineering Education.

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Donald G. Fink, (A'35–SM'45–F'47), for a photograph and biography, please see page 2 of the January, 1958 issue of PROCEEDINGS.

❖

Arnim Frei was born on May 26, 1931 in Zürich, Switzerland. He attended high school at Zürich and the Swiss Federal Institute of Technology from 1952 to 1956. He received the M.Sc. degree in electrical engineering in 1956. Since 1956, he has been an assistant and research fellow in the Department of Advanced Electrical Engineering, Swiss Federal Institute of Technology, Zürich, Switzerland.

A. FREI

Mr. Frei has published papers on universal impedance curves and on resistance-capacitance networks as analog computers.

❖

Donald E. Johansen (M'57) was born in Los Angeles, Calif. on March 20, 1932. He received the A.B. degree in physics at the University of California at Berkeley in 1953. Since then he has done postgraduate work at the University of California and at Massachusetts Institute of Technology. From 1953 to 1954, Mr. Johansen was associated with Chromatic Television Laboratories, where he participated in the analysis of tricolor tube design parameters. For the next two years he served with the U. S. Army as a physical science assistant. His duties included digital computer programming and instrument design. Since May, 1956, Mr. Johansen has been with the technical staff of Hycon Eastern, Inc., where he has worked in the fields of control systems design and communications.

D. E. JOHANSEN

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Ralph C. Johnston (S'53–M'57) was born in Fremont, Neb. on May 22, 1933. He attended Iowa State College, Ames, Iowa, where he received the B.S. degree in electrical engineering in 1955. The following year he was a National Science Foundation Fellow at the Massachusetts Institute of Technology. From 1956 to 1958, he was connected with the M.I.T. Lincoln Laboratory as a staff associate and attended M.I.T. as a research assistant and as a full-time student. In 1958, Mr. Johnston received the S.M. degree in electrical engineering and the degree of Electrical Engineer.


R. C. JOHNSTON

He is a member of Tau Beta Pi and Eta Kappa Nu.

❖

Eugene W. Kinaman (M'51) was born in Albany, N. Y., on February 15, 1927. He received the B.S. degree in electrical engineering in 1949 from Rensselaer Polytechnic Institute and the M.S. degree in electrical engineering in 1953 from New York University. From 1949 to 1951, he was engaged in design work on the A. B. DuMont 5-kilowatt aural and visual television transmitter. From 1951 to 1953, he worked at New York University as a research assistant on several projects, including a microwave impedance plotter. He joined RCA Microwave Engineering in 1953 and is presently responsible for low-noise traveling-wave tube development and has technical responsibility for developmental sample tubes supplied by Microwave Engineering to customers.


E. W. KINAMAN

He is a member of Eta Kappa Nu.

❖

Max Magid was born on June 18, 1926, in Paterson, N. J. He received both the B.S.E.E. and M.S.E.E. degrees from the Newark College of Engineering, Newark, N. J. in 1950 and 1953, respectively. From 1950 until 1951, he was with the Boeing Airplane Company, Seattle, Wash., where he worked on the B-52 Electrical and Radio Systems. From 1951 until 1953, he was with the Western Union Telegraph Company in New York working on the development of a microwave radio relay system. Since 1953, he has been with the Radio Corporation of America, Microwave Tube Activity, Harrison, N. J., where he is engaged in application problems of traveling-wave tubes and magnetrons.


M. MAGID

Mr. Magid is a member of Tau Beta Pi.

❖

A. Papoulis (SM'55) was born in Greece, in 1921. He studied mechanical and electrical engineering at the Polytechnic Institute of Athens, Athens, Greece, from 1937 to 1942. He came to the U. S. in 1945 and continued his studies at the University of Pennsylvania, Philadelphia, Pa., where he received the M.S., M.A., and Ph.D. degrees.


A. PAPOULIS

Dr. Papoulis is an associate professor of electrical engineering at the Polytechnic Institute of Brooklyn, Brooklyn, N. Y., where he has been teaching since 1952, and he has been a consultant with the Burroughs Corporation since 1951. He also taught at the University of Pennsylvania and at Union College, Schenectady, N. Y.

❖

John N. Pierce was born on July 9, 1933 in Norwood, Mass. He received the Bachelor's and Master's degrees in electrical engineering at the Massachusetts Institute of Technology, Cambridge, Mass. in 1955.


J. N. PIERCE

Since 1952, Mr. Pierce has been employed at the Air Force Cambridge Research Center, Bedford, Mass., where he worked on uhf data transmission systems, long-range, antijamming systems, and high-reliability scatter transmission. At present he is engaged in studies of communication theory and reliability.

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Harrison E. Rowe (S'49–A'53) was born in Chicago, Ill. on January 29, 1927. He received the B.S. degree in 1948, the M.S. degree in 1950, and the D.Sc. degree in 1952, all from the Massachusetts Institute of Technology, Cambridge, Mass. From 1948 to 1952, he also served as a research assistant in the storage tube and microwave tube laboratories at M.I.T.


H. E. ROWE

Dr. Rowe joined the technical staff of Bell Telephone Laboratories in 1952, in the research department at Holmdel, N. J. He was initially associated with a group engaged in systems research. More recently, he has been working on mode conversion problems arising in multimode waveguides.

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

❖

Seymour Stein (SM'57) was born in Brooklyn, N. Y., in April, 1928. He received the B.E.E. degree from the City College of New York in 1949, and the S.M. and Ph.D. degrees in applied physics from Harvard University, Cambridge, Mass., in 1950 and 1955, respectively. In 1951–1952 and 1952–1953, he held an RCA Fellowship in electronics under the National Research Council.


S. STEIN

From 1953 to 1956, he was an engineering specialist with the Waltham Laboratories of Sylvania Electric Products, Inc. Since 1956, he has been with the technical staff of Hycon Eastern, Inc., where he has worked on a variety of problems, especially in propagation and communications system theory.

Dr. Stein is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

❖

Max J. O. Strutt (SM'46–F'56) was born on October 2, 1903 in Surakarta, Java. He studied at the University of Munich, Institute of Technology at Munich, and Institute of Technology at Delft, The Netherlands. From the latter he received the M.Sc. degree in electrical engineering and the Doctor of Tech.Sc. degree (cum laude) in 1926 and 1927, respectively. He was a research engineer at the Philips Company, Ltd., Eindhoven, The Netherlands, from 1927 to 1948. Since 1948, Dr. Strutt has been professor and director of the Department of Advanced Electrical Engineering, Swiss Federal Institute of Technology, Zürich, Switzerland.


M. J. O. STRUTT

Dr. Strutt holds more than 60 U. S. patents on electron tubes and circuits, especially at vhf and uhf. Among his awards are the Doctor Honoris Causa (1950) conferred by the Institute of Technology, Karlsruhe, Germany and the Carl Friedrich Gauss Medal (1954) of the Society of Sciences, Brunswick, Germany. In 1955 he was elected an honorary member of the Society of Sciences, Brunswick.

Dr. Strutt is a member of the Swiss Society of Electrical Engineers, the German Society of Electrical Engineers, the Swiss Society of Sciences at Berne, Switzerland, the German Physical Society, the Swiss Mathematical Society, and the Zürich Physical Society.

Scanning the Transactions

Transmitting digital data by telephone opens up potentially enormous possibilities for sending information accurately and at high speed between business machines and computers which are geographically remote from one another. An experimental system was recently announced by which tape recorded digital messages may be sent over ordinary telephone lines at a speed of about 750 words per minute. The magnetic tape record can be prepared by any of a number of business machines. In the actual tests of the system, a modified electric typewriter was used. At the receiving end the message is again recorded for subsequent processing by the machine at that end. The purpose of the magnetic tape is to make optimum use of the information capacity of the telephone line. For example, a typewriter operates at rates far below 750 words per minute. By recording the typewriter output, the message can then be speeded up to match the full capabilities of the telephone line. The prospect of machines telephoning one another all over the country is almost frightening to contemplate. Nevertheless, the potential demand for such a system is enormous. (W. A. Malthaner, "Experimental data transmission system," 1957 IRE WESCON CONVENTION RECORD, Part 8.)

Creativity. Much has been written of late concerning one of mankind's most precious commodities, how it arises, and how it may be fostered and best utilized. Among the many excellent definitions of creativity that have been offered, we think the one suggested recently by the co-founder and Editor Emeritus of the IRE is especially well worth repeating: "Creativity involves extracting from the crucible of truth something more than the sum of the components which have been put into it. Somehow, in the flame of both inspiration and reason, there will then have occurred a strange welding and combination of the contents of the crucible resulting in the new and, often enough, the unexpected." (A. N. Goldsmith, "Creative engineering methods," IRE TRANS. ON ENGINEERING MANAGEMENT, March, 1958.)

Is automation still regarded as a revolution? With a decade's experience behind us now, the answer seems to be no. For one thing, we can point to no radically new invention. We have seen instead the application of a number of old and previously well-known inventions. The computer, for example, represents no more than large scale circuit combinations of the 50-year-old electron tube, the century-old principle of Babbage's calculating "engine," and the 100-year-old symbolic algebra of Boole. Neither have we seen any sudden or widespread technological changes in industry as a whole. Only about 20 per cent of manufacturing is potentially adaptable to automation, and this adaptation will be a gradual process.

On the other hand, the impact of automation will be felt well beyond just the technological field. It is believed that as productivity increases, substantial segments of the nation's working force will be able to enjoy more leisure time with their families. Due to the increased complexity of machines, the skilled worker of today will be trained to be a more highly skilled technician tomorrow, and it is likely that ratio of male employees will increase in manufacturing plants while more women will remain at home. The incentive for greater intellectual attainment will no doubt be greater in an automation environment. Automation, then, is of broad significance to all fields of human endeavor—social, intellectual and moral, as well as the strictly technological and economic fields. If we cannot term automation a "revolution," perhaps we can refer to it as a "renaissance." (J. J. Lamb, "Automation—its social, moral, and spiritual implications," IRE TRANS. ON ENGINEERING MANAGEMENT, March, 1958.)

The measurement of bioelectric potentials generated within cells has become one of the many fruitful areas in which electronics is aiding biological and medical research. One of the relatively new tools for this work is the electrolyte-filled micropipette electrode. The measuring technique involves impaling a living cell upon the end of a microelectrode which has a tip diameter of only one micron or less, and amplifying the feeble pulses that arise from within the cell. Special amplifiers and associated circuits are required because of the extremely high resistance (100 megohms) of the electrodes and in order to neutralize input capacitance effects. These special requirements were recently the subject of a symposium, the results of which should provide important impetus to bioelectric research. (IRE TRANS. ON MEDICAL ELECTRONICS, March, 1958.)

What factors are important in hiring an engineer or scientist for research and development work? To shed some more light on this question a statistical study was made of hundreds of engineers and scientists employed by a large laboratory. Out of 27 factors analyzed, only eight were determined to be statistically significant in evaluating prospective employees. These eight factors related to the type of college degree, subject majored in, and the number of papers, honor societies, patents and fellowships involved. These results are not surprising. Of greater interest is the fact that, according to the study, such revered factors as class standing, participation in extra-curricular activities and particular college attended do not affect salary growth. (R. A. Martin and James Pachares, "Evaluating engineers and scientists for a research and development activity," 1957 IRE WESCON CONVENTION RECORD, Part 10.)

The use of the transistor as a switch has long been attractive. In fact, the old point-contact transistor was generally a better switch than an amplifier, to the dismay of circuit engineers seeking amplification. Interest in transistor and semiconductor diode switches, stimulated initially by computer and telephone switching developments, has since extended to a broad array of control circuit and data processing applications. It is significant that at the two major transistor conferences last year, the Transistor Circuits Conference in Philadelphia and the Semiconductor Device Research Conference in Boulder, transistor switches received the lion's share of attention. Four of the new devices discussed at the latter conference are reported in a current issue of TRANSACTIONS. Two of these switches operate in much the same fashion as a thyratron, *i.e.*, a very small power is used to control a much larger current flow, but with the added advantage that they can be readily turned off. (IRE TRANS. ON ELECTRON DEVICES, January, 1958.)

Automatic-tracking radar systems introduced the problem of analyzing sampled-data and its effect on control system performance. The increasing use of digital computers in control system applications has awakened further interest in the sampled-data field because digital computers can accept inputs and deliver outputs only at discrete instants of time. Special techniques have been developed to facilitate design of these systems and to predict their performance. Although a very extensive number of papers are available, relatively little theory of sampled data systems has been published in book form. To simplify the task of searching the literature to obtain either a general acquaintance with the field or to obtain information on specific aspects of it, a bibliography is now available which includes only that material which in the authors' opinion represents either a significant contribution to the field or has tutorial value. It is arranged alphabetically by author, and it has a subject index cross-referenced to the author index. This bibliography should be of interest to anyone who has an interest in systems which utilize intermittent data. (H. Freeman and O. Lowenschuss, "A bibliography of sampled-data control systems and Z-transform," IRE TRANS. ON AUTOMATIC CONTROL, March, 1958.)

Small transformers can be added to the growing list of electronic items that are being designed with the aid of computers. Aside from the obvious result of saving engineering time (the computer does the job in 2.5 minutes), some incidental advantages accrue from the use of computers. Designs tend to be more uniform and not dependent on individual whims. Therefore, they are easier to check and control. It is not difficult to envisage a further transition from the computer type of design to automatic manufacture and test of transformers. The day may come when it is possible to feed in the rating required and have computer-controlled automatic machinery produce transformers with little or no hand work anywhere along the line. (W. Etchison, M. B. Meunier, and R. Lee, "Computer design of small electronic transformers," IRE TRANS. ON COMPONENT PARTS, March, 1958.)

Books

An Introduction to Semiconductors by W. C. Dunlap, Jr.

Published (1957) by John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 407 pages+9 index pages+xxi pages. Illus. 9¼×6. $11.75.

According to its preface this book is intended for two separate groups. "The main group includes those engineers, technicians, and research workers who are entering upon active work with semiconductors . . ." and the second group includes "the student, perhaps in his senior year or in graduate school, who is interested in supplementing his usual course work in solid-state theory with material on both the scientific and the technical aspects of semiconductors." The material covered is perhaps indicated as well as it may be by a list of chapter titles:

1. Introduction
2. Some Facts About Crystals and Their Structure
3. Theory of the Solid State
4. Imperfections in Crystals
5. Statistical Mechanics for Metals and Semiconductors
6. The Electron Theory of Metals and semiconductors
7. Contact and Surface Properties of Semiconductors.
8. Properties of *p-n* Junctions
9. Experimental Measurements on Semiconductors
10. General Methods of Preparing Semiconductor Materials
11. Properties of the Elemental Semiconductors
12. Semiconducting Compounds
13. Rectifiers
14. Transistors
15. Photocells, Thermistors, Hall Effect, and Other Semiconductor Applications

Thus it will be seen that the range of topics is quite broad, so much so in fact that there is not space enough to really get down to cases with any of them. The treatment is not unduly mathematical although the theory behind each of the topics is presented. Here again the requirements of space have had their influence to the point where it is difficult to believe that a reader of either group to whom the book is directed can achieve enough mastery of the subject from this presentation alone to claim to understand, let alone to be able to use for further development. There are, however, satisfactory lists of references with each chapter that permit further pursuit of any chosen topic. The book contains no problems or exercises. The general format, index, and table of symbols are all good.

The number of errors does not seem to be excessive, eleven on a somewhat hasty reading, though a specialist might find still other points with which to quarrel. One puzzling error occurs in Figure 2.10 attributed to Lonsdale in which the symmetries shown for the cubic lattice are obviously incorrect. All in all the reviewer must confess to a certain unhappiness about the book. The intent is certainly praiseworthy and the book will be useful, but it would be so nice if it could also have the lucidity of some of the famous texts such as Courant-Hilbert, Richtmyer, or Sommerfeld.

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Handbook of Noise Control ed. by C. M. Harris

Published (1957) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 1012 pages+23 index pages. Illus. 9¼×6¼. $16.50.

Problems of acoustical noise affect the quality of all electronic, radio, television and electrical equipment. This new *Handbook of Noise Control*, edited by Professor Cyril M. Harris of the Electrical Engineering Department of Columbia University, provides an invaluable reference on all facets of the subject of acoustical noise and its control.

A general summary of this handbook would be a statement to the effect that the sub-title of this book could be "Acoustical Noise: what it is, its causes and cures."

Noise is defined as "unwanted sound"; the handbook, in forty sections, discusses all important aspects of the problem. Reference to this handbook will enable an electronic engineer to proceed rapidly to the understanding of, and solution of, acoustical noise problems in electronic equipment.

Some of the sections of particular interest to the radio engineering and allied professions are the following: "Physical Properties of Noise and Their Specification;" "Propagation of Sound in the Open Air;" "The Hearing Mechanism;" "The Loudness of Sounds;" "Principles of Vibration Control;" "Vibration Isolation;" "Vibration Damping;" "Vibration Measurement;" "Instruments for Noise Measurement;" "Noise Measuring Techniques;" and "Reduction of the Noise of Iron-Core Transformers and Choke."

Other subjects which are covered thoroughly, which one would normally not relate to electronic problems unless specifically indicated, are such topics as "fan noise," a field of importance in electronic equipment which today so frequently requires auxiliary forced air cooling.

Each of the forty sections, written by well-recognized authorities in the field of sound and vibration, is thoroughly illustrated, and all terms are carefully defined by Prof. Harris in the introduction. This group of definitions makes for unusual value when only certain sections are being used as references.

Equations are worked into the text in logical fashion, and tables are conveniently located to the referenced textual material. At the end of each section there is an extensive bibliography.

This new handbook of noise control will prove a boon to all in the radio engineering profession who must control acoustical noise, whether it is to reduce microphonics in sensitive electronic amplifying equipment, whether it is necessary to meet customer specifications, or whether one merely wishes, for esthetic reasons, to manufacture a piece of quiet equipment.

VICTOR WOUK
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Industrial Electronics Handbook, 2nd ed., by R. Kretzmann

Published (1957) by Philosophical Library, Inc., 15 E. 40 St., N. Y. 16, N. Y. 288 pages+4 bibliography pages+6 index pages+vii pages. Illus. 6¼×9 $12.00.

This book, a basic reference work, deserves a place on the book shelf of engineers and technical people dealing with electronic tubes, components and circuits in industry.

The first section describes the "building blocks"—the various types of electronic tubes and their basic circuits. Amplifying and transmitting tubes, rectifiers, thyratrons, senditrons, ignitrons and excitrons, photocells, trigger tubes and cathode-ray tubes are covered. Graphs, tables, photos and diagrams augment the text. Mathematical equations are helpfully used to describe the parameters involved.

The second section takes the "building blocks" previously stated and puts them into electronic devices for industrial purposes. The range of the devices covered is quite large and includes electronic relays, counting circuits, timers, rectifiers, lamp dimming, speed and temperature control, resistance welding control, motor control, induction and dielectric heaters and special devices such as ultrasonic soldering, inverters, and dust precipators. Approximately seventy circuits are diagrammed and an explanation of each circuit is included. Application diagrams and photos along with the rest of the material should be helpful to the reader who is searching for ideas to bring his problem quickly to a solution.

To cover the industrial electronics field adequately in a book of about 300 pages is very difficult but the author has done an excellent job of selecting the material. This book was published in the Netherlands and the tube references are European. Some tube data is given in the appendix. It would be more helpful if data were given on all the tubes shown in the circuits. An interchangeability list referring to U. S. tubes would enhance the usefulness of this book to readers in this country. However, in spite of this criticism, this book should prove a valuable source of information for those concerned with industrial electronics.

Industrial Electronics Circuits by R. Kretzmann

Published (1957) by Philosophical Library, Inc., 15 E. 40 St., N. Y. 16, N. Y. 187 pages+2 content pages+2 bibliography pages+2 index pages+2 pages Philip's Technical Library book lists+1–19×9 folded diagram. Illus. 6½×9. $10.00.

Industrial Electronic Circuits is a *must* companion volume to *Industrial Electronics Handbook* by the same author. The two books form a continuous whole. In this book there are over 90 separate circuits covering the general categories of photoelectrically controlled apparatus, counting circuits, temperature, voltage and level stabilizing circuits, contact and control devices, oscillator and amplifier circuits and rectifier and motor control circuits. The components used in the circuits are identified and there is a reasonable amount of detail in the explanation of the circuits. This should enable the reader to more thoughtfully evaluate the selected circuit for his particular needs. These features should enable the user to intelligently modify the selected circuit to conform to the specific demands of the problem at hand.

All of the circuits shown have been proved either in practice or in the laboratory. The selection of the circuits described is very good. While there is no pretense that every conceivable situation is covered in detail, the circuits shown should provide a practical, useful answer to most of the problems encountered in industry. By having this book on his shelf, the reader will save himself countless hours in developing circuits which the author has already compiled for the reader.

W. L. Atwood
Consulting Engineer
Warren, Ohio

Basic Electric Circuit Theory by W. W. Lewis with the assistance of C. F. Goodheart.

Published (1958) by The Ronald Press Co., 15 E. 26 St., N. Y. 10, N. Y. 643 pages+6 index pages+v pages. Illus. 6¼×9¼. $9.00.

This is a textbook for the electrical student at the intermediate level. Its organization follows the traditional approach. The aim is essentially to provide a basic understanding of fundamental analysis of electric circuits under both steady-state and transient conditions. Although the topics have been covered in many textbooks, this particular treatment is written with consummate clarity of thought. Careful consideration is given to the learning difficulties of undergraduates. Points likely to be misunderstood are fully stressed with excellent illustrations.

The subject matter has been divided into two major parts. Part I consists of nine chapters that cover the basic principles necessary for the solution of circuits. Part II consists of seven chapters that bring out important secondary principles and applications. Unique features of this book include a chapter on Laplace transforms, which supplements the classical solution of transients, a chapter on hyperbolic and exponential functions, and a chapter on short circuits utilizing symmetrical components. The mksa rationalized system of units is used throughout, based upon the adoption of the "ampere" as a fundamental unit by the International Electrotechnical Commission in 1956.

The first chapter centers itself around fundamental concepts associated with the generation, representation, and manipulation of alternating electromotive forces. The next chapter describes properties of time functions as applied to voltage, current, and power. The following one discusses parameters, their combinations in elementary circuits, and the concept of complex impedance and impedance triangle. The fourth chapter introduces complex numbers and the general theory underlying calculation with complex quantities. It is followed by a chapter on series circuits and a chapter on parallel and series-parallel circuits, which emphasize circle diagrams and resonance phenomena. Chapter Seven concerns itself with network aspects involving mesh analysis, transformations, theorems, and nodal analysis. The two final chapters of Part I deal with coupled circuits and mutual inductance, and polyphase circuits of the two-phase and three-phase variety including computational and measurement techniques.

The range of topics covered in Part II is quite impressive. At the beginning of Chapter Ten, there is a very straightforward exposition of nonsinusoidal waves and their representation by the Fourier series which is followed by essentials of Fourier analysis and the theory of circuit analysis involving nonsinusoidal quantities in single phase and polyphase systems. The authors then proceed to discuss in the next chapter mathematical aspects of hyperbolic and exponential functions required for the understanding of more advanced topics. In Chapter Twelve, transients are considered in d-c and a-c circuits utilizing the classical approach. It is followed by a chapter on the solution of transients by Laplace transforms. Classical filters comprised of prototype sections, *m*-derived full sections, and *m*-derived end-sections are dealt with in the next chapter to achieve desired impedance and attenuation characteristics. Chapter Fifteen considers transmission lines applicable to both the communications and power fields. The final chapter stresses various methods of evaluating short circuits and unbalanced faults, including the method of symmetrical components.

In attempting to keep a work of this kind within the space of one volume, the authors have had to use considerable judgment as to the choice of material. There is no question that, on the whole, the work is well balanced, that the selection of material is excellent, and that both the casual reader and the serious student will find it most appealing. The volume is distinguished for its good exposition and is certain to gain ready acceptance in curricula which follow traditional patterns.

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Closed Circuit TV System Planning by M. A. Mayers and R. D. Chipp

Published (1957) by John F. Rider, Inc., 116 W. 14 St., N. Y. 11, N. Y. 223 pages+2 pages of bibliography+12 pages of appendix+6 index pages+xii pages. Illus. 8½×11. $10.00

This is a book which endeavors to give a comprehensive description of the history, practice, and equipment of closed circuit television. In spite of this ambitious object, it succeeds admirably in many respects. It is divided into three main sections, entitled respectively "Application of Closed Circuit Television," "How Closed Circuit Television Works," and "Equipment." Within each of these broad classifications there is a rather jumbled arrangement of descriptive material.

While the book is professedly addressed to the user, who is not necessarily interested or qualified in the technical details of the equipment, there are a number of errors and confusions which should be cleared up in a second edition. Among these may be noted the statement on page 116 that "the luminance value of a monochrome picture on an average home receiver is in the order of 25 to 60 foot lamberts, making it possible to view the picture under conditions of normal room lighting without noticeable flicker." Surely what is meant is "without objectionable loss of contrast." On page 118, figure 2–20 should be corrected since the rays shown are inaccurate, or it should be omitted. On page 127, we read that the iris diaphragm controls the "quantity and distribution of the light which reaches the sensitive surface." What is meant by *distribution* is not clear. The formulas given on page 137 for bandwidth requirements do not agree;

actually the simpler formula is correct, while the more elaborate one (and hence ostensibly, the more accurate) has neglected blanking time. On page 140, in describing the angular coverage of a camera lens, it is not stated that the angular values given are for half the field—a very unusual method of specification, and inconsistent with the tables of horizontal field in feet, which are based on the full width. There are several other errors of this general nature, small but misleading at least to the non-technical reader.

Despite these criticisms, the book is felt by this reviewer to be of real value to those contemplating the installation and use of TV for closed circuit applications. Particularly to be commended are the many excellent photographs of commercial equipment and of typical installations. These certainly serve to display the great potential possessed by closed circuit television, and should be very helpful in system planning.

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Electrical Discharges in Gases by F. M. Penning

Published (1958) by The Macmillan Co., 60 Fifth Ave., N. Y. 11, N. Y. 68 pages+3 bibliography pages +3 index pages+viii pages. 29 figures. 9¼×6¼. $3.00.

This reviewer can do no better than to quote from the introduction by Dr. W. de Groot: "The publication, in 1955, two years after his death, of the first edition of his (Penning's) book *Electrische Gasontladingen*, soon prompted the view that the contents, set out with such remarkable clarity and with such a wealth of information, deserved to be made available to a wider circle of readers than they would find in Dutch-speaking countries alone." It is refreshing to find the fundamentals of the "classical" discharges in gases presented so clearly and briefly. This little monograph is an excellent introduction to the subject and may be read with profit by both the student, who will later pursue the subject in much more detail, and by administrators and engineers who feel the need of an introductory understanding of this old but still expanding field.

The processes discussed are those of the low-current, low-frequency or direct current discharges, *i.e.* "classical" gaseous conduction phenomena. Discharges at high currents and the more exotic effects at high frequencies in shock phenomena, in fusion studies, etc., are not considered. The titles of the short chapters indicate quite well the scope of the book, namely: Gas Discharges, Natural and Man-Made; The Conduction of Electricity in Metals and Gases; The Non-Self-Sustaining Discharges; the Movement of Electrons and Ions Through a Gas; the Non-Self-Sustaining Arc Discharge; The Townsend Discharge and Breakdown; Sparks and Lightning; The Glow Discharge; The Self-Sustaining Arc Discharge; The Positive Column. A short bibliography, mostly to the European literature, concludes the book. This book is certainly a worthy addition to the literature.

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Passive Network Synthesis by J. E. Storer

Published (1957) by McGraw-Hill Book Co., 330 W. 42 St., N. Y. 36, N. Y. 315 pages+3 index pages +x pages. Illus. 9¼×6¼. $8.50.

Dr. Storer in the preface to the book indicates as his aim "to provide a concise survey of the field." In the opinion of this reviewer, he succeeds admirably in the accomplishment of that aim. His book is fluently written and well illustrated by numerous examples of network synthesis. It provides a reasonably thorough survey of a large and highly developed analytical subject in no more than 315 pages.

The book is organized into four main sections: (1) driving-point impedance synthesis, (2) network synthesis using image parameters, (3) modern realization methods for two-terminal pair networks, and (4) rational-fraction approximations. Each of these sections will provide the novice with an insight into the field and the more advanced scholar with a delightfully succint condensation.

A thorough knowledge of network analysis is assumed as a prerequisite for reading the book. While the novice with inadequate mathematical background may follow the development of the many examples of synthesis, a full appreciation of the analytical subtleties requires a thorough knowledge of both the LaPlace transform and functions of a complex variable. This is particularly evident in the fourth main section, rational-fraction approximations. Unlike the preceding three sections, this section, while it is excellently written, is deficient in examples and it is so analytical as to be extremely difficult reading for the novice. The rational-fraction approximation section, in the opinion of this reviewer, is of benefit largely to the advanced scholar of network synthesis.

In surveying in 315 pages a field so complex and extensive as network synthesis, Dr. Storer has chosen his topics wisely and expressed himself carefully. However, in so compact a survey, it is inevitable some relevant topics will be omitted. Perhaps the most outstanding example of such an omission is the absence of development of the relationship between the even and odd parts of both driving-point and transfer functions of networks.

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Solid State Physical Electronics by Aldert van der Ziel

Published (1957) by Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. 593 pages+9 index pages+x pages. Illus. 8¼×6. $13.00.

The application of the physics of solids to practical engineering has blossomed so fantastically in the past few years that it is no longer adequate to treat solid state devices in an extra chapter of a standard "electronics" (vacuum tube) textbook. Even a book like the present one, which is wholly devoted to solid state devices, covers a tremendous range of material from oxide cathodes to solid state masers. In spite of the diversity of topics, however, the treatment is rarely superficial, so that the careful reader will obtain a solid understanding of the operation of solid state devices sufficient for most purposes other than actual research in the field. The discussion of photo and secondary emission is particularly good, as is the chapter on the physical basis of modern semiconductor amplifiers. Analysis and examples are built around representative devices, rather than around the physics *per se*. This approach runs the risk of rapid obsolescence, since new devices are invented weekly. However, sufficient discussion of basic principles is included to make the book useful to seniors, graduate students and practicing engineers and physicists for many years to come.

Considering the fairly detailed nature of the material, there are few references to the original literature. Beginning students may welcome this since it leads to a more unified approach to many topics, without ifs and buts. Advanced workers may have their doubts. Taken as a whole the book is a successful exposition of what is undoubtedly a difficult and complex subject.

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Princeton, N. J

Recent Books

Buchanan, J. P. *Handbook of Piezoelectric Crystals for Radio Equipment Designers (WADC Technical Report 56-156).* ASTIA Document Service Center, Knott Bldg., Dayton 2, Ohio.

Churchman, C. W., Ackoff, R. L., and Arnoff, E. L., *Introduction to Operations Research.* John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $12.00.

Cogan, E. J., and Norman, R. Z., *Handbook of Calculus, Difference and Differential Equations.* Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. $6.00.

DeFrance, J. J., *Electron Tubes and Semiconductors.* Prentice-Hall, Inc., 70 Fifth Ave., N. Y. 11, N. Y. $9.00.

Geyger, W. A., *Magnetic-amplifier Circuits.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $7.00.

Glasstone, Samuel, *Sourcebook on Atomic Energy*, 2nd ed. D. Van Nostrand Co., 257 Fourth Ave., N. Y. 10, N. Y. $4.40.

Jefferson, Sidney, *Radioisotopes, A New Tool for Industry.* Philosophical Library, Inc., 15 E. 40 St., N. Y. 16, N. Y. $4.75.

Lytel, Allan, *Printed Circuitry.* Instruments Publishing Co., Inc., 845 Ridge Ave., Pittsburgh 12, Pa. $2.00.

Proceedings of the EIA Symposium on Numerical Control Systems for Machine Tools. Engineering Publishers, Inc., G.P.O. Box 1151, N. Y. 1, N. Y. Obtainable from abroad at Interscience Publishers, 250 Fifth Ave., N. Y. 1, N. Y. $5.00.

Schure, A., *Basic Television*, Volumes One through Five. John F. Rider Inc., 116 W. 14 St., N. Y. 11, N. Y. Per vol., $2.25; per set, $10.00; set in cloth binding, $11.50.

The Radio Amateur's Handbook, 35th ed., 1958. American Radio Relay League, W. Hartford 7, Conn. $3.50 in the U. S.; $4.00, U. S. possessions and Canada; $4.50, elsewhere.

Volk, William, *Applied Statistics for Engineers.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $9.50.

Wilkes, M. V., *Automatic Digital Computers.* John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $7.00.

Abstracts of IRE Transactions

The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
Automatic Control	PGAC-4	$1.00	$1.50	$3.00
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Medical Electronics	PGME-10	0.90	1.35	2.70

*** Public libraries and colleges may purchase copies at IRE Member rates.**

Automatic Control

PGAC-4, March, 1958

Standard Terminology for Feedback Control (p. 1)

The Issue in Brief (p. 2)

How the Bandwidth of a Servo Affects Its Saturated Response—G. A. Biernson (p. 3)

All realizable servomechanisms saturate when large step inputs are applied. The response of the servo during saturation differs from the predicted linear response, and sometimes this may be deterimental to system performance. It is desirable to know when the saturation effects may be tolerated. This paper shows not only how the saturated response can be predicted from system parameters, but how system parameters can be selected to make the saturated response allowable for a step input which causes velocity and acceleration saturation.

By reducing a complex system to a simplified form, the paper illustrates that the number of overshoots which result from a velocity saturating step response is related to the maximum acceleration, the maximum velocity, and the gain-crossover frequency (bandwidth) of the primary servo loop. To limit the number of overshoots or to produce an optimum response, the bandwidth of the servo can be adjusted. However, if this reduces the bandwidth to an excessively low value, it is possible to use nonlinear compensation to allow a stable transient for a saturating input and yet maintain a high gain and bandwidth for small, nonsaturating inputs.

Analog Study of Dead-Beat Posicast Control—G. H. Tallman and O. J. M. Smith (p. 14)

This paper supplements an earlier paper, "Posicast Control of Damped Oscillatory Systems," by Smith in the September, 1957 issue of PROCEEDINGS. Although either paper may be understood without reference to the other, the interesting principles of posicast compensation can be more easily grasped if both papers are read. The paper in this issue describes analog tests on a system where the posicast compensation is varied from its nominal theoretical configuration to illustrate the effects of imperfect compensation that might be encountered in a practical application.

Posicast control is also described as a method of splitting the input excitation into fragments which are delayed in time before being applied to the system. When properly used, it eliminates oscillations and overshoot in the response of a servomechanism having very lightly damped poles. This can be done in a time considerably less than one cycle of the uncompensated oscillation. The frequency response of the compensated system can be made flat through and beyond the resonant frequency of the system. Compensation is considered for simple and complex systems, and for special inputs and disturbances anywhere in the forward path to the control system.

On Closed Form Expressions for Mean Squares in Discrete-Continuous Systems—Jack Sklansky (p. 21)

If a system is to be optimized with respect to the mean square of some variable, it is usually desirable to obtain a closed form expression for that mean square. This has been an unsolved problem in the optimization of discrete-continuous systems where both sampled-data and continuous subsystems are used.

A method of finding the desired closed form expressions is presented in this paper. It includes a complete introduction to the subject, and the essential steps in the mathematical procedure for analyzing, discrete-continuous systems with random inputs are included. The technique is illustrated by deriving closed form expressions for the output and ripple of discrete-continuous systems, and for the control error of a sampled-data feedback loop. An application to a "track-while-scan" system is also included.

Bibliography of Sampled-Data Control Systems and *Z*-Transform Applications—H. Freeman and O. Lowenschuss (p. 28)

This bibliography represents an attempt to simplify the task of searching the literature to obtain a general acquaintance with the field of sampled-data control and to obtain information on certain specific aspects of it. Due to the close association of the *Z*-transform and sampled-data systems, a number of papers dealing exclusively with the *Z* transform are included in the bibliography.

The bibliography includes only that material which in the authors' opinion represents either a significant contribution to the field or has tutorial value. The bibliography is arranged alphabetically according to the name of the author, or the first-named author in cases of coauthorship. A subject index with cross references to the author list is also provided.

Proposed IRE Standard Terminology for Feedback Control Systems (p. 31)

PGAC News (p. 32)

Roster of PGAC Members (p. 33)

IRE WESCON Papers Deadline Set for May 1, 1958 (p. 39)

Broadcast & TV Receivers

Vol. BTR-4, No. 2, March, 1958

Development of the 12-Volt Plate-Voltage Hybrid Automobile Radio Receivers—AM, Signal Seeker and FM—Chih Chi Hsu (p. 1)

Local Oscillator Radiation from TV and FM Sets—W. G. Peterson (p. 32)

Spurious radiation that is generated by any type of superheterodyne type of receiver has presented a mammoth problem to the industry. That is, limits set up by the FCC had to be met by the industry. The amount of work that has been done by the industry toward this end has been tremendous.

It is the scope of this paper to bring to light some of the more significant points of interest in this field of endeavor, as well as some of the problems of accurate measurement techniques evolved through recent years.

Techniques Involved in Meeting FCC Radiation Requirements at UHF—John Bell (p. 38)

To meet FCC radiation limits, uhf strips and tuners must be designed to take full advantage of normal double-tuned circuit skirt attenuation so that a prohibitive increase in cost and complication can be avoided. UHF strips must also meet the normal radiation limits on the oscillator fundamental, as well as the limits on used and unused harmonics, on both the vhf and uhf antenna connections. Techniques and typical designs are described, and typical field measurements are given.

Minimizing the Effect of Cutoff in TV Vertical Oscillators—S. F. Love (p. 52)

The vertical oscillator of a TV set, whether a multivibrator or a blocking oscillator, can be very dependent upon the cutoff characteristic of the tube. It can affect the speed or frequency of the oscillator, the linearity and the synchronizing characteristic. There are several classes of cutoff current of the tube, namely that which is controlled by the grid voltage and that which is uncontrolled or "dead" plate current. Both of these have different effects on the performance of the circuit and tube. Some circuits are sensitive to cutoff currents as low as 10^{-4} times the peak current drawn during the cycle. Circuits can be modified to mitigate the effect of tube cutoff and several ways of doing this are described in the paper.

Automatic Fine Tuning Circuitry in Television Receivers—K. E. Farr and L. J. Sienkiewicz (p. 63)

Design of practical automatic fine tuning circuitry for production receivers involves problems of interaction with many of the functions of a television receiver including the tuner, IF amplifier, video and intercarrier sound stages, and automatic gain control circuits. This paper describes a receiver incorporating solutions to these problems as well as a method of programmed automatic fringe tuning.

Analysis and Synthesis of Magnetic Yokes Using Rotating Probes—H. S. Vasilevskis (p. 74)

A Synchronous Detector Using a Harmonic Pair Switching Wave—S. K. Altes (p. 88)

A synchronous detector is described in which the nonlinear element generates its own reference carrier from a pair of adjacent harmonics. This design is shown to offer many advantages, especially in triode circuits.

Some Notes on the Hybrid-Pi Transistor Equivalent Circuit—C. R. Wilhelmsen (p. 92)

Component Parts

Vol. CP-5, No. 1, March, 1958

Information for Authors (p. 1)

Photoelectric Cells—A Review of Progress—J. D. McGee (p. 2)

This paper is a tutorial review of the various devices that exhibit a photoelectric effect. An exposition of the underlying physical principles that produce photoelectric effects is presented.

Stability Characteristics of Vitreous Enamel Dielectric Capacitors—B. L. Weller (p. 24)

The critical dependency of all electrical devices upon their environment makes temperature stability one of the most important assets of any component. In this realm vitreous enamel dielectric capacitors have unusually predictable properties. This paper presents the results of experiments over a number of years and defines the ability of this component to retrace its capacity characteristic.

Aircraft Motor Generator with Secondary Standard Frequency Output—L. J. Johnson and S. E. Rauch (p. 28)

The paper describes an experimental ac motor alternator assembly which produces an output frequency exactly equal to a low-power secondary standard frequency signal source, such as a tuning fork or crystal oscillator. The frequency-controlled system employs a synchronizing type of control in contrast to the more generally employed error-detecting systems. As a result, no error exists between the output frequency and the secondary standard.

The control system requires only low-level signal power from the secondary standard reference. The main synchronizing power is drawn from the unregulated feeder line. No mechanical devices are needed for frequency correction or synchronization. The synchronizing system is completely electrical, requiring no mechanical or hydraulic compensating devices. The mechanical components employ only a single moving part, composed of rotor shaft and two bearings. The electrical energy supplied to the rotor is magnetically induced across air gaps so that no sliding contacts or brushes are used. Transient loads applied to the system do not produce any frequency transients; in other words, transient and steady-state frequencies are identical.

Characteristics of a 500-va experimental unit are presented.

Permanent Magnets in Audio Devices—R. J. Parker (p. 32)

The permanent magnet is considered as a component for changing the form of energy, and a brief review of the basic physics of the permanent magnet is included with emphasis on the nature of the magnetization process and how the permanent magnet functions in establishing external magnetic field energy.

Presently available characteristics of permanent magnets and future possibilities for improving the efficiency of the permanent magnet are discussed as well as the relationships between audio device performance and the unit properties of permanent magnets.

In using the permanent magnet, the choice of unit properties, volume, geometry, and magnetic circuit arrangement greatly influence the end performance and efficiency of audio devices. As an aid in exploiting the optimum combination of these variables an electrical analog system using lumped constants is introduced. Data on leakage permeance are presented for the more widely used permanent magnet arrangements in audio work. The analog technique is of general interest from the viewpoint of understanding the energy relationships involved in the efficient application of the permanent magnet and as an aid in predicting permanent magnet performance on a firm engineering basis.

A New Concept of Temperature-Rise Measurement of Transformers—Abraham Rand (p. 37)

Present methods of measuring transformer temperature rise require instrumentation with a high degree of accuracy, expensive temperature-control chambers, and relatively skilled operators. A new approach and method is described which provides equal or better accuracy while employing less accurate instruments, a lesser degree of operator skill, and without resorting to temperature-controlled test chambers.

This paper describes a bridge test circuit which utilizes a calibrated temperature-rise "standard" which acts as a reference in evaluating a rest unit. The bridge circuit provides a measure of compensation for varying electrical and ambient conditions. Data and curves are presented which indicate that the constant-temperature chamber could be dispensed with, except for the purpose of calibrating reference "standards."

Computer Design of Small Electronic Transformers—W. Etchison, M. B. Meunier, and R. Lee (p. 43)

Electronic computer program development for small power transformers is described. The methods used are applicable to simple open type designs rated up to 3 kw and up to 1000 volts working with primary and secondary taps. Computer operating time averages 2.5 minutes; most of this is the time required for feeding in the rating data and reading the computer output. Design information includes core, turns, turns per layer, wire size, winding height, winding resistances, regulation, secondary terminal voltage, and total of core and coil losses. An example is given to illustrate the procedure and some incidental advantages of computer design are indicated.

The Exact Design of Two Types of Single-Crystal, Wide-Band Crystal Filters—T. R. O'Meara (p. 53)

It is shown how a low-pass ladder filter may be transformed to one of two equivalent band-pass, unbalanced, crystal filter structures with three poles and two transmission zeros. The first structure is the bridged T, previously discussed by Mason on an image parameter basis, and the second is a ladder filter.

Either type may be designed on an exact insertion loss basis, and both types are equivalent if the Q's of the associated inductors are infinite. With finite inductor Q's one may "exactly" compensate for inductor losses with the bridged T while one may only partly compensate for inductor losses with the ladder. In spite of this, it is shown that the ladder filter does not suffer too large a degradation in its insertion loss function with reasonable Q's. Practical constraints on both types of filters are discussed.

Magnetostrictive Delay Line for Video Signals—G. I. Cohn, L. C. Peach, M. Epstein, H. O. Sorensen, and D. P. Kanellakos (p. 53)

Initial development of a magnetostrictive delay line capable of delaying an IF carrier which is amplitude modulated by radar video frequencies is presented. Pulses of IF energy up to 7 mc have been successfully delayed. A typical delay line operated at 5 mc has an insertion loss of 82 db for 70-μsec delay and an attenuation of approximately 8 db for each additional 70-μsec of delay. This line has the advantage that it can be conveniently provided with multiple pick-offs.

New Rugged Multicontact Connectors—J. Spergell and M. Pomerantz (p. 60)

Current and anticipated Army field tactics, requiring rugged communication ground equipment to withstand a wide range of environmental conditions such as humidity, temperature extremes, shock, vibration, sand, and mud, with a high degree of mobility, have established the need for a series of reliable general purpose, multicontact, power connectors which are capable of meeting these requirements. The AN series of connectors which have been employed to a great extent on military electronic equipments are considered unsuitable for present-day needs because of their high insertion and withdrawal forces, tedious coupling mechanisms, relatively high contact resistance characteristics, susceptibility to contact damage, poor mounting features, large number of maintenance parts, variety of connector classifications for specific applications, and general lack of ruggedness.

The Signal Corps Engineering Laboratories in conjunction with the Scintilla Division of the Bendix Aviation Corporation have recently completed a five-year program which resulted in the development of a new series of connectors, designated as the Q series, to replace the AN series for Army applications. This paper presents the design features and performance data on the Q connectors with emphasis being placed in those features which contribute to its ruggedness and reliability.

Contributors (p. 68)

Electron Devices

Vol. ED-5, No. 1, January, 1958

Editorial (p. 1)

The "Thyristor"—A New High-Speed Switching Transistor—C. W. Mueller and J. Hillibrand (p. 2)

A description is given of the construction and characteristics of a versatile and novel semiconductor device, called a "Thyristor," that may be operated as a bistable element switching to a high conductivity mode or as a more conventional high-frequency transistor, either in switching or amplifying circuitry. The Thyristor has thyratron-like characteristics that closely approach those of an ideal switch. However, the Thyristor, unlike the thyratron, can be turned off readily by the control element.

The open-state current is about 2 microamperes and the closed-state voltage drop is 0.3 to 0.5 volt. The unit can be switched into the high conductance mode in less than 0.1 microsecond with a pulse energy of 10^{-4} ergs. It can be turned off with a pulse energy of about 10^{-1} ergs in times of the order of 0.1 microsecond.

The bistable operation depends upon a new type of semiconductor contact that collects holes at low current densities and injects electrons of high current densities. The electron alpha of the injector increases as a power law function of current and greatly aids in obtaining device reproducibility. These injector properties can be described in terms of a funneling mechanism.

A New High Current Mode of Transistor Operation—C. G. Thornton and C. D. Simmons (p. 6)

An analysis is given of a new type of solid-state phenomena which appears as an abrupt transition to a low voltage circuit mode at high current densities under appropriate conditions. It is shown that a knowledge of this effect makes possible not only the consideration of a new family of thyratron devices, but also an understanding of heretofore inexplicable failures that have appeared in the form of collector-to-emitter shorts under particular circuit conditions.

Three-Terminal *P-N-P-N* Transistor Switches—I. M. MacIntosh (p. 10)

An investigation of the electrical properties of four-region silicon structures, with electrical contact made to both outer regions and to one of the inner base regions is described briefly. A satisfactory analytical understanding of the device has been achieved, but for simplicity the experimental results presented are discussed in nonmathematical and purely physical terms. The fuller theoretical discussion is being prepared for submission to the PROCEEDINGS OF THE IRE. In many respects, the behavior of this three-terminal device is found to be similar to the conventional thyratron.

Germanium Power Switching Devices—J. Philips and H. C. Chang (p. 13)

Two-terminal and three-terminal germanium power switching devices have been developed, utilizing a metal semiconductor contact as an electron injector in a multijunction device. The principles of operation, fabrication techniques, and electrical characteristics of this new device are discussed. Devices capable of switching up to 25 amperes and blocking up to 350 volts have been fabricated and applied to power control circuits. The low impedance voltage drop is of the order of 0.5 volt and the dynamic resistance is a few hundredths of an ohm. A switch-on time less than 0.1 microsecond has been measured with a switch-off time of the order of microseconds.

High Field Emission in Germanium Point-Contact Diodes—G. Wallis and J. F. Battey (p. 19)

The effects have been examined of small changes of barrier height on the reverse current of high inverse voltage germanium point-contact diodes. The barrier height was varied by changing the ambient gas. The experimental results are given quantitative interpretation by a field emission theory in the region from about 10 volts to about 150 volts, where other effects become important. Examination of some 50 diodes has shown that diodes which draw low reverse currents behave according to this theory.

A Wide-Band Bridge Yielding Directly the Device Parameters of Junction Transistors—J. Zawels (p. 21)

A method is described for determining on a bridge the nine elements of an equivalent circuit for junction transistors which is accurate at both low and high frequencies. These nine-device parameters, which include the collector-to-base interelectrode capacitance, could be determined in four independent steps of a multi-position switch, but for the sake of limiting the complexity of the equipment, five steps have been employed. All measurements are based entirely on the equivalent circuit and not on the choice of a particular set of four-pole quantities such as the z, y of h parameters. Fortuitously, the transistor terminations indicated are those most easily achieved in practice, *viz.*, a short circuit on the collector side and an open circuit on the base (or emitter) side.

The Parameters of Nonlinear Devices from Harmonic Measurements—F. Haber and B. Epstein (p. 26)

A method is discussed for determining the parameters of a nonlinear device by measurement of the harmonics generated in the output when a sinusoidal wave is applied at the input. Formulas are developed, in terms of the measured quantities, for determining the coefficients of the power series describing the input-output characteristic and the coefficients of the Fourier series giving the harmonic conversion transconductance.

Leak Detection—Ultrasensitive Techniques Employing the Helium Leak Detector—J. L. Lineweaver (p. 28)

The development of closures for electron devices often requires that leaks far below the usual range of the helium mass spectrometer-type leak detector be localized and measured. It has been possible, by using new techniques, to extend the sensitivity of detection several orders of magnitude beyond the normal limit of the instrument. Two highly sensitive methods with similar techniques but separate sensitivity ranges are discussed.

In seals similar to those of color television picture tube closures, it is noted that the character of small leaks is such that long times are required to reach a constant maximum rate of helium leakage. Methods for predicting this maximum rate from early observations are discussed. Methods for establishing procedures necessary for sensitivity requirements for any particular seal type are given. Time and sensitivity limitations are discussed.

Increases in the sensitivity of detection of the order of 40,000 are attained easily. Leaks as small as 1.5×10^{-10} μliters/second have been detected and measured in 22-inch rectangular color television picture tube closure seals. In one year, a leak of this size would only cause an increase in pressure in an ungettered tube of 10^{-7} mm Hg. This represents an increase in the sensitivity of detection of about one million.

The Design of Periodic Magnetic Focusing Structures—J. E. Sterrett and H. Haffner (p. 35)

Charts are presented which facilitate the design of permanent-magnet periodic structures for focusing electron beams. These charts include curves showing the peak magnetic field required for periodic focusing in terms of the electron-beam parameters, the magnet and pole-piece dimensions required to obtain this peak field, and the resultant weight of the focusing system. Thus, sufficient information is given to design completely a periodic permanent-magnet focusing structure for a given electron beam. The design of a focusing structure for a 1-watt X-band tube is carried through in detail to exemplify use of the charts.

Experimental Notes and Techniques (p. 43)

Book Reviews (p. 45)

Contributors (p. 46)

Engineering Management

VOL. EM-5, NO. 1, MARCH, 1958

The Teacher Is a Grafter—C. S. Manning (p. 1)

Discussion—C. R. Burrows (p. 2)

Engineering Management Development—G. W. Jernstedt (p. 3)

Automation—Its Social, Moral, and Spiritual Implications—J. J. Lamb (p. 6)

Help Wanted: U.S. Brainpower—H. L. Bevis (p. 10)

Administration of a Military Communications-Electronics Systems Engineering Function—C. J. Schauers (p. 12)

A Report on the Organization and Management of a Research and Development Technical Planning Group—A. H. Hausman (p. 19)

Creative Engineering Methods—A. N. Goldsmith (p. 22)

WESCON Papers Deadline Set for May 1, 1958 (p. 27)

Medical Electronics

PGME-10, MARCH, 1958

Foreword (p. 1)

Program of Symposium on Electrodes and Amplifiers in Biological Research (p. 2)

Measurement of Bioelectric Potentials With Microelectrodes and Neutralized Input Capacity Amplifiers—Ernest Amatniek (p. 3)

The biological aspects of bioelectric potentials are reviewed, equivalent electric circuits of the bioelectric generator derived, and grid current magnitude deleterious to a living cell evaluated. The pipette microelectrode is superior to the metallic microelectrode for simultaneous recording of resting and action potentials. However, the low-pass circuit formed by the high electrode resistance (1 to 100 megohms) and the amplifier input capacity causes distortion of the action potential waveform. The distortion can be reduced by amplifier equalization, capacity reduction, and capacity neutralization. Some of the limitations of these methods are: 1) Compensation for waveform distortion due to the low-pass circuit results in an increase of the amplifier noise, approximately as the inverse characteristic of the low-pass circuit. A filter for improving the signal-to-noise ratio is considered. 2) Neutralization is accomplished with a generated negative capacity which approximates a pure capacity only over a limited bandwidth. This limitation may be evaluated with a six-element synthetic circuit closely approximating a realizable negative capacity. 3) The accuracy of neutralization adjustment is inherently limited by the difficulty of inserting a calibrating voltage in place of the bioelectric generator. Moreover, the distributed capacitance of the microelectric cannot be exactly neutralized by a pure negative capacity. Despite these limitations, neutralization is the most applicable of the methods discussed, and highly successful neutralized input capacity amplifiers have been designed and used in several laboratories. Vacuum tube and transistorized versions of neutralized amplifiers are briefly described.

The Use of Transistors in Physiological Amplifiers—E. F. Macnichol, Jr. and T. Bickart (p. 15)

Preliminary experiments have indicated that it is feasible to construct low-level dc differential amplifiers using inexpensive hearing aid transistors. The noise level of such amplifiers was found to be roughly proportional to source impedance; it is less than that of a vacuum tube circuit for sources having an impedance of less than 10,000 ohms.

A physiological amplifier having a voltage gain of 10,000 was designed on the basis of these experiments. Several of these units have been constructed which show a bandwidth of about 60 kc and a noise level of about 2 μv rms with a 50-ohm input. For measuring nerve potentials a pair of electrometer tubes connected as cathode followers are used as an input probe. With this circuit arrangement the noise level is the order of 15 μv.

On Amplifiers Used for Microelectrode Work—C. C. Yang, J. P. Hervey, and P. F. Smith (p. 25)

Footnotes on a Headstage—J. Y. Lettvin, B. Howland, and R. C. Gesteland (p. 26)

This electrometer pulls itself up by both boot straps.

WESCON Papers Deadline Set for May 1, 1958 (p. 28)

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 Index to the Abstracts and References published in the PROC. IRE from February, 1957 through January, 1958 is published by the PROC. IRE, May, 1958, Part II. It is also published by *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, and included in the March, 1958 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. 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.

ACOUSTICS AND AUDIO FREQUENCIES

534.1:541.135 991
U-Effect, II, an Electrokinetic Phenomenon—W. W. Fain, S. L. Brown, and A. E. Lockenvitz. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 902–908; August, 1957.) "When a glass capillary tube filled with a column of alternate layers of mercury and an electrolytic solution is forced to vibrate mechanically, an ac voltage is generated across the ends of the column. Certain aspects of this electrokinetic phenomenon have been investigated both experimentally and theoretically."

534.2–14 992
Comparison of Experimental Underwater Acoustic Intensities of Frequency 14.5 kc/s with Values Computed for Selected Thermal Conditions in the Sea—F. H. Sagar. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 948–965; August, 1957.) See 1277 of 1956.

534.2–8 993
Rendering Standing Ultrasonic Fields Visible and Acoustic-Optical Image Conversion—G. Keck. (*Acustica*, vol. 6, no. 6, pp. 543–548; 1956. In German.) A survey of existing with further details of new photographic methods [see, *e.g.*, 963 of 1956 (Hauer and Keck)].

534.23+621.396.677 994
Some Aspects of the Design of Strip Arrays—D. G. Tucker. (*Acustica*, vol. 6, no. 5, pp. 403–411; 1956.) The design and performance of linear arrays are examined from the point of view of far-field directional patterns. Patterns are synthesized by the linear superposition of curves of the sin x/x type; this method clarifies the relation between directional pattern and the distribution of excitation (or sensitivity, in the case of reception) over the length of the array.

534.232:537.228.1 995
Piezoelectric Transducers—A. C. Dobelli. (*Acustica*, vol. 6, no. 4, pp. 346–356; 1956.) Review of the properties and parameters of the most widely used piezoelectric materials. Typical applications are mentioned and the choice of materials and the appropriate dimensions for the element are discussed. Reference is made to recent developments of piezoelectric polycrystalline ceramics.

534.232:621.317.35 996
Measurements on Electromechanical Transducers—Diestel. (See. 1214)

534.232:621.372.5 997
The Ideal Forms of Electroacoustic Transducers and the Characteristics of Coupled Systems Formed by Them—F. A. Fischer. (*Acustica*, vol. 6, no. 5, pp. 421–424; 1956. In German.) The ideal form of the four classes of electroacoustic transducer (piezoelectric, dielectric, electrodynamic, and electromagnetic) is discussed. The characteristics of the electric quadripoles resulting from the mechanical coupling of two ideal transducers of identical or different classes are investigated.

534.52 998
Scattering of Sound by Sound—P. J. Westervelt. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 934–935; August, 1957.) Earlier analysis of the mutual nonlinear interaction of two plane sound waves (2657 of 1957) is extended to include arbitrary directions of travel of one wave with respect to the other. An exact solution for the first-order scattering process is obtained.

534.612 999
Acoustical Radiation Pressure due to Incident Plane Progressive Waves on Spherical Objects—G. Maidanik and P. J. Westervelt. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 936–940; August, 1957.)

534.615 1000
Automatic Recorder of Lines of Equal Phase Change—V. Gavreau and A. Calaora. (*Acustica*, vol. 6, pp. 539–542; 1956. In French.) Apparatus for tracing curves of equal phase change in a plane section of a three-dimensional sound field is described. Its application to the sound field of a loudspeaker and the effects of interposed obstacles are illustrated.

534.641:621.395.92 1001
The Acoustical Impedance Presented by some Human Ears to Hearing-Aid Earphones of the Insert Type—J. Y. Morton and R. A. Jones. (*Acustica*, vol. 6, pp. 339–345; 1956.) Report of measurements made over the frequency range 220 cps–4 kc on persons of normal hearing. The impedance of the ear at the ear drum is calculated from that measured 1.6 cm from the ear drum with a complete seal between the ear mould and the ear canal. A continuously variable acoustic impedance developed for the investigation is described.

534.75 1002
Recorded Group Audiometer Test Comparisons at the 1956 Southern California Exposition—J. C. Webster and P. O. Thompson. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 895–899; August, 1957.)

534.75:621.391 1003
The Basic Elements for Determining Informing Capacity in Hearing—E. Zwicker. (*Acustica*, vol. 6, no. 4, pp. 365–381; 1956. In German.) Audibility changes are considered as a function of just perceptible levels of AM and FM for pure tones or white noise. By means of a model based on the perception of a change in sound intensity of 1 db within a group of frequencies with a mean time constant of 20 ms, the effects of phase differences as well as masking are explained. The maximum information perceptible is estimated and its application to the intelligibility of syllables is illustrated.

534.846 1004
Orthophonic Surfaces in Auditorium Design—C. Codegone. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 885–888; August, 1957.) "A study is made, in a particular case, of the shape to be given to the ceiling of a large auditorium such that the sum of the direct intensity and the intensity reflected by the ceiling is constant."

621.395.61:534.844.1 1005
Investigation of the Directional Characteristics of the Pick-Up Microphone in Pulse Measurements of Room Acoustics—H. Niese. (*Hochfreq. u. Elektroak.*, vol. 65, pp. 192–200; May, 1957.) A stereophonic microphone simulating the directional characteristics of the human head is described, which is used in tests to assess subjectively the effect of the angle of incidence of echoes, and of amplitude differences at the ears. See also 2658 of 1957.

621.395.623.5 1006
Wedge-Shaped Acoustic Horns for Underwater Sound Applications—C. M. McKinney and W. R. Owens. (*J. Acoust. Soc. Amer.*, vol.

29, pp. 940–947; August, 1957.) Report of measurements on horns constructed of rubber-covered Al plates [1241 of 1955 (McKinney and Anderson)] modified to operate at 10 kc.

621.396.623.64+534.833 1007
Attenuation of Ear Protectors by Loudness Balance and Threshold Methods—J. Hershkowitz and L. M. Levine. (*J. Acoust. Soc. Amer.*, vol. 29, pp. 889–894; August, 1957.) Results of measurements on earmuffs and earphone sockets using three different methods are compared.

621.395.623.7 1008
Design for a Folded Corner Horn—H. J. F. Crabbe. (*Wireless World*, vol. 64, pp. 57–62; February, 1958.) Details are given of the construction of low- and middle-frequency horns in a compact unit to match on 8-inch speaker, for domestic use.

621.395.625.3 1009
The Infrared Transparency of Magnetic Tracks—G. Lewin. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 66, pp. 517–522; Discussion, p. 522.) Magnetic tracks are substantially transparent to infrared light. The infrared sensitivity of the PbS photoconductive cell enables excellent reproduction to be obtained from an optical track completely covered by a magnetic stripe, eliminating the need for half-width stripes and giving superior magnetic recording quality.

621.395.625.3:389.6 1010
The International Standardization of Magnetic-Tape Recording—P. H. Werner. (*Tech. Mitt. PTT*, vol. 35, pp. 266–273; July 1, 1957. In French and German.) Summary of the principal international standards for professional and amateur recordings, including CCIR recommendations.

621.395.625.3:681.85 1011
Magnetic Recording Tape—Manufacture and Properties—H. G. M. Spratt. (*Brit. Commun. Electronics*, vol. 4, pp. 418–421; July, 1957.) Tables of representative British recording tapes and magnetic-tape base materials are given.

ANTENNAS AND TRANSMISSION LINES

621.315.212 1012
Coaxial Transmission Lines—S. Mahapatra. (*Electronic Radio Eng.*, vol. 35, pp. 63–67; February, 1958.) Approximate calculations are made for the distributed constants R, L, G, and C of a coaxial transmission line with the inner conductor of elliptical cross section. In conditions of restricted space higher Q and lower attenuation may be possible than with a conventional coaxial line.

621.315.212.029.63:621.316.543 1013
A Contactless Wide-Band Switch for 20-cm Wavelengths—L. Mollwo. (*Hochfreq. u. Elektroak.*, vol. 65, pp. 181–188; May, 1957.) The mechanical construction and electrical characteristics of a rotary-type coaxial-line switch with matching stubs are detailed. The pass band is 80 mc at 19.5 cm λ with more than 20-db attenuation of the blocked circuit.

621.315.212.029.63:621.372.512] +621.372.832.43 1014
Criteria for the Design of Loop-Type Directional Couplers for the *L* Band—P. P. Lombardini, R. F. Schwartz, and P. J. Kelly. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 234–239; October, 1956. Abstract, PROC. IRE, vol. 45, pp. 113–114; January, 1957.)

621.372.029.64:537.226 1015
The Use of Dielectric Materials to Enhance the Reflectivity of a Surface at Microwave Frequencies—G. B. Walker and J. T. Hyman. (*Proc. IEE*, pt. B, vol. 105, pp. 73–76; January, 1958.) The condition that the reflection at the dielectric shall be better than at the (metal) surface alone is derived. As a test, a disk of TiO_2 of thickness $\lambda/4$ was built into a metal cavity, and was estimated to produce a reflection about twice as great as that from the associated metal surface alone.

621.372.2 1016
Open Wire Lines—G. Goubau. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 197–200; October, 1956. Abstract, PROC. IRE, vol. 45, p. 113; January, 1957.)

621.372.2+621.372.8]:537.226 1017
Propagation of Microwaves along a Single Conductor Embedded in Three Coaxial Dielectrics—Part 2—S. K. Chatterjee and R. Chatterjee. (*J. Indian Inst. Sci.*, sect. B, vol. 39, pp. 71–82; April, 1957.) "The characteristic equation for the EH wave has been derived. It is shown, as a special case, that the asymmetric wave EH_1 cannot be propagated along a solid conductor embedded in free space due to very high attenuation. Field components in terms of the axial power flow have been derived." Part 1: 16 of 1957.

621.372.413+621.372.8]:518.5 1018
Special Slide Rule for Calculating the Internal Wavelength of E. M. Waves in Waveguides and Cavity Resonators—W. Otto. (*NachrTech.*, vol. 7, pp. 294–296; July, 1957.)

621.372.8 1019
Contribution to the General Theory of Irregular Waveguides—B. Z. Kastenelenbaum. (*Dokl. Ak. Nauk S.S.S.R.*, vol. 116, pp. 203–206; September 11, 1957.) Waveguides of nonuniform circular cross section are examined and their main characteristics obtained by an approximation method. See also 676 of 1957.

621.372.8 1020
Miniaturization of Microwave Assemblies—L. Lewin. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 261–262; October, 1956. Abstract, PROC. IRE, vol. 45, p. 114; January, 1957.)

621.372.821:621.372.86 1021
The Excitation of Surface Waveguides and Radiating Slots by Strip-Circuit Transmission Lines—A. D. Frost, C. R. McGeoch, and C. R. Mingins. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 218–222; October, 1956. Abstract, PROC. IRE, vol. 45, pp. 113; January, 1957.)

621.372.825 1022
A Note on the Fourier Series Representation of the Dispersion Curves for Circular Iris-Loaded Waveguides—P. N. Robson. (*Proc. IEE*, pt. B, vol. 105, pp. 69–72; January, 1958.) A rapid way of determining the dispersion curve to the accuracy necessary when designing slow-wave structures for electron accelerators. See also 19 of 1956 (Grosjean).

621.372.83:621.372.029.6 1023
Recent Advances in Finline Circuits—S. D. Robertson. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 263–267; October, 1956. Abstract, PROC. IRE, vol. 45, p. 114; January, 1957.) See also 2555 of 1955.

621.372.832.6 1024
Recent Advances in Waveguide Hybrid Junctions—P. A. Loth. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 268–271; October, 1956. Abstract, PROC. IRE, vol. 45, p. 114; January, 1957.)

621.372.832.8 1025
A Broad-Band Microwave Circulator—E. A. Ohm. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 210–217; October, 1956. Abstract, PROC. IRE, vol. 45, p. 113; January, 1957.)

621.372.832.8 1026
The Turnstile Circulator—P. J. Allen. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 223–227; October, 1956.) Abstract, PROC. IRE, vol. 45, p. 113; January, 1957.

621.372.85:621.318.134 1027
Propagation in Ferrite-Filled Transversely Magnetized Waveguide—P. H. Vartanian and E. T. Jaynes. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 140–143; July, 1956. Abstract, PROC. IRE, vol. 44, p. 1898; December, 1956.)

621.372.8:52.363:621.318.134 1028
Improved Rectangular-Waveguide Resonance Isolators—M. T. Weiss. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 240–243; October, 1956. Abstract. PROC. IRE, vol. 45, p. 114; January, 1957.)

621.372.853:537.562 1029
Properties of Ion-Filled Waveguides—L. D. Smullin and P. Chorney. (PROC. IRE, vol. 46, pp. 369–361; January, 1958.) A mathematical analysis suggests that two pass bands exist at frequencies below the usual pass band for the empty waveguide.

621.396.67:537.226 1030
Some Investigations on Dielectric Aerials—R. Chatterjee and S. K. Chatterjee. (*J. Inst. Telecommun. Engs., India*, vol. 3, pp. 280–284; September, 1957.) A comparative study of the expressions obtained for the radiation pattern of a circular dielectric rod antenna, excited in the HE_{11} mode (see 3637 of 1956 and *J. Indian Inst. Sci.*, sect. B, vol. 39, pp. 134–140; July, 1957) using 1) Schelkunoff's equivalence principle, 2) the application of Huyghen's principle over the whole rod, and 3) the theory of Halliday and Kiely (3350 of 1948).

621.396.674.3:621.396.11 1031
Radiation resulting from an Impulsive Current in a Vertical Antenna Placed on a Dielectric Ground—C. L. Pekeris and Z. Alterman. (*J. Appl. Phys.*, vol. 28, pp. 1317–1323; November, 1957.) The electromagnetic field produced by passing a current with the form of a delta function $\delta(t)$ through a vertical dipole on a dielectric ground is calculated with the aid of an electronic computer, and the solutions discussed. This method of treatment has certain advantages over the assumption of a periodic current and also can be generalized to deal with an arbitrary current variation.

621.396.676.012.12 1032
Currents Excited on a Conducting Surface of Large Radius of Curvatures—J. R. Wait. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 143–145; July, 1956. Abstract, PROC. IRE, vol. 44, p. 1898; December, 1956.) See also 2296 of 1956.

621.396.677+534.23 1033
Some Aspects of the Design of Strip Arrays—Tucker. (See 994.)

621.396.677:621.372.826 1034

An Investigation of Periodic Rod Structures for Yagi Aerials—J. O. Spector. (*Proc. IEE*, vol. 105, pp. 38–44; January, 1958.) Experiments suggest that the radiation mechanism can be explained in terms of a surface wave propagating along the antenna together with a radiating aperture at the end. The side-lobe structure is explained by interference with direct radiation from the driven element caused by inefficient launching of the surface wave.

621.396.677.029.62:523.164 1035

A Radio Telescope—J. Firor. (*QST*, vol. 41, pp. 32–36; September, 1957.) A description of an interferometer system including constructional details of an antenna system suitable for tracking an earth satellite on a frequency of 108 mc.

621.396.677.029.62:523.164:629.19 1036

Mark II Minitrack Base-Line Components—R. L. Easton. (*QST*, vol. 41, pp. 37–41; September, 1957.) Full constructional details of an antenna system suitable for earth satellite tracking on 108 mc. See also *ibid.*, vol. 40, pp. 38–41, 134; July, 1956.

621.396.677.31:[523.164+523.5 1037

An Antenna Array for Studies in Meteor and Radio Astronomy at 13 Metres—P. B. Gallagher. (Proc. IRE, vol. 46, pp. 89–92; January, 1958.) Short description of a broadside array at Stanford, Calif., designed for radar studies of very small meteors. Preliminary results indicate that the system should be sensitive to meteors down to the 15th visual magnitude. The theoretical half-power azimuthal beam width is 1.2°. Other possible applications of the system are noted.

621.396.677.83:523.164 1038

Radio-Telescope Antennas of Large Aperture—J. D. Kraus. (Proc. IRE, vol. 46, pp. 92–97; January, 1958.) A discussion of the detection and resolving powers of radio telescopes illustrates the need for larger apertures, and a design of a low-cost antenna system which provides a large effective aperture at 300 mc is described. It consists of a combination of long parabolic and flat sheet reflectors built on the ground.

621.396.677.833.2.095 1039

The Radiation Diagram of the Paraboloid under Various Conditions of Illumination—G. Barzilai, C. Montebello, and F. Serracchioli. (*Note Recensioni Notiz.*, vol. 6, pp. 525–538; July/August, 1957.) Radiation diagrams obtained at 9375 mc with different types of radiator are reproduced and discussed.

621.396.677.833.2.095 1040

Microwave Antenna Characteristics in the Presence of an Intervening Ridge—R. Vikramsingh, M. N. Rao, and S. Uda (*J. Instn Telecommun. Engs, India*, vol. 3, pp. 274–279; September, 1957.) A description of parabolic-antenna characteristics at 1860 mc and 1940 mc as measured over a 14-km path. The polar diagrams, and the effects of polarization, defocusing, and varying angle of elevation are recorded.

AUTOMATIC COMPUTERS

681.142 1041

Logical Design of a Computer for Business Use—R. J. Froggat. (*J. Brit. IRE*, vol. 17, pp. 681–696; December, 1957.) Paper presented at the Convention on Electronics in Automation, Cambridge, England, June, 1957, describing an E.M.I. serial machine operating at a clock rate of 115 kc with magnetic-drum storage.

681.142 1042

Torsional Waves in Wires: Disc-Loading Increases Computer Memory—(*Engineering, London*, vol. 183, p. 787; June 21, 1957.) Note on the characteristics of a disk-loaded delay line machined from a solid brass rod.

681.142 1043

Digital/Analogue Converter Provides Storage—H. N. Putschi, J. A. Raper, and J. J. Suran. (*Electronics*, vol. 30, pp. 148–151; December 1, 1957.) A transistorized converter changes eight binary bits, received in parallel from a shift register, to 128 steps in amplitude of a 400-cps sine wave. One binary bit is used to obtain phase information. Operation is performed within a 4-ms sampling period occurring at an average rate of 20 cps.

681.142:621.039 1044

Analogue and Digital Computers in Nuclear Engineering—(*Nucleonics*, vol. 15, pp. 53–88; May, 1957.) A group of articles outlining the principles of operation and giving a survey of applications in the U.S.A.

681.142:621.314.7 1045

A.D.A.—a Transistor Decimal Digital Differential Analyser—N. W. Allen. (*J. Instn Engs. Aust.*, vol. 29, pp. 255–263; October/November, 1957.) Circuit techniques and assembly and construction methods are discussed. The specification of the machine is summarized and some typical problems are considered.

681.142:621.318.57:537.227 1046

Signals from Switched Ferroelectric Memory Capacitors—Pulvari and McDuffie. (See 1053.)

618.142:621.372.4/.5 1047

Electric Nonlinear Computation Performed by Linear Elements—Abow-Hussein. (See 1061.)

681.142:621.375.4 1048

Application of Junction Transistors to Carrier-Frequency Computing Amplifiers—W. A. Curtin. (*Commun. and Electronics*, no. 28, pp. 746–752; January, 1957.) A discussion of the properties required by a transistorized summing amplifier in a carrier-frequency analog computer. Details are given of a practical 400-cps amplifier using standard commercial transistors.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.049.75 1049

Moulding moves in on Printed Circuits—N. L. Greenman. (*Product Eng.*, vol. 28, pp. 90–93; September 30, 1957.) The moulding process described permits the use of heavy-gauge conductor material in circuits of high current-carrying capacity.

621.3.078:621.372.852.3 1050

An Automatic Gain Control System for Microwaves—J. P. Vinding. (IRE Trans. on Microwave Theory and Techniques, vol. 4, pp. 244–245; October, 1956. Abstract, Proc. IRE, vol. 45, p. 114; January, 1957.

621.314.2:621.372.51 1051

Wide-Band Balun Transformer—A. I. Talkin and J. V. Cuneo. (*Rev. Sci. Instr.*, vol. 28, pp. 808–815; October, 1957.) Description of a balun transformer developed for use with a cro deflection amplifier. Bandwidth extends from below 50 kc to above 500 mc. Input impedance data and pulse response characteristics are shown. One balun coil can be used alone as a passive wide-band pulse inverter.

621.314.2.029.55/.62 1052

Design of Wide-Band R. F. Transformers utilizing a Synthesized Equivalent Network—H. H. Kajihara. (*Commun. and Electronics*, no. 28, pp. 802–805; January, 1957.)

621.318.57:537.227:681.142 1053

Signals from Switched Ferroelectric Memory Capacitors—C. F. Pulvari and G. E. McDuffie, Jr. (*Commun. and Electronics*, no. 28, pp. 681–685; January, 1957.) The equivalent circuit for a bistable storage capacitor is a current generator with a resistance in parallel. The properties of the output signals from switched and unswitched stores are discussed using this model.

621.318.57:621.314.63 1054

High-Speed Microwave Switching of Semiconductors—R. V. Garver, E. G. Spencer, and R. C. LeCraw. (*J. Appl. Phys.*, vol. 28, pp. 1336–1338; November, 1957.) A microwave switch using an *n*-type Ge diode has been shown to give pulse rise and decay times as low as 3 mμs.

621.318.57:621.314.7 1055

Transistor NOR Circuit Design—W. D. Rowe and G. H. Royer. (*Commun. and Electronics*, no. 31, pp. 263–267; July, 1957.) A NOR circuit gives a signal output only when no input signal is present. Examination of transistor characteristics gives a stable design for an arbitrary number of inputs and outputs on a single Type-2N109 transistor. Crosstalk factor and speed of operation are considered.

621.318.57:621.314.7 1056

A New Family of Transistor Switching Circuits—M. Rubinoff. (*Commun. and Electronics*, no. 31, pp. 286–289; July, 1957.) Two pairs of output levels are obtained from the "dual range" circuits described which incorporate both *n-p-n* and *p-n-p* transistors. The possibility of lower standby power dissipation and greater switching speed than the DCTL circuits (see 2855 of 1955) is offered.

621.318.57:621.314.7 1057

Transistor 2-Terminal Switches—A. Har'el. (*Commun. and Electronics*, no. 31, pp. 328–338; July, 1957.) Discussion showing how the blocking-voltage concept may be applied to obtain negative-resistance characteristics using a junction transistor, a point-contact transistor, a modified double-base diode, or a modified tetrode. A few applications mainly in telephone systems are indicated.

621.372:621.396.822 1058

A Note on Noise Temperature—P. D. Strum. (IRE Trans. on Microwave Theory and Techniques, vol. 4, pp. 145–151; July, 1956. Abstract, Proc. IRE, vol. 44, p. 1898; December, 1956.) Expressions are given for the effective noise temperature of a lossy passive network, with internal noise sources. Their application in different receiving systems is discussed.

621.372.029.6:621.372.83 1059

Recent Advances in Finline Circuits—Robertson. (See 1023.)

621.372.2:512.831 1060

Equivalent Admittance, Impedance and Scattering Matrices—G. C. Corazza and F. Serracchioli. (*Note Recensioni Notiz.*, vol. 6, pp. 502–510; July/August, 1957.) (See also 694 of 1958.)

621.372.4/5:681.142 1061
Electric Nonlinear Computation Performed by Linear Elements—M. S. M. Abou-Hussein. (*Commun. and Electronics*, no. 32, pp. 378–380; September, 1957.) Simple linear circuits are shown for performing multiplication, division, and inversion, and forming powers, positive and negative roots, logarithms, antilogarithms, and hyperbolic functions.

621.372.5:621.3.015.3 1062
Interpretation of Network Theorems of Laplace Transforms—V. M. Narbutt. (*J. Instn Telecommun. Engs, India*, vol. 3, pp. 304–309; September, 1957.)

621.372.54:621.396.62 1063
Greater Selectivity with the C.W. Clipper Filter—L. I. Albert. (*QST*, vol. 41, pp. 24–26; September, 1957.) A two-stage amplifier with variable bandwidth.

621.372.54(083.57) 1064
Charts Simplify Passive *LC* Filter Design—D. R. J. White. (*Electronics*, vol. 30, pp. 160–163; December 1, 1957.) "Universal design data permit design of Butterworth and Tchebycheff filters of prescribed steady-state insertion-loss characteristics. Design of band-pass prototypes of lumped-element configuration is given."

621.372.54.029.6 1065
A Travelling-Wave Directional Filter—F. S. Coale. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 256–260; October, 1956. Abstract, PROC. IRE, vol. 45, p. 114; January, 1957.)

621.372.54.049 1066
Design and Manufacture of Practical Filter Circuits—S. Boyle. (*Electronics*, vol. 30, pp. 154–157; December 1, 1957.) Discusses the effects of distributed capacitance, component proximity, encapsulating material, and impedance mismatch.

621.372.56:621.385.029.6 1067
Travelling-Wave-Tube Limiters—Fank and Wade. (See 1288.)

621.372.6:621.3.016.35 1068
A Phasor Method of Nonlinear Network Analysis—J. P. Neal. (*Commun. and Electronics*, no. 28, pp. 630–636; January, 1957. Discussion, p. 636.) The network is represented by the sum of a finite number of simple sinusoidal functions with time, chosen from a knowledge of the network, experiment, and preliminary analysis. An Appendix illustrates stability tests and an application of the method to a typical problem.

621.373:517.942.932 1069
Fine Structure of Response Curves of Frequency—Entrained Oscillations—C. A. Ludeke and J. D. Blades. (*J. Appl. Phys.*, vol. 28, pp. 1326–1328; November, 1957.) Jump and hysteresis phenomena suggested by Cartwright (see 2740 of 1948), together with a dependence of entrainment range on forcing amplitude, have been observed in an electromechanical analog of a forced self-excited system.

621.373.029.4:621.396.822 1070
A Low-Frequency Random-Signal Generator—J. C. West and G. T. Roberts. (*J. Sci. Instr.*, vol. 34, pp. 447–450; November, 1957.) The 5-kc and 8-kc noise components, each in an over-all bandwidth of 250 cps, are selected from the noise output of an argon-filled thyratron operating in a magnetic field. After peak rectification the resulting noise signals are added so that their dc components cancel. The amplitude distribution of the resultant output then approximates very closely to a Gaussian distribution. The output noise spectrum is flat from zero frequency up to 100 cps. A full circuit diagram is given.

621.373.029.45:621.372.5 1071
Measurements of the Quality of Very-Low-Frequency RC Oscillators—A. Zanini. (*Note Recensioni Notiz.*, vol. 6, pp. 459–477; July/August, 1957.) A "figure of merit" is calculated for a number of different RC networks, and the experimental determination of this coefficient is described.

621.373.421.14.029.63 1072
Double Excitation in Decimetre-Wave Oscillators with Coaxial Lines—W. Rohde. (*NachrTech.*, vol. 7, p. 311; July, 1957.) The simultaneous oscillation at a frequency corresponding to a setting of about $\lambda/4$ in a disk-seal triode oscillator tuned to $3\lambda/4$ is investigated and the calculations are verified by experiment.

621.373.431.1:621.314.7:621.397.621 1073
Monovibrator has Fast Recovery Time—A. I. Aronson and C. F. Chong. (*Electronics*, vol. 30, pp. 158–159; December 1, 1957.) "Use of complementary transistors desreases recovery time of monostable multivibrator. This prevents erratic operation when circuit is used in a television sync generator. Since both transistors are off during timing cycle, circuit is relatively insensitive to transistor variations and operates reliably from -50 to $+70°C$ for input frequencies from 250 cps to 1 mc."

621.373.44 1074
A Simple Rectangular-Pulse Generator—H. L. Armstrong. (*Canad. J. Phys.*, vol. 35, pp. 1250–1252; October, 1957.) The cathode-coupled multivibrator circuit discussed produces pulses of up to 100-v amplitude with rise times from 0.1 to 1 μs.

621.373.44 1075
Method of Obtaining Integral Mark/Space Ratios of High Accuracy with a Continuously Variable Control Frequency—R. Rickert. (*NachrTech.*, vol. 7, pp. 303–304; July, 1957.) A pulse generator circuit is described in which the mark/space ratio can be adjusted to within 0.1 per cent over the range 1:1 to 1:9 for control frequencies from 10 cps to 1 kc.

621.373.5:621.3.018.41(083.74) 1076
Increasing Quartz Oscillator Stability—(*Engineering, London*, vol. 184, p. 496; October 18, 1957.) Brief description of a 5-mc quartz servo oscillator forming part of frequency-standard equipment Type RD101. A frequency error of less than 1 in 10^{10} occurs for a 5 per cent change in supply voltage or 5°C change in temperature.

621.374.32 1077
Fast Gray Wedge Analyser for High Input Rates—J. T. Flynn and F. A. Johnson. (*Rev. Sci. Instr.*, vol. 28, pp. 867–874; November, 1957.) A full description is given of the circuits and operation of a pulse height analyzer designed for positive input pulses at rates up to 10^7/s with less than 10 per cent distortion of the spectrum. The resolution is approximately 10 mμs.

621.374.5:621.396.962.3 1078
Pulse Compression: Part 2—R. Krönert. (*NachrTech.*, vol. 7, pp. 305–308; July, 1957.) Description of an experimental realization of the method discussed in 72 of 1958.

621.375.1.011.6 1079
The Time Behaviour of Logarithmic Amplifier Input Circuits—T. P. Flanagan. (*J. Sci. Instr.*, vol. 34, pp. 450–452; November, 1957.) "The nonlinear nature of the logarithmic amplifier input circuit causes the time constant to vary with current, and three definitions of effective time constant are examined, based on an analysis of the time behavior of the input circuit to a step change of input current."

621.375.122 1080
New Types of D.C. Amplifier: Part 1—The Cascade-Balance System—D. J. R. Martin. (*Electronic Radio Eng.*, vol. 35, pp. 2–7; January, 1958.) The Owen-Prinz method of zero drift correction [see, *e.g.*, 1602 of 1948 (Prinz)] is applied to the first of two identical amplifier stages; the residual drift is then balanced against that of the second stage. The over-all drift is reduced 100 times as compared to the parallel balance system. A practical design is fully discussed.

621.375.122 1081
New Types of D.C. Amplifier: Part 2—The Reflex-Monitor System—D. J. R. Martin. (*Electronic Radio Eng.*, vol. 35, pp. 56–62; February, 1958.) A correcting amplifier or monitor, identical to the input stage of the main amplifier, corrects alternately its own drift and that of the main amplifier. A differential input is used which rejects in-phase drift components, and residual drifts are balanced between the two amplifiers. Full details are given of a practical circuit which has certain advantages over the simpler cascade-balance system described in Part 1 (1080 above).

621.375.127 1082
Push-Pull Amplifier Pesign—R. G. Christian. (*Electronic Radio Eng.*, vol. 35, pp. 72–73; February, 1958.) "A method of designing class AB push-pull amplifiers which eliminates the need to plot composite characteristics is discussed, and some examples are given which compare favorably with practical results."

621.375.132 1083
Total Differential Feedback—J. C. H. Davis. (*Electronic Radio Eng.*, vol. 35, pp. 40–44; February, 1958.) A device for squaring the effective possible feedback in a feedback amplifier.

621.375.3 1084
The Simple Reactor Circuit; its Operation and Mode Transition—J. F. Ringelman and A. L. Fenaroli. (*Commun. and Electronics*, no. 28, pp. 660–668; January, 1957. Discussion, pp. 668–669.)

621.375.3 1085
The Series Magnetic Amplifier—R. C. Barker. (*Commun. and Electronics*, no. 28, pp. 819–830; January, 1957. Discussion, pp. 830–831.) A theoretical analysis is given of the steady-state operation of the amplifier for saturation states of the cores. The control-circuit-impedance and control-voltage relations determining the mode of operation are established, and the gain characteristic and time constant are considered.

621.375.3 1086
Analysis of the Full-Wave Magnetic-Amplifier Circuits considering the Change of the Width of the Dynamic Hysteresis Loop—T. Kikuchi. (*Commun. and Electronics*, no. 31, pp. 241–249; July, 1957.)

621.375.3 1087
The Operation of the Current-Type Self-Balancing Magnetic Amplifier—A. D. Krall. (*Commun. and Electronics*, no. 32, pp. 380–384; September, 1957.) A theoretical analysis giving the voltage gain and time constant in terms of

the circuit parameters. Within the range of ideal components the input impedance is zero.

621.375.4 1088
High-Frequency Amplification using Junction Transistors—L. E. Jansson. (*Mullard Tech. Commun.*, vol. 3, pp. 174–187; October, 1957.) The maximum possible stage gain of a unilateralized grounded-emitter narrow-band amplifier stage is calculated and the design of interstage coupling transformers is described. A typical 470-kc IF amplifier stage is designed and its stability factor calculated.

621.375.4 1089
An Analysis of Transient Response of Junction-Transistor Amplifiers—J. C. Bhattacharyya. (*J. Instn. Telecommun. Eng., India*, vol. 3, pp. 297–303; September, 1957.) An exact solution of the one-dimensional diffusion equation is obtained by the method of Laplace's transform. The short-circuited output collector current is calculated for a step-input forcing function. Experimental results agree with the theory, and for ordinary conditions the physical process underlying transistor action must be diffusion of minority carriers across the base region.

621.375.4+621.376.233]:539.169 1090
How Transistors Operate under Atomic Radiation—R. L. Riddle. (*Electronics*, vol. 30. pp. 125–127; December 1, 1957.) Results of tests on an amplifier/detector system show that degrading effects of irradiation can be controlled to some degree by negative feedback. Measurements were also made on a single separate transistor and a coaxial cable.

621.375.4:621.396.822 1091
Internal Noise of Transistor Amplifiers—J. J. Brophy and A. R. Reinberg. (*Rev. Sci. Instr.*, vol. 28, pp. 965–966; November, 1957.) The internal noise levels of two commercial transistor amplifiers were measured for various input impedances and the results are shown graphically.

621.375.4.029.3 1092
Transistors in Speech Equipment—H. J. Albrecht. (*QST*, vol. 41, pp. 19–22; September, 1957.) The use of transistors in af circuits is described and constructional details are given of a speech amplifier.

621.375.9:538.221:538.569.4.029.63/.64 1093
Theory of the Ferromagnetic Microwave Amplifier—H. Suhl. (*J. Appl. Phys.*, vol. 28, pp. 1225–1236; November, 1957.) "All three possible types of operation, using respectively, two electromagnetic cavity modes, two sample modes, and one sample and one cavity mode, are discussed. One especially simple case, that of a sphere in the first type of operation, is treated separately. Thereafter all three cases are discussed in terms of scalar and vector potentials. An Appendix deals with the gain-bandwidth problem and gives an expression for the equivalent 'negative Q' of the sample."

621.375.9:538.569.4.029.6 1094
Twin Cavity for NH_3 Masers—J. Bonanomi, J. Herrmann, J. De Prins, and P. Kartaschoff. (*Rev. Sci. Instr.*, vol. 28, pp. 879–881; November, 1957.) "A system of two coupled cavities is described replacing the single cavity of an NH_3 maser. Using this system the curve of the oscillation frequency against cavity temperature presents a plateau, thus reducing considerably the 'pulling' effects of the cavity."

621.375.9:538.569.4.029.6:621.396.822 1095
Maser Noise Considerations—J. Weber. (*Phys. Rev.*, vol. 108, pp. 537–541; November 1, 1957.) A discussion of the effect of the saturation rf field on the noise figure of a three-level maser. For the conditions considered, the effect is small. The spontaneous-emission equivalent temperature for a free-electron vacuum-tube amplifier is one-half that of a maser. In principle, more general quantum-mechanical amplifiers can be constructed which will not have spontaneous-emission noise.

621.376.32:621.318.134 1096
Compact Microwave Single-Sideband Modulator using Ferrites—J. C. Cacheris and H. A. Dropkin. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 152–155; July, 1956. Abstract, PROC. IRE, vol. 44, pp. 1898–1899; December, 1956.) See also 3183 of 1954 (Cacheris.)

621.376.332 1097
Foster-Seeley Discriminator—C. G. Mayo and J. W. Head. (*Electronic Radio Eng.*, vol. 35, pp. 44–51; February, 1958.) A detailed analysis is given of the circuit and the use of the design formulas is illustrated. With the circuit arrangement described a sensitivity of 1.4 v (af signal) per ma (IF) can be obtained with less than 0.1 per cent second-harmonic distortion at 100 per cent modulation.

GENERAL PHYSICS

531.19–2 1098
Statistical Properties of an Isotropic Random Surface—M. S. Longuet-Higgins. (*Phil. Trans. A*. vol. 250, pp. 157–174; October 17, 1957.)

535.37 1099
Possibility of Luminescent Quantum Yields Greater than Unity—D. L. Dexter. (*Phys. Rev.*, vol. 108, pp. 630–633; November 1, 1957.) "It is shown than an excited sensitizer can transfer its energy simultaneously to two activators, under suitable conditions, leading to two emitted photons per incident higher energy photon. The probability of this transfer process is computed, and the process is shown to be experimentally feasible."

537+538]:621.38 1100
Physics and the New Electronics—R. A. Smith. (*J. Sci. Instr.*, vol. 34, pp. 377–382; October, 1957.) Vacuum electronics and the new crystalline-solid electronics are compared. A brief account of recent work on semiconductors and maser amplifiers is given and the effect of these new developments on the training of electronic engineers is discussed.

537.122 1101
Some Remarks about Electron Correlation—O. Krisement. (*Phil. Mag.*, vol. 2, pp. 245–258; February, 1957.) A comparison is made between theories representing different approaches to the inclusion of correlation in the wave-mechanics treatment of a free electron gas.

537.122 1102
Correlation Energy of an Electron Gas at High Density: Plasma Oscillations—K. Sawada, K. A. Brueckner, N. Fukuda, and R. Brout. (*Phys. Rev.*, vol. 108, pp. 507–514; November 1, 1957.) An analysis of the contribution made by zero-point oscillations to the correlation energy of an electron gas at high density. This contribution is expressed in terms of that from the scattering states by using the analytic properties of the scattering amplitudes. The results are compared with those of Bohm and Pines (1375 of 1954).

537.122 1103
Correlation Energy of a High-Density Gas: Plasma Coordinates—R. Brout. (*Phys. Rev.*, vol. 108, pp. 515–517; November 1, 1957.) "The model Hamiltonian of Sawaba [*ibid.*, vol. 106, pp. 372–383; April 15, 1957] which describes electron correlation at high density is examined. It is shown that the set of scattering modes for momentum transfers below a certain q_{max} is not complete. It is completed by the plasma mode. $(q_{max})^{-1}$ is the natural Debye length of the theory."

537.122 1104
Characteristic Energy Loss of Electrons passing through Metal Foils: Momentum-Exciton Model of Plasma Oscillations—R. A. Ferrell and J. J. Quinn. (*Phys. Rev.*, vol. 108, pp. 570–575; November 1, 1957.) Previous work is extended to obtain a unified theory of collective and individual electron effects in the excitation of a degenerate electron gas by fast incident electrons. The theory is equivalent to the screening out by the conduction electrons of the external field set up by the incident electrons.

537.311.1 1105
Quantum Theory of Electrical Transport Phenomena—W. Kohn and J. M. Luttinger. (*Phys. Rev.*, vol. 108, pp. 590–611; November 1, 1957.) Using a simple model a technique is developed which gives the entire density matrix of the system of charge carriers in the steady state. The model consists of noninteracting free (or Bloch) electrons scattered by "random" rigid impurity centers. The density matrix is developed in ascending powers of the strength of the scattering potential. The Boltzmann transport equation represents an approximation valid in the limiting cases of very weak or very dilute scatterers. Higher-order corrections are given.

537.311.31:536.63 1106
The Specific Heat of the Electrons of Metals—F. Kaschluhn. (*Ann. Phys., Leipzig*, vol. 19, pp. 94–101; December 20, 1956.) A new approximation method is used to calculate the specific heat for low temperatures taking account of electron interaction. Results are compared with those obtained by other methods.

537.525 1107
Exact Nonlinear Plasma Oscillations—I. B. Bernstein, J. M. Greene, and M. D. Kruskal. (*Phys. Rev.*, vol. 108, pp. 546–550; November 1, 1957.) A solution is given for the problem of a one-dimensional stationary nonlinear electrostatic wave in a plasma free from interparticle collisions. In the limiting case of small-amplitude waves, the linearized theory can still be applied by using singular first-order distribution functions.

537.525 1108
Contribution to the Diffusion Oscillations in Gas-Discharge Plasma—O. V. Prudkovskaya. (*Dokl. Ak. Nauk S.S.S.R.*, vol. 117, pp. 601–604; December 1, 1957.)

537.525.8:538.569 1109
Quenching of the Negative Glow by Microwaves in Cold-Cathode Gaseous Discharges—J. M. Anderson. (*Phys. Rev.*, vol. 108, pp. 898–899; November 1, 1957.) A discussion of results obtained in He at a pressure of 10.2 mm Hg using a 40 μs pulse at 9375 mc. Enhanced emission was observed in Ne, A, and Xe discharges at 1–10 mm Hg.

537.533.8 1110
The Theory of the Surface Effect of Secondary Emission—W. Brauer and W. Klose.

(*Ann. Phys., Leipzig*, vol. 19, pp. 116–132; December 20, 1956.) For primary energies of about 100 ev about 30 per cent of the secondary-electron yield in light metals can be ascribed to surface effects. For higher energies and heavier metals surface effects can be neglected.

537.56 1111
Effective Electron-Electron and Electron-Ion Cross-Sections in Plasmas—M. Bayet. (*J. Phys. Radium*, vol. 18, pp. 380–386; June, 1957.) Cross sections for elastic scattering are infinite for strictly Coulomb interaction, but finite values are found by using Debye's theory of electrolytes.

538:061.3 1112
Report on the 4th Course of the International School of Physics of the Italian Physical Society, Varenna, 15th July–4th August 1956—(*Nuovo Cim.*, vol. 6, suppl. no. 3, pp. 801–1237; 1957.) Report of the proceedings of the course on magnetic properties held at the Villa Monastero, Varenna. The text is given of lectures and communications, including the following:

1) **Paramagnetism in Crystals**—M. H. L. Pryce (pp. 817–856, in English).
2) **Magnetic Properties of Metals**—J. H. Van Vleck (pp. 857–886, in English).
3) **Paramagnetic Relaxation**—C. J. Gorter (pp. 887–894, in English).
4) **Ferromagnetism**—C. Kittel (pp. 895–922, in English).
5) **Antiferromagnetism**—C. J. Gorter (pp. 923–941, in English).
6) **Metamagnetics or Critical-Field Antiferromagnetic Materials**—L. Néel (pp. 942–960, in French).
7) **Nuclear Magnetism and Nuclear Relaxation**—E. M. Purcell (pp. 961–992, in English).
8) **Line Breadths and the Theory of Magnetism**—J. H. Van Vleck (pp. 993–1014, in English).
9) **The Influence of Electrons on Nuclear Spin Resonance in Diamagnetic Materials: "Chemical" Displacement and Indirect Interaction**—A. Abragam (pp. 1015–1062; in French).
10) **Stochastic Theory of Magnetic Resonance**—R. Kubo (pp. 1063–1080, in English).
11) **The Concept of Temperature in Magnetism**—J. H. Van Vleck (pp. 1081–1100, in English).
12) **Magnetism at Very Low Temperatures and Nuclear Orientation**—N. Kurti (pp. 1101–1139, in English).
13) **Cyclotron Resonance in Crystals**—C. Kittel (pp. 1140–1147, in English).
14) **Optical Methods for Radio-Frequency Resonance**—A. Kastler (pp. 1148–1167, in French).
15) **Magnetic Properties of Superconductors**—C. J. Gorter (pp. 1168–1176, in English).
16) **Hall Effect in Ferromagnetics**—J. Smit (pp. 1177–1182; in English).
17) **Critical Scattering of Neutrons from Ferromagnets**—W. Marshall (pp. 1183–1184, in English. Discussion, pp. 1184–1185).
18) **The Van Vleck Model of Ferromagnetism**—W. Marshall (pp. 1186–1187, in English).
19) **Influence of the Apparatus on Nuclear Magnetic Resonance**—H. Pfeifer (pp. 1188–1189, in English).
20) **Paramagnetic Resonance of Impurities in a Semiconductor**—A. Abragam and J. Combrisson (pp. 1197–1211, in French. Discussion, p. 1212). See also 165 of 1957.
21) **The Electric Analogue to Antiferromagnetism: Antiferroelectricity**—H. Gränicher (pp. 1220–1222; in English. Discussion, p. 1223).
22) **Magnetic Susceptibility of Electrons in Periodic Fields**—C. P. Enz (pp. 1224–1229, in English).

538.3 1113
Equivalence Theorems for Electromagnetic Fields—G. C. Corazza. (*Ricerca sci.*, vol. 27, pp. 2109–2119; July, 1957.) Theorems are considered with reference to electric and magnetic surface currents.

538.3 1114
Forces and Stresses in an Electromagnetic Field—T. H. Lee. (*Commun. and Electronics*, no. 31, pp. 267–271; July, 1957. Discussion, pp. 271–274.) The theories of Faraday, Larmor and Livens, and the "energy method" are compared and applied to some illustrative examples. The three theories may lead to different results for nonrigid magnetic materials, and the Larmor and Livens theory may not be applicable to magnetostriction. A discussion by R. M. Lichtenstein giving a further "thermodynamic" method is included.

538.311 1115
Coils for the Production of High-Intensity Pulsed Magnetic Fields—S. Foner and H. H. Kolm. (*Rev. Sci. Instr.*, vol. 28, pp. 799–807; October, 1957.) The design, performance, and applications of a pulsed-field system for 750,000 G, are described. It consists of a 2000 μF capacitor bank charged to 3 kv which is discharged through a Be-Cu helix of 3/16 inch internal diameter and $\frac{1}{2}$ inch long. Detailed design data are also given for a range of coils giving increased volume and field uniformity at lower field intensities and include a coil providing transverse access to the field.

538.311 1116
Production and Use of High Transient Magnetic Fields: Part 2—H. P. Furth, M. A. Levine, and R. W. Waniek. (*Rev. Sci. Instr.*, vol. 28, pp. 949–958; November, 1957.) The mechanical and thermal limitations of solid metal coils are discussed and illustrated by experiment. These limitations can be removed by force-free coil designs. Part 1: 3030 of 1956 (Furth and Waniek).

538.311 1117
Variable and Reversible Magnetic Fields Obtained from a Permanent Magnet—J. A. Dalman and L. S. Goodman. (*Rev. Sci. Instr.*, vol. 28, pp. 961–962; November, 1957.) A permanent magnet forms the core of a cylinder whose surface is made of two armco iron shoes and two brass shoes. Rotation of the cylinder directs the flux either to a shunt or across the gap of the magnet, for deflection of an atomic beam.

538.566:537.122 1118
Oscillations of Electron Cloud in External Fields—M. Sumi. (*J. Phys. Soc. Japan*, vol. 12, pp. 1110–1117; October, 1957.) The density fluctuations are analyzed in a pure electron gas not neutralized by positive ions and under the influence of an external magnetic field.

538.566:621.396.677.85 1119
A Semi-infinite Array of Parallel Metallic Plates of Finite Thickness for Microwave Systems—R. I. Primich. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 156–166; July, 1956. Abstract, PROC. IRE, vol. 44, p. 1899; December, 1956.)

538.569.4:538.221 1120
On the Line Width in Ferromagnetic Resonance—S. Takeno. (*Progr. Theor. Phys.*, vol. 18, pp. 448–449; October, 1957.) It is suggested that the line width is modified by a nonuniformity of the demagnetization field throughout the specimen. It is shown that the broadening is greater for a disk magnetized normal to its plane than for one magnetized in its plane, and also that the effect would be greater for ferrites than metals; both results are in agreement with experiment.

538.569.4.029.63/.64:538.221:621.375.9 1121
Theory of the Ferromagnetic Microwave Amplifier—Suhl. (See 1093.)

538.569.4.029.65:535.34 1122
High-Temperature Molecular-Beam Microwave Spectrometer—A. K. Garrison and W. Gordy. (*Phys. Rev.*, vol. 108, pp. 899–900; November 1, 1957.) A description of measurements at 3 mm λ on the $J=12\rightarrow13$ rotational transition of KCl.

538.652 1123
Effective Magnetic Anisotropy and Magnetostriction of Monocrystals—P. K. Baltzer. (*Phys. Rev.*, vol. 108, pp. 580–587; November 1, 1957.) A nine-constant expression is derived which describes empirically the spontaneous magnetostriction to the sixth order in the direction cosines of the magnetization. The relation between the effective anisotropy in domains and domain walls is also studied.

538.691 1124
Diffusion of Charged Particles in a Homogeneous Electromagnetic Field—Kh. Ya. Khristov. (*Dokl. Ak. Nauk, S.S.S.R.*, vol. 116, pp. 213–216; September 11, 1957.) An approximate solution is considered.

539.11:548.4 1125
The Elastic Impurity-Centre Model—J. Teltow. (*Ann. Phys., Leipzig*, vol. 19, pp. 169–174; December 20, 1956.) An approximation is derived for the purely elastic interaction between two extraneous atoms in a crystal lattice.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164 1126
Radio Astronomy—(PROC. IRE, vol. 46, January, 1958.) The main part of this issue is devoted to a group of 50 papers dealing with practical and theoretical aspects of radio astronomy. Abstracts of some of the papers are given individually; titles of others are as follows.

Introduction to Radio Astronomy—F. T. Haddock (pp. 3–12).

The Discovery and Identification by Karl Guthe Jansky of Electromagnetic Radiation of Extraterrestrial Origin in the Radio Spectrum—C. M. Jansky, Jr, (pp. 13–15).

Early Radio Astronomy at Wheaton, Illinois—G. Reber (pp. 15–23).

The Telescope Program for the National Radio Astronomy Observatory at Green Bank, West Virginia—R. M. Emberson and N. L. Ashton (pp. 23–35).

Noise Levels at the National Radio Astronomy Observatory—J. W. Findlay (pp. 35–38).

Radio Astronomy at the Meudon Observatory—E. J. Blum, J. F. Denisse, and J. L. Steinberg (pp. 39–43).

A High-Resolution Radio Telescope for Use at 3.5 M—B. Y. Mills, A. G. Little, K. V. Sheridan, and O. B. Slee (pp. 67–84).

The Sydney 19.7-MC Radio Telescope—C. A. Shain (pp. 85–88).

Measurements of Solar Radiation and At-

mospheric Attenuation at 4.3 Millimetres Wavelength—R. J. Coates (pp. 122–126).

Scanning the Sun with a Highly Directional Array—W. N. Christiansen and D. S. Mathewson (pp. 127–131).

A Wide-Band Antenna System for Solar Noise Studies—H. Jasik (pp. 135–142).

The Radio Spectrum of Solar Activity—A. Maxwell, G. Swarup, and A. R. Thompson (pp. 142–148).

A Swept-Frequency Interferometer for the Study of High-Intensity Solar Radiation at Meter Wavelengths—J. P. Wild and K. V. Sheridan (pp. 160–171).

The Cornell Radio Polarimeter—M. H. Cohen (pp. 183–190).

A Time-Sharing Polarimeter at 200 MC—S. Suzuki and A. Tsuchiya (pp. 190–194).

A Polarimeter in the Microwave Region—K. Akabane (pp. 194–197).

Radio Sources and the Milky Way at 440 MC—N. G. Roman and B. S. Yaplee (pp. 199–204).

Flux Measurements of Cassiopeia A and Cygnus A between 18.5 MC and 107 MC—H. W. Wells (pp. 205–208).

The Distribution of Cosmic Radiation Background Radiation—H. C. Ko (pp. 208–215).

A Galactic Model for Production of Cosmic Rays and Radio Noise—L. Marshall (pp. 215–220).

Hydrogen-Line Study of Stellar Associations and Clusters—T. K. Menon (pp. 230–234).

Extragalactic 21-CM-Line Studies—D. S. Heeschen and N. H. Dieter (pp. 234–239).

Measurements of Planetary Radiation at Centimeter Wavelengths—C. H. Mayer, T. P. McCullough, and R. M. Sloanaker (pp. 260–266).

Planetary and Solar Radio Emission at 11 Meters Wavelength—J. D. Kraus (pp. 266–274).

Radio Emission from Comet 1956 h on 600 MC—R. Coutrez, J. Hunaerts, and A. Koeckelenbergh (pp. 274–279).

Lunar Thermal Radiation at 35 KMC—J. E. Gibson (pp. 280–286).

Lunar Radio Echoes—J. H. Trexler (pp. 286–292).

Radar Echoes from the Moon at a Wavelength of 10 CM—B. S. Yaplee, R. H. Bruton, K. J. Craig, and N. G. Roman (pp. 293–297).

A Phase Tracking Interferometer—H. Penfield (pp. 321–325).

Radio Astronomy Measurements at VHF and Microwaves—J. Aarons, W. R. Barron, and J. P. Castelli (pp. 325–333).

Cosmical Electrodynamics—J. H. Piddington (pp. 349–355).

523.164 1127

Spectral Lines in Radio Astronomy—A. H. Barrett. (Proc. IRE, vol. 46, pp. 250–259; January, 1958.) A review of atomic and molecular resonance lines occurring in the radio spectrum. It is shown how radiation from the galaxy is modified by interstellar gas.

523.164 1128

Radio Astronomy Polarization Measurements—M. H. Cohen. (Proc. IRE, vol. 46, pp. 172–183; January, 1958.) A survey of possible techniques for measuring polarization, and for studying the Faraday effect, shows which components should be measured. Methods for deducing the electron density along the path are discussed and applied to data on radiation from the sun and the Crab Nebula.

523.164 1129

Absorption Techniques as a Tool for 21-CM Research—A. E. Lilley and E. F. McClain. (Proc. IRE, vol. 46, pp. 221–229; January, 1958.) Absorption lines in the spectra of radio stars are used to study the distribution of interstellar gas, minimum distances to radio stars, and problems related to cosmology.

523.164 1130

Excitation of the Hydrogen 21-CM Line—G. B. Field. (Proc. IRE, vol. 46, pp. 240–250; January, 1958.) The influence of certain mechanisms on the spin temperature, such as electron collision and interaction with light, are considered, and their importance to radio observations is discussed. The deuterium line at $\lambda=91.6$ cm is also considered.

523.164:53.088 1131

Restoration in the Presence of Errors—R. N. Bracewell. (Proc. IRE, vol. 46, pp. 106–111; January, 1958.) A discussion of the relationship between observed antenna temperatures and the true distribution of a celestial source. Techniques for allowing for certain errors are described.

523.164:621.317.794:621.396.822 1132

A Broad-Band Microwave Source Comparison Radiometer for Advanced Research in Radio Astronomy—Drake. (See 1229.)

523.164:621.317.794.089:621.396.822 1133

A Method of Calibrating Centimetric Radiometers using a Standard Noise Source—Hey and Hughes. (See 1230.)

523.164:621.396.677.029.62 1134

A Radio Telescope—Firor. (See 1035.)

523.164+523.5]:621.396.677.31 1135

An Antenna Array for Studies in Meteor and Radio Astronomy at 13 Metres—Gallagher. (See 1037.)

523.164:621.396.677.83 1136

Radio-Telescope Antennas of Large Aperture—Kraus. (See 1038.)

523.164:629.19:621.396.677.029.62 1137

Mark II Minitrack Base-Line Components—Easton. (See 1036.)

523.164.2 1138

The Large-Scale Structure of the Galaxy—F. J. Kerr, J. V. Hindman, and M. S. Carpenter. (*Nature, London*, vol. 180, pp. 677–679; October 5, 1957.) Observations at 21 cm λ are being made at Sydney, Australia, using an antenna of 36 feet diameter and beam width 1.5° and a receiver with bandwidth 40 kc. A composite diagram of the galaxy is shown based on these observations and similar observations at Leyden, The Netherlands.

523.164.3:523.6 1139

Radio Observations of the Comet Arend-Roland—G. R. Whitfield and J. Högbom. (*Nature, London*, vol. 180, p. 602; September 21, 1957.) Observations made at Cambridge between March 12 and May 13, 1957 at 38 and 81.5 mc using 5 different instruments gave negative results. Emission at 600 and 27.6 mc has previously been reported elsewhere.

523.164.32 1140

Duration of Transients in Solar Radio Noise—Ö. Elgaröy. (*Nature, London*, vol. 180, pp. 808–809; October 19, 1957.) Histograms of observations made at 200 mc at Oslo are given. Reber's theory (see 1637 of 1955) is confirmed for this frequency. See also 439 of 1958 (de Groot). (Please note change of U.D.C. number.)

523.164.32 1141

Frequency Drift of Short-Time Transients in Solar Radio Noise—Ö. Elgaröy. (*Nature, London*, vol. 180, p. 862; October 26, 1957.) An analysis of records made at Oslo on a two-channel receiver is being made to give a detailed study of solar-noise storm bursts. Channels are centered on 199 and 200.5 mc, bandwidth is 0.3 mc, time constant 0.01 s and noise factor about 4. It is suggested that the transients in most cases have a frequency drift in either direction ranging from about 10 mc^2 to higher values.

523.164.32:53.088 1142

Discussion of 10.7-CM Solar Radio Flux Measurements and an Estimation of the Accuracy of Observations—W. J. Medd and A. E. Covington. (Proc. IRE, vol. 46, pp. 112–118; January, 1958.) Revision of previously published data is recommended.

523.164.32:538.566.2(083.57) 1143

Critical Frequency, Refractive Index, and Cone of Escape in the Solar Corona—R. N. Bracewell and C. V. Stableford. (Proc. IRE, vol. 46, pp. 198–199; January, 1958.) "Nomograms give refractive index governing radio propagation in the solar corona and the semi-vertical angle of the cone of escape."

523.164.32:523.75 1144

Studies at the McMath-Hulbert Observatory of Radio-Frequency Radiation at the Time of Solar Flares—H. W. Dodson. (Proc. IRE, vol. 46, pp. 149–159; January, 1958.) Records of radiation at 80, 200, and 2800 mc during several hundred flares are compared with associated data obtained from spectroheliograms.

523.164.32:621.317.75 1145

A Dynamic Spectrum Analyzer for Solar Studies—J. Goodman and M. Lebenbaum. (Proc. IRE, vol. 46, pp. 132–135; January, 1958.) The range 90–580 mc is swept three times a second in three overlapping bands. Results appear as a frequency-intensity display on film.

523.164.42:535.417 1146

Radio Interferometry of Discrete Sources—R. N. Bracewell. (Proc. IRE, vol. 46, pp. 97–105; January, 1958.) "Salient features of the theory and practice of radio interferometry are presented with special attention to assumptions and to the specifically two-dimensional aspects of the subject. A theorem is proved according to which only certain discrete stations on a rectangular lattice need be occupied for full determination of a discrete source distribution. Procedures in interferometry are discussed in the light of this result and an optimum procedure is deduced. Current practice is considered over conservative; *e.g.*, independent data in the case of the sun are obtainable only at station spacings of about 100 wavelengths on the ground, a fact which has not hitherto been taken into account."

550.38 1147

Preliminary Report of Geomagnetic Observations at Prince Harald Coast, Antarctica—T. Nagata, T. Oguti, and K. Momose. (*Rep. Ionosphere Res. Japan*, vol. 11, pp. 41–49; June, 1957.) The results of geomagnetic observations during the first Japanese Antarctic Research Expedition (1956/1957) are briefly summarized. The geomagnetic total intensity was found to be 10 per cent less than the value in Vestine's world map. Correlations between geomagnetic and ionospheric phenomena are discussed.

550.389.2:629.19 1148

Apsidal Motion of an I.G.Y. Satellite Orbit—L. Blitzer. (*J. Appl. Phys.*, vol. 28, p. 1362;

November, 1957.) An error in previous work [756 of 1957 (Blitzer *et al.*) and *ibid.*, vol. 28, p. 279; February, 1957 (Blitzer and Wheelon)] is corrected, concerning the movement of the line of apsides relative to a mode. (Please note change of U.D.C. number.)

551.510.53:[551.557+551.524.7 1149

Winds and Temperatures between 20 km and 100 km—a Review—R. J. Murgatroyd. (*Quart. J. R. Met. Soc.*, vol. 83, pp. 417–458; October, 1957.) The results of measurements, by various methods, of temperature and wind velocity in the northern hemisphere are briefly outlined. The results are used to obtain consistent systems of zonal wind and temperature for the winter and summer seasons. A standard atmosphere is suggested for northern latitudes to a height of 60 km, extended to 100 km for latitudes 30°N to 60°N, for both seasons.

551.510.535:523.164 1150

The Use of Radio Stars to Study Irregular Refraction of Radio Waves in the Ionosphere—H. G. Booker. (PROC. IRE, vol. 46, pp. 298–314; January, 1958.) Information from various sources on the amplitude and phase scintillation of radio-star radiation is reviewed, and the experimental data compared with theory. There is general agreement on most, but not all, of the characteristics of the scintillation and its variation with time and zenith angle. The rate of scintillation increases under magnetically disturbed conditions. There is good correlation with spread F ionospheric reflections and some with sporadic E reflections. The scale of the irregularities causing the scintillation is of the order of a kilometer. The most puzzling problem is the cause of the night-time maximum in scintillation. Forty-one references.

551.510.535:523.164 1151

An Investigation of the Perturbations Imposed upon Radio Waves Penetrating the Ionosphere—R. S. Lawrence. (PROC. IRE, vol. 46, pp. 315–320; January, 1958.) A new method is described for measuring continuously the phase deviations introduced by the ionosphere into the radiation from discrete sources. Records at 53, 108, and 470 mc are shown. In future work vertical incidence measurements will be made of the ionopshere at the point where the extraterrestrial waves penetrate.

551.510.535:523.164 1152

Some Measurements of High-Latitude Ionospheric Absorption Using Extraterrestrial Radio Waves—C. G. Little and H. Leinbach. (PROC. IRE, vol. 46, pp. 334–348; January, 1958.) A theory of absorption and a technique of measurement are described. Observations at 30 mc have shown that regions of anomalous absorption have dimensions in excess of 100 km and that marked differences can occur, in disturbed periods, between station 800 km apart. Most of the absorption occurs below the E region and it correlates well with the geomagnetic *K* index.

551.510.535 "1957":621.396.11 1153

Ionosphere Review 1957—T. W. Bennington. (*Wireless World*, vol. 64, pp. 77–78; February, 1958.) Sunspot activity has reached unprecedented high values; average F_2-layer critical frequencies have increased by a factor of just over two since 1954; monthly mean values for noon and midnight were considerably higher than at last sunspot maximum. MUF's as high as 50 mc have been reported, and examples of long-distance reception at 41.5 and 45 mc are noted. Though sunspot maximum may have occurred before the end of 1957 little change in propagation conditions is expected over various paths in 1958 as compared with those during similar months of 1957. "Quasi-maximum" conditions may well prevail throughout 1959.

LOCATION AND AIDS TO NAVIGATION

621.396.932.2 1154

"S.A.R.A.H.": a U.H.F. (243-Mc/s) Pulse Coded Air/Sea Rescue System—D. Kerr. (*J. Brit. IRE*, vol. 17, pp. 669–680; December, 1957.) The portable beacon emits 160–240 pairs of 5–10-μs pulses per sec and the pulse spacing can be varied from 100–300 μs. The double pulse permits discrimination between several beacons, and facilitates synchronization at the receiving station. Speech may be transmitted by pulse frequency modulation. The search equipment can detect pulses at a range of 145 km at a height of 3000 m, or 9 km at sea level.

621.396.933 1155

Some Radio Aids for High-Speed Aircraft—J. S. McPetrie. (*Proc. IEE*, pt. B, vol. 105, pp. 11–13; January, 1958.) Discusses some of the newer navigational aids which might be expected to come into civil and military use during the next few years.

621.396.962.3:621.374.5 1156

Pulse Compression: Part 2—R. Körnert. (See 1048.)

621.396.963.3:621.396.822 1157

Detection of Pulse Signals in Noise: Trace-to-Trace Correlation in Visual Displays—M. I. Skolnik and D. G. Tucker. (*J. Brit. IRE*, vol. 17, pp. 705–706; December, 1957.) Comment on 3149 of 1957 and author's reply.

621.396.969.3 1158

"Angels" on Centimetric Radars caused by Birds—W. G. Harper. (*Nature, London*, vol. 180, pp. 847–849; October 26, 1957.) A study of unexplained echoes recorded at East Hill 30 miles NW of London from 1952 onwards suggests that these are due to birds only and not to insects or meteorological effects. 10-cm-λ ppi equipment was used with peak power 500 kw, pulse length 2 μs, and half-power beam width $1\frac{1}{4}$°.

535.215:546.289 1159

Copper-Doped Germanium as a Model for High-Resistivity Photoconductors—P. J. van Heerden. (*Phys. Rev.*, vol. 108, pp. 230–238; October 15, 1957.) In the electrical behavior of high-resistivity copper-doped Ge the nature of the electrodes plays an essential role. The primary and secondary photocurrents can be sharply distinguished experimentally and form the basic concepts for an explanation of the photocurrent. The secondary photocurrent can be quantitatively expressed in terms of the primary photocurrent and the variation of the dark current with a voltage step-function. Observations are given on the space-charge-limited current and some light is thrown on the mechanism of electrical breakdown.

535.215:546.682.86 1160

New Infrared Detectors using Indium Antimonide—D. G. Avery. (*J. Sci. Instr.*, vol. 34, pp. 394–395; October, 1957.) Description of the properties of photoconductive InSb detectors at room temperature, and of experimental work on *p-n* junctions cooled to 90°K used as photovoltaic cells.

535.215:546.817.231 1161

The Reaction of Oxygen with Lead Selenide—R. H. Jones. (*Proc. Phys. Soc.*, vol. 70, pp. 1025–1032; November 1, 1957.) Oxidation processes were investigated at various temperatures from ambient to 280°C, to gain fundamental information about the photoconductive sensitization of evaporated layers of lead selenide by oxygen. Results indicate that two processes occur during oxidation and this postulation is supported by earlier electrical measurements on similar layers. Theories of photoconductivity in such layers are discussed in relation to these experiments. (See also 3887 of 1957.)

535.37:53.082.5 1162

The Measurement of the Decay of Phosphors by means of a New Type of Phosphoroscope—D. Hahn and H. J. Kösel. (*Z. angew. Phys.*, vol. 9, pp. 137–140; March, 1957.) Decay curves for periods ranging from 10^{-1} to 10^{-4} s can be obtained.

535.37:546.472.21 1163

Influence of the Intensity and Duration of Excitation on the Photoelectric Properties of Copper-Activated Zinc Sulphide—B. Hagène and J. J. Le Fèvre. (*J. Phys. Radium*, vol. 18, pp. 412–413; June, 1957.)

535.372 1164

The Effectiveness of Some Fluorescence Quenchers in Relation to the Position of the Fluorescence Spectrum—V. V. Zelinskiĭ, V. P. Kolobkov, and N. I. Kondaraki. (*Dokl. Ak. Nauk S.S.S.R.*, vol. 117, pp. 391–394; November 21, 1957.) The effect of aniline and triethylamine iodide on several types of amino-N-methylphthalimide are examined and results tabulated.

535.376:546.472.21 1165

The Electroluminescence of ZnS Phosphors as an Equilibrium Process—W. Lehmann. (*Optik, Stuttgart*, vol. 14, pp. 319–327; July/August, 1957.) The time average of electroluminescence intensity is characterized by dynamic equilibrium between monomolecular excitation and a recombination mechanism. The dependence of electroluminescence on field strength, temperature, activator, and electron-trap concentrations, is considered theoretically and a comparision is made with experimental results.

537.226/.227:546.431.824-31:621.317.335.3.029.64 1166

The Dielectric Constant of Barium Titanate at 10 Gc/s—H. J. Schmitt. (*Z. angew. Phys.*, vol. 9, pp. 107–111; March, 1957.) The results of measurements at 9.4 kmc are discussed. Values of dielectric constant obtained range from 173 to 560 according to the grade of material. Comparison with 2-kc values indicates a relaxation frequency of about 10^{10} cps which appears to be increased when a strong steady field is applied.

537.227:546.431.824-31 1167

Domain Conversion of Multidomain Barium-Titanate Single Crystals—P. H. Fang, S. Marzullo, and W. S. Brower. (*Phys. Rev.*, vol. 108, pp. 242–243; October 15, 1957.) "It is found that complete domain conversion of single crystal $BaTiO_3$ can be achieved by passing the crystal through the orthorhombic-tetragonal transition under an applied dc field. By using this process either complete *a* or *c* domain crystals can be prepared."

537.311.33 1168

On the Electron-Lattice Interaction in Nonpolar Semiconductors—S. Koshino. (*Progr. Theor. Phys.*, vol. 18, pp. 23–32; July, 1957.) The acoustical-mode scattering mobility is derived for the two-phonon process which is shown to be predominant at higher temperatures. The transition from the one-phonon to the two-phonon region can account for the dependence of mobility on temperature for both electrons and holes in Ge and Si.

537.311.33 1169

Studies on Group III–V Intermetallic Compounds—C. Kolm, S. A. Kulin, and B. L. Averbach. (*Phys. Rev.*, vol. 108, pp. 965–971; November 15, 1957.) The effects of additions of Si, Ge, Sn, and Pb to GaAs and InSb are described. The forbidden energy gap (determined by infrared absorption measurement), varies inversely with the lattice spacing for the GaAs alloys, except for the GaAs-Ge alloy where the reverse effect occurs. From infrared transmission and electrical resistivity data it is concluded that the Group IV elements substitute for nearest-neighbor pairs in GaAs, but with Si and Ge in InSb it appears that the substitution of only one atom in the unit polyhedron occurs. Similar measurements on InSb-GaSb alloys are also described.

537.311.33 1170

Calculation of the Electrical Conductivity of a Semiconductor with High Impurity Concentration—R. Ziegenlaub. (*Dokl. Ak. Nauk S.S.S.R.*, vol. 117, pp. 395–398; November 21, 1957.) A brief mathematical analysis based on a new method for the evaluation of conductivity in semiconductors without the use of kinetic equations. The scattering of current carriers by ionized impurity centers is investigated.

537.311.33 1171

Electron and Hole Mobility in a Liquid Semiconductor—I. Z. Fisher. (*Dokl. Ak. Nauk S.S.S.R.*, vol. 117, pp. 399–402; November 21, 1957.) A mathematical analysis of the diffusion of electrons and holes in a liquid is presented and an expression for electron or hole mobility is derived. The diffusion coefficient is estimated to be approximately 1 cm².

537.311.33:535.214 1172

Light-Induced Plasticity in Semiconductors—G. C. Kuczynski and R. F. Hochman. (*Phys. Rev.*, vol. 108, pp. 946–948; November 15, 1957.) "It has been found that light of wavelength between 2.0 and 4.0μ, or shorter than 0.4μ, decreases the hardness of the surface layer of *n*-type germanium by 10 to 60 per cent. The softened layer extends to a depth of one to two microns. A similar but less intense effect was observed in *p*-type germanium and *n*-type InSb and InAs. On the other hand, *p*-type silicon seems to soften to a greater extent than germanium. The effect is proportional to light intensity and is affected by surface preparation."

537.311.33:535.215 1173

Measurement of Minority-Carrier Lifetimes with the Surface Photovoltage—E. O. Johnson. (*J. Appl. Phys.*, vol. 28, pp. 1349–1353; November, 1957.) The method is based on the junction-like properties of a semiconductor surface. With a capacitive contact the photovoltage (in the millivolt range) may be detected, and is a linear function of excess-carrier density. This gives the same carrier decay constant as the photoconductivity method when the lowest diffusion mode prevails.

535.311.33:535.37 1174

Anomalous Variation of Band Gap with Composition in Zinc Sulpho- and Selenotellurides—S. Larach, R. E. Shrader, and C. F. Stocker. (*Phys. Rev.*, vol. 108, pp. 587–589; November 1, 1957.) "A monotonic variation of band gap with composition occurs for many binary solid solutions. Of some Group II–Group VI systems, ZnS-ZnSe shows this type of variation of band gap with composition, whereas ZnSe-ZnTe, ZnS-ZnTe show an anomalous minimum in a plot of band gap vs composition of the solid solution."

537.311.33:539.16 1175

Radiation-Induced Expansion of Semiconductors—D. Kleitman and H. J. Yearian. (*Phys. Rev.*, vol. 108, p. 901; November 1, 1957.) Small areas (3 mm$\times$3mm) of polished GaSb, InSb, and Ge were irradiated with 9-MeV deuterons at temperatures below -130°C. Surface contours were then determined by interferometers.

537.311.33:546.28 1176

Effect of Oxygen on the Carrier Lifetime in Silicon—D. H. Roberts, P. H. Stevens, and P. H. Hunt. (*Nature, London*, vol. 180, pp. 665–666; September 28, 1957.) Results of measurements made using an infrared absorption technique to determine oxygen content show that the presence of oxygen increases the carrier lifetime. This is explained by the hypothesis of recombination centers which can be nullified by chemical reaction with the oxygen or which compete with it for a place in the lattice.

537.311.33:546.28 1177

Behaviour of Oxygen in Plastically Deformed Silicon—S. Lederhandler and J. R. Patel. (*Phys. Rev.*, vol. 108, pp. 239–242; October 15, 1957.) For a deformed sample with a dislocation density of 10^7/cm², the amplitude of the 9-μ infrared absorption band is reduced appreciably in 15 min at 1000°C, while the undeformed sample with a dislocation density of 10^4/cm² requires about 12 h. The effect of deformation on light scattering is described.

537.311.33:546.28 1178

Energy Band Structure in Silicon Crystals by the Orthogonalized Plane-Wave Method—F. Bassani. (*Phys. Rev.*, vol. 108, pp. 263–264; October 15, 1957.) Energies at the point $\mathbf{K}=2\pi a^{-1}(1, 0, 0)$ are calculated, and energy curves are drawn as a function of **K** in the [100] direction.

537.311.33:546.28 1179

Infrared Absorption in *n*-Type Silicon—W. Spitzer and H. Y. Fan. (*Phys. Rev.*, vol. 108, pp. 268–271; October 15, 1957.) Experiments using samples of various impurity in the spectral region 1–45μ, show an absorption band at 1.5–5μ as well as a smooth rise with wavelength. The absorption is proportional to carrier concentration.

537.311.33:546.28 1180

Annealing of Electron Bombardment Damage in Silicon Crystals—G. Bemski and W. M. Augustyniak. (*Phys. Rev.*, vol. 108, pp. 645–648; November 1, 1957.) The annealing was studied between 200°C and 400°C by observing the recovery of the minority-carrier lifetime. The annealing was found to proceed with an activation energy of 1.3 ev. The significance of the results is discussed briefly.

537.311.33:546.28 1181

Thermal Breakdown in Silicon *p-n* Junctions—J. Tauc and A. Abrahám. (*Phys. Rev.*, vol. 108, pp. 936–937; November 15, 1957.) "It is shown that the normal avalanche breakdown of well-etched silicon *p-n* junctions can pass at higher temperatures into a kind of thermal breakdown at which most of the current flows through a small hot hole in the potential barrier."

537.311.33:546.28 1182

Surface Protection and Selective Masking during Diffusion in Silicon—C. J. Frosch and L. Derick. (*J. Electrochem. Soc.*, vol. 104, pp. 547–552; September, 1957.) In the apparatus described a carrier gas containing impurity elements passes over heated Si. Carriers such as H_2 or He produce serious pitting of the Si surface; oxidizing carriers, however, cause a protective SiO_2 layer to form. This layer also provides a selective mask against diffusion of certain donors and acceptors into Si.

537.311.33:546.28 1183

Preparation of Large-Area *p-n* Junctions in Silicon by Surface Melting—E. Billig and D. B. Gasson. (*J. Appl. Phys.*, vol. 28, pp. 1242–1245; November, 1957.) The surface melting was produced either by using direct rf power coupling or by using radiation heating. The junctions, which were formed by overdoping of the molten material, sustained a peak inverse voltage of approximately 40ρ_n and 10ρ_p for *n*-type and *p*-type base material of resistivity ρ_n and ρ_p, respectively.

537.311.33:546.28:621.314.63 1184

Semiconductor Properties of Recrystallized Silicon in Aluminium Alloy Junction Diodes—R. A. Gudmundsen and J. Maserjian, Jr. (*J. Appl. Phys.*, vol. 28, pp. 1308–1316; November, 1957.) The Al concentration was found to be largest in the region first recrystallized decreasing in a linear manner to the surface immediately beneath the Al-Si eutectic; the average concentration was about 7×10^{18} cm^{-3}. The average Hall hole mobility was about 55 cm²/v sec while the conductivity varied from 72 to 35 mho. cm^{-1}.

537.311.33:546.289 1185

Dislocation Arrays in Germanium—W. W. Tyler and W. C. Dash. (*J. Appl. Phys.*, vol. 28, pp. 1221–1224; November, 1957.) "Dislocation arrays have been observed in deformed germanium using a technique based on the selective etching of lithium precipitates which prefer to nucleate on or near dislocations. Evidence is presented which indicates that both internal Frank-Read sources and surface dislocation sources are of importance in the deformation of germanium."

537.311.33:546.289 1186

Hillocks, Pits, and Etch Rate in Germanium Crystals—B. W. Batterman. (*J. Appl. Phys.*, vol. 28, pp. 1236–1241; November, 1957.) Symmetric, faceted etch hillocks have been produced on certain Ge surface orientations by an H_2O_2-HF etchant. General characteristics of hillock formation are explained in terms of the variation of etch rate with surface orientation.

537.311.33:546.289 1187

Surface Studies on Single-Crystal Germanium—S. G. Ellis. (*J. Appl. Phys.*, vol. 28, pp. 1262–1269; November, 1957.) The condition of the surface has been studied by chemical, electron-microscope, electron-diffraction, and other techniques after several of the standard etching procedures. It was found that the surface was often partially covered by particles believed to be GeO. Details of an etch which gives a controlled thickness of GeO and of an etch which minimizes the oxide formations are given.

537.311.33:546.289:538.24 1188

Magnetic Susceptibility of Germanium—R. Bowers. (*Phys. Rev.*, vol. 108, pp. 683–689; November 1, 1957.) A study of the contribution of extrinsic charge carriers to the magnetic susceptibility of Ge as deduced from measurements of susceptibility between 1.3°K and 300°K. The results are compared with theoretical estimates based on the effective-mass values given by cyclotron-resonance experiments. The susceptibility of high-purity Ge is found to be independent of temperature below 60°K.

537.311.33:546.289:538.63 1189
Effect of Impurity Scattering on the Magnetoresistance of *n*-Type Germanium—M. Glicksman. (*Phys. Rev.*, vol. 108, pp. 264–267; October 15, 1957.) A decrease in anisotropy of the magnetoresistance with increased ionized-impurity content and decreased temperature is shown. Deviations from symmetry conditions are observed for electron concentrations $>4\times10^{16}cm^{-3}$ at 77°K.

537.311.33:546.47–31 1190
Hall-Effect Studies of Doped Zinc Oxide Single Crystals—A. R. Hutson. (*Phys. Rev.*, vol. 108, pp. 222–230; October 15, 1957.) Doping with H or interstitial Zn or Li permitted a single-donor-level analysis. The conductivity and Hall coefficient were measured between 55°K and 300°K. An experimental curve shows Hall mobility between 80°K and 600°K. At 300°K it is 180 cm²/v sec, and increases with decreasing temperature (T). The probable active scattering modes are discussed. Optical-mode scattering appears important above 200°K. Curves of electron concentration and carrier concentration as functions of $1/T$ are given. The low-frequency dielectric constant of ZnO was found to agree with Kamiyoshi's value of 8.5.

537.311.33:546.561–31:539.23 1191
Crystal Structure and Magnetic Susceptibility of Rectifying Cuprous Oxides—K. R. Dixit and V. V. Agashe. (*Indian J. Phys.*, vol. 31, pp. 466–482; September, 1957.) The properties of films formed at different temperatures have been studied. Above 800°C rectification is appreciable and is accompanied by changes in crystal structure and susceptibility. A mechanism of film formation giving crystallites of Cu_2O with an excess of oxygen is suggested. See also 1140 of 1957 (Dixit and Agashe).

537.311.33:546.682.86:537.312.9 1192
Piezoresistance of Indium Antimonide—R. F. Potter. (*Phys. Rev.*, vol. 108, pp. 652–658; November 1, 1957.) The piezoresistivity coefficients were measured for both *n*- and *p*-type InSb over the range 77°K–300°K. The results for extrinsic *p*-type material are similar to those for *p*-type Si and Ge. The results for extrinsic *n*-type material confirm the picture of a conduction band having its minimum at the center of the Brillouin zone.

537.311.3:546.682.86:538.63 1193
Galvanomagnetic Effects in *n*-type Indium Antimonide at Very Low Temperatures—Y. Kanai and W. Sasaki. (*J. Phys. Soc. Japan*, vol. 12, p. 1169; October, 1957.) An anomalous Hall effect was investigated.

537.311.33:546.817.241:538.63 1194
The Galvanomagnetic Effects in Single Crystal of PbTe—K. Shogenji and S. Uchiyama. (*J. Phys. Soc. Japan*, vol. 12, p. 1164; October, 1957.) Results of measurements of magnetoresistance and planar Hall effect in *p*-type material are given.

537.311.33:546.824–31 1195
Electrical Resistivity of some Titanium Dioxide Semiconductors—V. Andresciani, L. Nicolini, and D. Sette. (*Note Recensioni Notiz.*, vol. 6, pp. 511–524; July/August, 1957.) Report of experimental investigations on three specimens, one being pure rutile, and the others containing small additions of Ba and Ca, respectively. See also 757 of 1954 (Breckenridge and Hosler).

537.312.62:538.569.4 1196
Microwave Measurements of the Energy Gap in Superconducting Aluminium—M. A. Biondi, M. P. Garfunkel, and A. O. McCoubrey. (*Phys. Rev.*, vol. 108, pp. 495–497; October 15, 1957.) Experimental evidence is presented for a temperature-dependent energy-gap, extrapolating to a value between 3.0 and 3.5 kT_c at absolute zero.

537.312.62:538.569.4 1197
Millimetre-Wave Studies of Superconducting Tin—M. A. Biondi, A. T. Forrester, and M. P. Garfunkel. (*Phys. Rev.*, vol. 108, pp. 497–498; October 15, 1957.) A description of measurements of the temperature variation of the power transmitted by a superconducting tin waveguide, and brief comparison with theory.

537.312.62:539.23 1198
Conductivity of Superconducting Films for Photon Energies between 0.3 and $40kT_c$—R. E. Glover III, and M. Tinkham. (*Phys. Rev.*, vol. 108, pp. 243–256; October 15, 1957.) An investigation of far-infrared and mm-wave transmission through thin superconducting lead and tin films are reported, including measurements in the unexplored frequency region in which superconduction changes to normal. At $T=0$, σ_1 is very small for photon energies below 3 kT_c; above $3kT_c$, σ_1 rises rapidly to a limit σ_n at 20 kT_c. A gap in the excitation spectrum, of width 3 kT_c, is suggested.

538.22 1199
Magnetic Properties of the $FeTiO_3$-Fe_2O_3 Solid Solution Series—Y. Ishikawa and S. Akimoto. (*J. Phys. Soc. Japan*, vol. 12, pp. 1083–1098; October, 1957.) Experiments on $xFeTiO_3\cdot(1-x)Fe_2O_3$, showed that its properties could be classified into three groups according to x. Simple molecular field theory does not explain all magnetic properties although a preferential distribution of Ti ions could account for some of them.

538.22 1200
Magnetic Properties of $NiTiO_3$-Fe_2O_3 Solid Solution Series—Y. Ishikawa. (*J. Phys. Soc. Japan*, vol. 12, p. 1165; October, 1957.)

538.221 1201
Effect of Isothermal Magnetic Annealing on the Magnetic Properties at Room Temperature in Ni-Co Alloys—H. Masumoto, H. Saitô, and M. Takahashi. (*Sci. Rep Res. Inst. Tohoku Univ. Ser. A*, vol. 9, pp. 374–394; October, 1957.) The magnetostriction and the magnetization at room temperature in a wide range of Ni-Co alloys were measured after annealing at high temperatures under external magnetic fields of various strengths. The results of the measurements are presented in 36 families of graphs.

538.221:[621.318.124+621.318.134 1202
Convention on Ferrites—(*Proc. IEE*, pt B, vol. 104, suppl. no. 5, pp. 127–266; 1957.) The following papers were included among those read at the IEE Convention held in London October 29–November 2, 1956.

Chemical and Physical Properties and Preparation

The Chemistry and Crystal Structure of Ferrites and Other Magnetic Oxides—A. J. E. Welch (pp. 138–140).

Structural and Magnetic Properties of Solid Solutions of Lithium Ferrite with Cadmium Ferrite and with Lithium Aluminate—A. E. Carter, P. A. Miles, and A. J. E. Welch (pp. 141–144).

The Physical and Chemical Properties of some Nickel-Zinc Ferrite Compounds—N. C. Tombs and J. Watkins (pp. 145–151).

A Note on Crystal Formation in the Metallic Oxides—L. R. Maxwell and S. J. Pickart (pp. 152–153).

Ferrite Materials for Faraday Rotation at Wavelengths of 3, 6 and 10 cm—P. E. Ljung (pp. 154–158).

The Preparation of Magnesium-Manganese Ferrite for Microwave Applications—A. E. Robinson (pp. 159–164.)

The Properties of Manganese-Zinc Ferrites and the Physical Processes Governing them—C. Guillaud (pp. 165–173).

Discussion (pp. 174–178).

Magnetic Spectra

Magnetic Spectra—J. B. Birks (pp. 179–188).

Losses in Ferrites: Single-Crystal Studies—J. K. Galt (pp. 189–197).

Effect of Magnetocrystalline Anisotropy on the Magnetic Spectra of Mg-Fe Ferrites—G. T. Rado, V. J. Folen, and W. H. Emerson (pp. 198–205).

Ferrimagnetic Resonance in a Magnesium-Manganese Ferrite—K. J. Standley and J. Peters (pp. 206–208). See also 1153 of 1957 (Standley and Stevens).

Complex Susceptibility of High Resistivity Ferrites—P. M. Prache and B. Chiron (pp. 209–212).

Discussion (pp. 213–216).

Molecular Interaction

Neutron-Diffraction Studies of the Manganese-Magnesium Ferrite System—R. Nathans, S. J. Pickart, S. E. Harrison, and C. J. Kriessman. (pp. 217–220). See also 510 of 1957 (Callen *et al.*).

Magnetic Exchange Mechanisms in Magnesium-Manganese Ferrites—W. P. Osmond (pp. 221–227).

Discussion (pp. 228–230).

Direct-Current and Low-Frequency Properties

Heat Changes Accompanying Magnetization Processes in Ferrites—L. F. Bates and D. A. Christoffel (pp. 231–237). See also 3569 of 1957 (Christoffel).

Magnetocrystalline Anisotropy in Cobalt-Substituted Magnetite Single Crystals—L. R. Bickford, Jr., J. M. Brownlow, and R. F. Penoyer (pp. 238–244).

Discussion (pp. 245–248).

New Materials

A New Class of Oxidic Ferromagnetic Materials with Hexagonal Crystal Structures—G. H. Jonker, H. P. J. Wijn, and P. B. Braun (pp. 249–254).

Saturation Magnetization of Some Ferrimagnetic Oxides with Hexagonal Crystal Structures—E. W. Gorter (pp. 255–260).

Crystalline Structure and Magnetic Properties of Ferrites having the General Formula $5Fe_2O_3\cdot3M_2O_3$—F. Bertaut and R. Pauthenet (pp. 261–264).

Discussion (pp. 265–266).

538.221:[621.318.124+621.318.134 1203
Synthesis of Some Ferrites—H. Kedesdy and A. Tauber. (*J. Metals, New York*, vol. 9, pp. 1140–1148; September, 1957.) Results are discussed of systematic investigations of the formation of a basic ferrite, such as Ni ferrite, a mixed ferrite, such as NiZn ferrite, and a territe involving a complex formation process, such as Mn ferrite. Trends in magnetic characteristics are related to changes in firing cycle and furnace atmosphere.

538.24:621.318.1 1204
Investigations of the Internal Demagnetization Factor—G. Vogler. (*Ann. Phys., Leipzig*, vol. 19, pp. 229–232; December 20, 1956.) Report on magnetic measurements on substances consisting of spherical ferromagnetic particles embedded in nonmagnetic material. The demagnetization factor is defined and its characteristics are discussed.

621–762.8:621.318.134 1205
Hermetic Seal for Ferrites—A. H. Iversen. (*Rev. Sci. Instr.*, vol. 28, pp. 797–799; October, 1957.) The electrical characteristics of ferrites are adversely affected by moisture. A hermetic seal is obtained by coating the ferrite surface with a thin layer of glass after sintering a thin layer of finely ground ferrite surface to reduce porosity. The glass seal has good mechanical and thermal shock properties and negligible effect on electrical properties.

537.311.33 1206
Rectifying Semiconductor Contacts—[Book Review]—H. K. Henisch. Publishers: Clarendon Press, Oxford, 372 pp. 70s, 1957. (*Nature, London*, vol. 180, pp. 566–567; September 21, 1957.) "... a basic fund of knowledge upon which any investigator of surface and contact phenomena may usefully draw."

MEASUREMENTS AND TEST GEAR

53.082.5:519.24:621.397.3 1207
Optical Autocorrelation Measurement of Two-Dimensional Random Patterns—L. S. G. Kovásznay and A. Arman. (*Rev. Sci. Instr.*, vol. 28, pp. 793–797; October, 1957.) A function of two independent variables can be regarded as an image. Using a novel optical method the autocorrelation coefficient of such a function is formed as another image. The basic optical equipment is improved by adopting electronic scanning techniques.

621.3.018.41(083.74):621.396.11.029.45 1208
Comparison of Caesium Resonators by Transatlantic Radio Transmission—L. Essen, J. V. L. Parry, and J. A. Pierce. (*Nature, London*, vol. 180, pp. 526–528; September 14, 1957.) A resonator at Teddington was indirectly compared with one of a different type at Camden, N. Y., by means of the GBR 16-kc transmission from Rugby at about 0300 G.M.T. A table of results from November 16, 1956 to June 1, 1957 is given. Simultaneous measurements of suitable duration using daytime transmission should give results accurate to at least one part in 10^{10}.

621.317.2.013.782 1209
A Test Chamber Magnetically Screened by Magnetized Dynamo Laminations—W. Albach and G. A. Voss. (*Z. angew. Phys.*, vol. 9, pp. 111–115; March, 1957.) Three concentric cylinders of lamination material of 120, 190, and 300 cm diameter form a screened enclosure. A 25-fold increase of the screening effect is achieved by magnetizing the toroidally wound cylinders with 50 cps ac.

621.317.335:518.4 1210
Rapid Measurement of Dielectric Constant and Loss Tangent—D. M. Bowie and K. S. Kelleher. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. 4, pp. 137–140; July, 1956. Abstract, PROC. IRE, vol. 44, p. 1898; December, 1956.)

621.317.335.2:621.319.4 1211
Precision Measurements on an Imperfectly Screened Three-Plate Capacitor—W. Wiessner. (*Z. angew. Phys.*, vol. 9, pp. 120–125; March, 1957.) Sources of errors and their elimination are discussed.

621.317.335.3.029.64:537.226.8 1212
A Three-Centimetre Microwave Bench for Studying the Pressure and Temperature Effects on Dielectric Behaviour of Gases—Krishnaji and G. P. Srivastava. (*J. Sci. Indus. Res.*, vol. 16A, pp. 289–294; July, 1957.)

621.317.34.029.63/.64:621.372.5 1213
Three-Point Method of Measuring U.H.F. Quadripoles—L. Mollwo. (*Hochfreq. u. Elektroak.*, vol. 65, pp. 188–192; May, 1957.) A simplified method of determining network characteristics is described. It is applicable to simple types of quadripole and is based on locating three node positions.

621.317.35:534.232 1214
Measurements on Electromechanical Transducers—H. G. Diestel. (*Acustica*, vol. 6, pp. 357–360; 1956. In German.) Apparatus is described for the continuous recording of frequency response curves of electromechanical transducers. Feedback into the driving system is compensated by means of a control circuit.

621.317.351.018.756 1215
A Simple Pulse Spectrograph—N. R. Hansen. (*J. Sci. Instr.*, vol. 34, pp. 402–404; October, 1957.) Pulses from a radiation detector are displayed as bright dots along a nonlinear vertical timebase on a cr tube screen. The timebase is deflected towards the right by an amount proportional to the pulse amplitude. A photograph of the screen with time exposure thus shows a variable density pattern. Lines of constant density represent pulse frequency as a function of pulse amplitude.

621.317.361 1216
Pulse-Type Frequency Measurement—J. C. Muller. (*Wireless World*, vol. 64, pp. 83–85; February, 1958.) Discusses the design of a beat-counting frequency comparator for the range 60–600 kc in which an unknown frequency is compared with a harmonic of a 1-kc standard. The harmonics are derived using a thyratron/delay-line circuit to generate a rectangular 1-μs pulse.

621.317.382.029.64:538.632 1217
The Hall Effect and its Application to Power Measurement at 10 Gc/s—H. E. M. Barlow and S. Katoka. (*Proc. IEE*, pt. B, vol. 105, pp. 53–60; January, 1958.) The instrument described is a transmission wattmeter based on the Hall effect produced in single crystals of n- and p-type Ge erected on the axis of a hollow metal rectangular waveguide carrying the power. The preparation of suitable crystals is described, and the reduction of errors due to the finite admittance of the crystal, thermal emf and residual rectifier output is discussed. The Hall coefficient at nearly 10 kmc appeared to be of the same order of magnitude as the theoretical dc value. See also 1488 of 1956 (Barlow and Stephenson.)

621.317.6.029.45 1218
Measurement of Voltage Ratio at Audio Frequencies—W. C. Sze. (*Commun. and Electronics*, no. 32, pp. 444–449; September, 1957.) Detailed description of bridge circuits for precise voltage ratio and phase angle measurements. An accuracy within 0.005 per cent in ratio and 0.2′ in phase angle is achieved at frequencies from 100 cps to 10 kc.

621.317.7:621.376.2:621.396.62 1219
Modulation Check Meter for the Receiver—M. W. Kirby. (*Short Wave Mag.*, vol. 15, pp. 350–351; September, 1957.) A circuit for direct modulation depth measurement.

621.317.7:621.383.27:621.396.822 1220
Photomultiplier Tubes as Standard Noise Sources—Chenette, Shimeda, and van der Ziel. (See 1286.)

621.317.714.023.43 1221
Clamp-on Microammeter Measures A.C. Current—G. F. Montgomery and C. Stansbury. (*Electronics*, vol. 30, pp. 152–153; December 1, 1957.) "Small toroidal current transformer clamped around unknown current measures 0 to 200-microampere range over frequency band of 50 cps to 100 kc with negligible reaction upon measured circuits. Feedback to tertiary winding supplies frequency correction to transistor amplifier."

621.317.725.082.72 1222
A Compact Rotating-Electrode Voltmeter for Wide Voltage Ranges—W. Knauer. (*Z. angew. Phys.*, vol. 9, pp. 115–119; March, 1957.) In the portable es voltmeter described the rotating multisegment electrode is motor-driven and contained in a pressure chamber together with the hv electrode. Linear indication is provided in 6 ranges, 0–3 v up to 0–300 kv.

621.317.733.029.5 1223
A 'Subtractor' for Radio-Frequency Bridges—E. A. Faulkner. (*J. Sci. Instr.*, vol. 34, p. 461; November, 1957.) A two-tube device with high-impedance input terminals. The output is proportional to the vector difference of the input potentials, with respect to earth, and can be fed into a normal receiver. Balancing is simple at frequencies up to 5 mc. A full circuit diagram is given.

621.317.755 1224
Sampling Oscilloscope for Statistically Varying Pulses—R. Sugarman. (*Rev. Sci. Instr.*, vol. 28, pp. 933–938; November, 1957.) Pulses of constant or random repetition rate may be displayed on the oscilloscope which has a rise time of 5.6×10^{-10} seconds, maximum sweep of 10^{-10} seconds/inches and a maximum sensitivity of 15 mv/inch.

621.317.755.087:621.3.015.3 1225
A High-Sensibility Cathode-Ray Tube for Millimicrosecond Transients—K. J. Germeshausen, S. Goldberg, and D. F. McDonald. (IRE TRANS. ON ELECTRON DEVICES, vol. 4, pp. 152–158; April, 1957. Abstract, PROC. IRE, vol. 45, p. 1164; August, 1957.)

621.317.784 1226
A Precision Thermoelectric Wattmeter for Power and Audio Frequencies—J. J. Hill. (*Proc. IEE*, pt. B, vol. 105, pp. 61–68; January, 1958.) Compensated thermojunctions are used with negative-feedback amplifiers to achieve accuracies within about 0.1 per cent over the range 200 cps–10 kc and 0.3 per cent for 50 cps–30 kc. A method of extending the frequency range to 100 kc is described.

621.317.794:621.372.822 1227
Transverse-Film Bolometers for the Measurement of Power in Rectangular Waveguides—J. A. Lane. (*Proc. IEE*, pt. B, vol. 105, pp. 77–80; January, 1958.) Describes a plunger-backed bolometer consisting of a relatively narrow absorbing strip (platinum sputtered on mica) placed symmetrically in the region of the transverse plane where the electric and magnetic fields are substantially uniform. Substitution of dc for microwave power is therefore used, the power range being 1–100 mw with an error of 3 per cent or less.

621.317.794:621.396.822 1228
Considerations in High-Sensitivity Microwave Radiometry—P. D. Strum. (PROC. IRE, vol. 46, pp. 43–60; January, 1958.) A discussion of the factors which limit sensitivity of receiving systems. Apart from fluctuations in the receiver and its antenna, background radiation from space, atmospheric oxygen, water vapor and earth-bound radiators must be considered. A square-wave switched type of receiver is most likely to yield satisfactory sensitivity and the optimum performance is discussed.

621.317.794:621.396.822:523.164 1229
A Broad-Band Microwave Source Comparison Radiometer for Advanced Research in Radio Astronomy—F. D. Drake and H. I. Ewen. (PROC. IRE, vol. 46, pp. 53–60; January, 1958.) A travelling-wave-tube radiometer operating at 8 kmc with bandwidth 1 kmc and sensitivity of the order of 0.01°K is described. Effects of gain fluctuations are eliminated by introducing compensating noise. Results obtained using a 28-foot reflector are discussed.

621.317.794.089:523.164 1230
A Method of Calibrating Centimetric Radiometers using a Standard Noise Source—J. S. Hey and V. A. Hughes. (PROC. IRE, vol. 46, pp. 119–121; January, 1958.) The method is suitable for aerial temperatures in the range 0–1000°K with an accuracy within 1°K. The source is an argon discharge tube.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.384.61 1231
Circular Electron Accelerators—M. Seidl. (*Slab. Obz., Praha*, vol. 17, pp. 698–702; December, 1956.) A description is given of particle accelerators made in Czechoslovakia including a 15 mev betatron made in 1955, earlier smaller betatrons and a 3-mev synchrotron.

621.384.61 1232
C.E.R.N. 600-MeV Synchrocyclotron—(*Engineer, London*, vol. 204, pp. 538–540; October 11, 1957.) Constructional details of the equipment.

621.384.613 1233
Engineering Problems in the Development of Glass Betatron Toroids—S. J. Morrison. (*J. Soc. Glass Tech.*, vol. 40, pp. 520T–541T; December, 1956.) Three different types of toroidal glass vacuum chambers which have been developed for use in betatrons operating at 16 and 20 mev are reviewed.

629.113:621.317.39 1234
Electronics in Automobile Research—C. H. G. Mills and J. F. Winterbottom. (*Brit. Commun. Electronics*, vol. 4, pp. 410–415; July, 1957.) Illustrated description of equipment used for electronic recording of stress measurements, speed measurement by means of lamp/photocell units, and vehicle noise measurement using $\frac{1}{3}$-octave band-pass filters.

681.61:621.318.57 1235
***L'Electrostyl* to Revolutionize Stenography?**—A. V. J. Martin. (*Radio-Electronics*, vol. 28, pp. 38–39; March, 1957.) Brief description of an electronically controlled typewriter, a French invention in which the conventional keyboard is replaced by a flat panel with contact studs representing characters. A single conducting probe is drawn across the panel completing a switching circuit whenever a contact is touched; this actuates the keys of a typewriter.

PROPAGATION OF WAVES

621.396.11:523.5 1236
Meteor Signal Rates Observed in Forward Scatter—E. L. Vogan and L. L. Campbell. (*Can. J. Phys.*, vol. 35, pp. 1176–1189; October, 1957.) Measurements at 49.98 mc over a path 860 km long in eastern Canada are described. Variations in meteor signal rate on an hourly, daily, and monthly time scale show a large spread, the maximum rate occurring in the summer months. An explanation of certain features in the diurnal variation is based on possible attenuation in the E region.

621.396.11:551.510.535 1237
New Type of Scattering Echo Observed by the Ship-Borne Ionospheric Sounder over the Sea—H. Okamoto, M. Ose, and K. Aida. (*Rep. Ionosphere Res. Japan*, vol. 11, pp. 50–54; June, 1957.) A note on new short-range echoes having variation with time of day, frequency, and type of terrain; records are shown. The scattered echoes appear to be related to tropospheric conditions.

621.396.11:551.510.535 1238
The Attenuation of Radio Waves Reflected from the E Region of the Ionosphere—R. W. Meadows. (*Proc. IEE*, pt. B. vol. 105, pp. 22–26; January, 1958.) The absorption measured at 5.1 mc over a 740-km path between Slough and Inverness is compared with that measured simultaneously at Slough at vertical incidence on 2-5 mc. It is suggested tentatively that the absorption measurements which have been made for many years on 2 mc at noon at Slough should prove a useful guide to the absorption likely to be obtained over other oblique E-region paths during any part of the sunspot cycle, and they have been formulated accordingly.

621.396.11:551.510.535:550.38 1239
The Effect of the Earth's Magnetic Field on Absorption for a Single-Hop Ionospheric Path—R. W. Meadows and A. J. G. Moorat. (*Proc. IEE*, pt. B, vol. 105, pp. 33–37; January, 1958.) Magneto-ionic calculations show that deviative absorption is not necessarily negligible at vertical incidence for waves reflected from the E region at frequencies considerably below the penetration value. Deviative absorption on paths sufficiently oblique, however, is negligible. The effect of the earth's magnetic field on Martyn's absorption theorem is shown to be similar to the effect of losses due to partial reflections. The absorption to be expected on a radio path might best be calculated by applying the conventional nondeviative formula to measurements made at oblique rather than at vertical incidence.

621.396.11:551.510.535:621.396.812.3 1240
The Effect of Fading on the Accuracy of Measurement of Ionospheric Absorption—R. W. Meadows and A. J. G. Moorat. (*Proc. IEE*, pt. B, vol. 105, pp. 27–32; January, 1958.) An analysis is made of measurements of the amplitude of first-order reflections from the E region. The accuracy of measurement of the smoothed value of field strength is defined as the range of values having a 99 per cent chance of containing the correct value. On this basis, the accuracy of noon absorption, as calculated from a single day's observations at one frequency, has been estimated to be within about +4 and −12 db at vertical incidence and +4 and −11 db for a particular oblique-incidence case.

621.396.11.029.55 1241
Broadcasting Plan: 26 Mc/s and 'Phase: June 70'—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, vol. 11, pp. 43–49; March, 1957.) Critical comparison of some muf charts issued for various epochs by the International Frequency Registration Board with predictions based on the Gea method (see, *e.g.*, 558 of 1957).

RECEPTION

621.396.62:621.317.794 1242
Present and Future Capabilities of Microwave Crystal Receivers—C. T. McCoy. (PROC. IRE, vol. 46, pp. 61–66; January, 1958.) The effective temperatures of present-day receivers using crystal mixers are about 600°K at frequencies up to 10 kmc but increase rapidly at higher frequencies. Future developments in crystals and the introduction of cooling mechanisms may reduce effective temperatures to 150°K at all frequencies up to 100 kmc.

621.396.62:621.375.2 1243
High-Selectivity I.F./A.F. Amplifier Unit—A. C. Edwards. (*Short Wave Mag.*, vol. 15, pp. 344–349; September, 1957.) Design details for amateur construction of a circuit incorporating many of the latest reception techniques.

621.317.7:621.376.2:621.396.62 1244
Modulation Check Meter for the Receiver—Kirby. (See 1219.)

621.396.621:621.314.7 1245
Audio Stages for All-Transistor Portables—O. J. Edwards and L. H. Light. (*Mullard Tech. Commun.*, vol. 3, pp. 188–194; October, 1957.) A driver stage is transformer coupled to a class B single-ended output stage. Two versions are described having outputs of 200 and 100 mw and battery drains of 20 and 10 ma, respectively.

621.396.621:621.396.215 1246
New Frequency-Shift Telegraphy System—(*Wireless World*, vol. 64, pp. 93–94; February, 1958.) Brief account of a system of frequency-diversity reception described in 2278 of 1957 (Allnatt *et al.*).

621.396.621.53:621.396.662.1 1247
A New Method of Analysing Tracking in Superheterodyne Receivers—J. Holownia. (*NachrTech.*, vol. 7, pp. 289–293; July, 1957.) The method described is independent of the IF, the tuning range, and tuning system of the receiver.

621.396.665:621.375.4 1248
Transistor Amplified A.G.C—W. Woods-Hill. (*Wireless World*, vol. 64, pp. 94–95; February, 1958.) Description of a circuit suitable for miniature receivers, in which very large variations of signal strength are encountered. A single junction transistor combines the function of second detector, af amplifier, and agc amplifier.

621.396.822:621.317.794 1249
Considerations in High-Sensitivity Microwave Radiometry—Strum. (See 1228.)

621.396.828 1250
The Problem of Eliminating Radio Interference Originating from High-Frequency Heating Installations—E. Rosemann. (*Nachr-Tech.*, vol. 7, pp. 297–392; July, 1957.) The design of screens and suppressors is discussed and the difficulties of measuring the interference field strength are outlined.

STATIONS AND COMMUNICATION SYSTEMS

621.376.3 1251
Distortion in F.M. Systems—A. Ditl. (*Slab. Obz., Praha*, vol. 17, pp. 609–716; November 1956.) See also 3291 of 1957.

621.396.3:621.396.41 1252
A Predicted Wave-Signalling Phase-Shift Telegraph System—E. T. Heald and R. C. Clabaugh. (*Commun. and Electronics*, no. 31, pp. 316–319; July, 1957. Discussion, p. 319.) Highly stable mechanical resonator filters in the detection equipment are turned on and off in step with the arrival and conclusion of each signal pulse; the detector is thus subjected to noise only when a signal is present. A typical 40-channel system ("kineplex") is described, and a superiority in signal/noise ratio of 8 db over a wide-band frequency-shift keyed system is claimed.

621.396.3:621.396.43:523.5 1253
The Janet Communication System—G. W. L. Davis. (*Brit. Commun. Electronics*, vol. 4, pp. 402–407; July, 1957.) Description of the method of operation and a note on equipment used. See 3533 of 1956 (Cocking).

621.396.41 1254
The Canadian Transcontinental Microwave System—S. Bonneville. (*Commun. and Electronics*, no. 32, pp. 473–477; September, 1957.) The system under construction will cover 4800 miles with about 137 repeater points. It operates in the 4-kmc band, with six radio channels in each direction. Each 2-way channel can handle two television programs or 600 telephone circuits.

621.396.41:621.314.7 1255
A Transistorized Time-Division Multiplex Telegraph Set—F. D. Biggan. (*Commun. and Electronics*, no. 28, pp. 780–785; January, 1957.) A description of a fully transistorized 4-channel equipment which saves 80 per cent in weight and volume and 95 per cent in power consumption compared with the equivalent tube equipment Type AN/FGC-5.

621.396.41:621.376 1256
The Third Method of S.S.B—H. F. Wright. (*QST*, vol. 41, pp. 11–15; September, 1957.) A description of the method previously given by Weaver (PROC. IRE, vol. 44, pp. 1703–1705; December, 1956) which produces the ssb signal directly on the required radio frequency without the use of wide-band af phase-shift networks.

621.396.44:621.395.97:621.396.82 1257
High-Frequency Broadcasting over Lines and R.F. Interference—J. Meyer de Stadelhofen. (*Tech. Mitt. PTT*, vol. 35, pp. 257–265; July 1, 1957. In French.) A description of methods used for measuring and eliminating interference from various sources in program transmission over telephone lines, with particular reference to the Swiss system.

621.396.65 1258
Light-Route Microwave Systems in Canada—A. J. Dinnin. (*Commun. and Electronics*, no. 32, pp. 488–492; September, 1957.) Description of systems and equipment used for auxiliary links in the Trans-Canada microwave system (1254 above). Frequencies in the 900-mc 2-, 4-, and 6-kmc bands are used, and diversity reception is found necessary, particularly for telephone transmission and at the higher frequencies.

621.396.933:621.396.41 1259
Dual-Purpose Circuitry cuts Transceiver Size—P. G. Wulfsberg and C. H. Kirkpatrick. (*Electronics*, vol. 30, pp. 134–138; December 1, 1957.) Reflex circuit techniques are applied in the design of a 1750-channel transceiver using only 35 crystals and 28 tubes. The unit is intended for fixed- or mobile-station use in a ground-to-air communications system in the frequency range 225–400 mc. Transmitter power is 15 w and receiver sensitivity better than 5 μv for a 10-db signal-plus-noise/noise ratio. The operation of a "squelch" circuit is controlled by this ratio.

SUBSIDIARY APPARATUS

621–519:621.318.57:621.314.7 1260
Transistor Relays Have Low Idling Current—D. W. R. McKinley. (*Electronics*, vol. 30, p. 147; December 1, 1957.) Two transistor circuits for remote control are shown, which operate a 2- or 3-w electromechanical relay from the output signal of a receiver. This may be a carrier modulated at 1 kc or a pulsed microwave signal. Using Si transistors the idling current is a few microamps.

621.311.6:621.314.7:621.316.72 1261
Magnetic-Regulation Transistor Power Supply—L. F. Lyons. (*Commun. and Electronics*, no. 28, pp. 643–645; January, 1957.) The method uses saturable reactors in the input and load circuit of the supply. For closer regulation a silicon-diode reference voltage may be used.

621.311.62 1262
Compact Supplies Have Wide-Range Regulation—W. F. Schreiber. (*Electronics*, vol. 30, pp. 168–169; December 1, 1957.) "Use of solid-state rectifiers and high-current regulator tubes eliminates power transformer and reduces size and cost of power supplies. Units have regulation better than 1 per cent, ripple of 0.01 per cent and stacking factor of about 150 ma capacity per inch of rack space."

621.311.62:621.314.2 1263
A Method of Designing Small Power Transformers for Communication Equipment—H. S. Sear. (*Commun. and Electronics*, no. 28, pp. 758–761; January, 1957.)

621.311.62:621.372.54 1264
The Effect of Capacitance on Power-Supply Filter Bounce—D. T. Geiser. (*QST*, vol. 41, pp. 27–30; September, 1957.) A discussion on the design of power-supply filters.

621.311.62:621.375.4 1265
Regulated Supply for Transistor Amplifiers—H. R. Lowry. (*Radio TV News*, vol. 58, pp. 55, 193–194; October, 1957.

621.314.63 1266
Semiconductor Rectifiers—D. Ashby. (*Elec. Rev., London*, vol. 161, pp. 587–592; October 4, 1957.) Summary of characteristics and review of industrial applications of Si, Ge, Se, and copper-oxide rectifiers.

621.314.63 1267
A New Voltage-Divider Circuit for Semiconductor Rectifiers—I. K. Dortort. (*Commun. and Electronics*, no. 31, pp. 356–358; July, 1957. Discussion, p. 358.) Auxiliary transformers divide the inverse voltage across high-power rectifiers.

621.314.63:546.28 1268
The Fused Silicon Rectifier—H. W. Henkels. (*Commun. and Electronics*, no. 28, pp. 733–746; January, 1957.) A comprehensive review dealing with 1) rectification theory, 2) basic fabrication techniques and the properties of Si, the counter-electrode alloy solder, the base solder, and the base electrode, 3) encapsulation and heat dissipation, 4) theoretical and experimental properties of Si rectifiers.

621.314.63:[546.289+546.28 1269
Rating and Application of Germanium and Silicon Rectifiers—F. W. Gutzwiller. (*Commun. and Electronics*, no. 28, pp. 753–757; January, 1957.) A discussion of the factors limiting the life and rating of semiconductor rectifiers. Practical design criteria are considered including fusing, surge voltage protection, and parallel and series operation.

621.314.63–713:546.289 1270
Direct-Water-Cooled Germanium Power Rectifier—R. E. Wahl. (*Commun. and Electronics*, no. 28, pp. 832–841; January, 1957.) A description of the development, performance, and operation of various types of water-cooled unit.

TELEVISION AND PHOTOTELEGRAPHY

621.397.5 (71) 1271
The Development of Television in Canada—J. E. Hayes. (*Commun. and Electronics*, no. 32, pp. 482–484; September, 1957.) General description of the growth of the television service, which now covers 80 per cent of the population. Discussion of improvements in production facilities and system design is included.

621.397.5:535.623 1272
Colour-Television-System Performance Requirements—R. Kennedy. (*Commun. and Electronics*, no. 28, pp. 653–659; January, 1957.) A discussion of the basic requirements of phase, amplitude, and transient response which a transmission system must meet to handle a standard color-television signal. The effect of departures from the required standards are illustrated and test procedures are described.

621.397.5:621.391 1273
Some Relations between Television Picture Redundancy and Bandwidth Requirements—K. H. Powers and H. Staras. (*Commun. and Electronics*, no. 32, pp. 492–496; September, 1957. Discussion, p. 496.) Analysis of the statistical redundancy; *i.e.*, correlations existing between the elements of a given picture or between corresponding elements of successive frames, in terms of information theory. It is concluded that the best way of achieving bandwidth reduction is by cleverly degrading the information in a picture in such a way that the observer would not notice it appreciably.

621.397.611.2:681.42.089 1274
New Equipment and Methods for the Evaluation of the Performance of Lenses for Television—W. N. Sproson. (*B.B.C. Eng. Div. Monographs*, no. 15, pp. 5–16; December, 1957.)

621.397.62 1275
An Original Television Receiver for 4 Standards and 12 Channels—K. Lamus. (*Télévision*, no. 71, pp. 56–60; February, 1957.) Circuit details are given of a commercial-type receiver with switching facilities for adapting the set to the 625-line Belgian or European and the 819-line Belgian or French standards.

621.397.62 1276
Black Level—the Lost Ingredient in Television Picture Fidelity—R. G. Neuhauser. (*J. Soc. Mot. Pict. Telev. Engs*, vol. 66, pp. 597–601; October, 1957. Correction, *ibid.*, vol. 66, p. 775; December, 1957.) The necessity for correct control of black level (dc restoration) in television pictures of good fidelity is discussed. Advances in equipment design to achieve this control are described and a waveform standard is suggested for uniformity in black-level reproduction.

621.397.62 1277

Television Line Structure Suppression—F. T. Thompson. (*J. Soc. Mot. Pict. Telev. Engs*, vol. 66, pp. 602–606; October, 1957. Discussion, p. 606.) Methods of eliminating line structure are described. Experimental results obtained in comparing pictures with and without line structure suppression are given together with photographic comparisons.

621.397.62.001.4:535.623 1278

Flat-Field Generator speeds Colour TV Testing—R. W. Cook. (*Electronics*, vol. 30, pp. 139–141; December 1, 1957.) Description of a generator providing video signals of any hue or saturation as well as a luminance signal variable from black to white.

TRANSMISSION

621.396.61:621–519 1279

A New Transmitter for Long Distance Communications—V. Němeček. (*Slab. Obz., Praha*, vol. 18, pp. 15–18; January, 1957.) The transmitter, using dsb or ssb, operates in the 3–27.5-mc band at powers of 6–60 kw. Remote tuning to any of eight carrier frequencies is possible.

621.396.61:621.376.22 1280

Transmitter Circuits for Suppressed-Carrier A.M.—J. P. Costas and R. W. French. (*Electronics*, vol. 30, pp. 128–131; December 1, 1957.) Description and circuit details of a suppressed-carrier dsb system applied to two communication transmitters. They operate with peak sideband power 100 w and 1 kw, respectively, over a frequency range 2–32 mc, using class-C push-push power amplifiers with push-pull screen-grid modulation.

621.396.61:621.376.3 1281

New Trends in the Development of Large Transmitters—J. Vackář. (*Slab. Obz. Praha*, vol. 18, pp. 2–9; January, 1957.) A survey of trends in broadcasting, communication, television, and vhf transmitters is given with special reference to modulation methods.

TUBES AND THERMIONICS

621.314.63 1282

On the Switching Transient in the Forward Conduction of Semiconductor Diodes—H. L. Armstrong (IRE Trans. on Electron Devices, vol. 4, pp. 111–113; April, 1957. Abstract, Proc. IRE, vol. 45, p. 1163; August, 1957.) See also 2624 of 1957.

621.314.63:537.311.33 1283

On the Forward Characteristic of Semiconductor Diodes—H. L. Armstrong. (Proc. IRE, vol. 46, p. 361; January, 1958.) One expression, frequently quoted, for the current/voltage characteristic is said to be restricted to small voltages. Alternative formulas are given which fit experimental curves more closely.

621.314.7 1284

Transistor Cut-Off Frequency—W. L. Stephenson. (*Electronic Radio Eng.*, vol. 35, pp. 69–71; February, 1958.) The variation of cutoff frequency with collector voltage is calculated from known physical relations and provides a method for the measurement of the Early feedback factor. (See 874 of 1953.)

621.314.7:546.289 1285

Some Measurements on Commercial Transistors and their Relation to Theory—F. J. Hyde. (*Proc. IEE*, pt. B, vol. 105, pp. 45–52; January, 1958.) The effective lifetimes of minority carriers in the bases of four types of a Ge *p-n-p* alloy transistor and on one Ge surface-barrier transistor were measured. One-dimensional small-signal internal-transistor equations could be adapted to explain current-gain and cutoff-frequency parameters, provided that the bulk lifetime and diffusion constant of minority carriers in the base were replaced by effective values.

621.383.27:621.396.822:621.317.7 1286

Photomultiplier Tubes as Standard Noise Sources—E. R. Chenette, K. Shimada, and A. van der Ziel. (*Rev. Sci. Instr.*, vol. 28, pp. 835–836; October, 1957.) A brief article indicating that photomultipliers may be used to produce very large "equivalent saturated-diode currents" with negligible flicker effect down to 1 cps.

621.385.002.2 1287

Background, Foreground and Horizon—the Radio Valve Industry in Prospect and Retrospect—T. E. Goldup. (*Proc. IEE*, pt. B, vol. 105, pp. 1–10; January, 1958.)

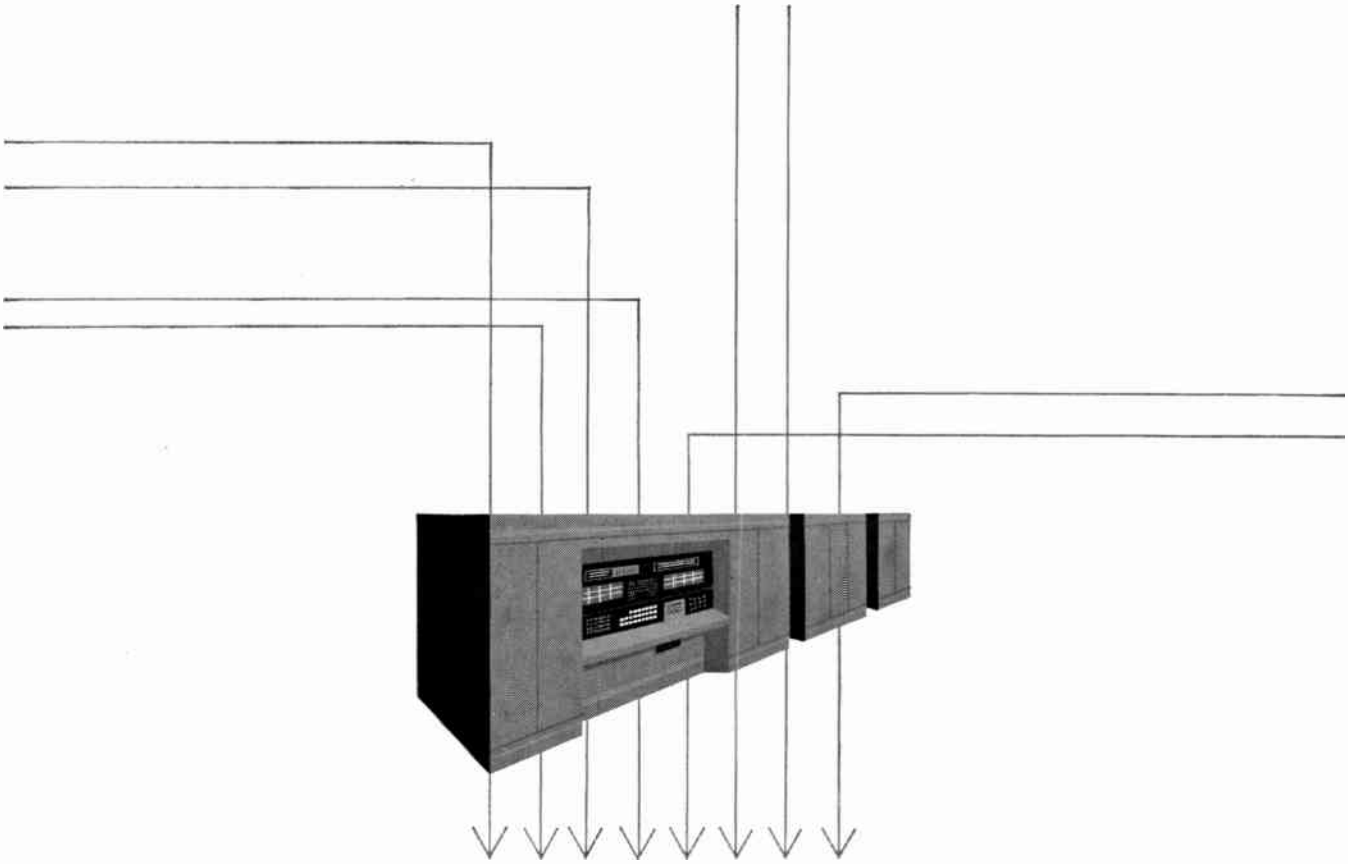
621.385.029.6:621.372.56 1288

Travelling-Wave-Tube Limiters—F. B. Fank and G. Wade. (IRE Trans. on Electron Devices, vol. 4, pp. 148–152; April, 1957. Abstract, Proc. IRE, vol. 45, p. 1164; August, 1957.)

MISCELLANEOUS

621.37/.39:061.3 1289

Radio Scientists Attend U.R.S.I. Conference—(*Tech. News Bull. Natl. Bur. Stand.*, vol. 41, pp. 187–189; December, 1957.) Brief summary of the proceedings of the 12th general assembly held in Boulder, Colo., from August 22 to September 5, 1957.



The **ORGANIZATION** and **RETRIEVAL** of **INFORMATION**

The organization and retrieval of large volumes of diverse types of information is rapidly becoming one of today's more serious problems. Major areas where the problem exists include business and industry, the military, the government, and the scientific and engineering community itself.

In its simpler forms, the problem may involve, for example, the automatic handling and analysis of business data such as payrolls, sales and manufacturing figures, insurance premiums. and other essentially statistical data. At the other extreme are certain complex military situations which require the concurrent interpretation, analysis, and integration on a very short time scale of data from a wide variety of sources, including field reports, photographs, news reports, estimates of industrial activity, and the like. In many of these situations, there is the additional requirement to translate the information from a foreign language into English.

The development in recent years of electronic data handling equipment is now making possible the practical solution of many of these problems. Such equipment has the capability to perform arithmetic operations, make decisions among alternatives, store and retrieve large quantities of information, and at high speed automatically perform long, complex sequences of operations.

At Ramo-Wooldridge, work is in progress on advanced information handling systems that are characterized by large volume and widely different forms of information, short time scales, and a variety of uses and users. The scope of the work includes the planning of systems and procedures, programming various types of data handling equipment, and formulation of requirements for new equipment. Research is also under way on the machine translation of foreign languages into English.

Engineers and scientists with experience in the following fields are invited to explore the wide range of openings now available:

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Computer Programming
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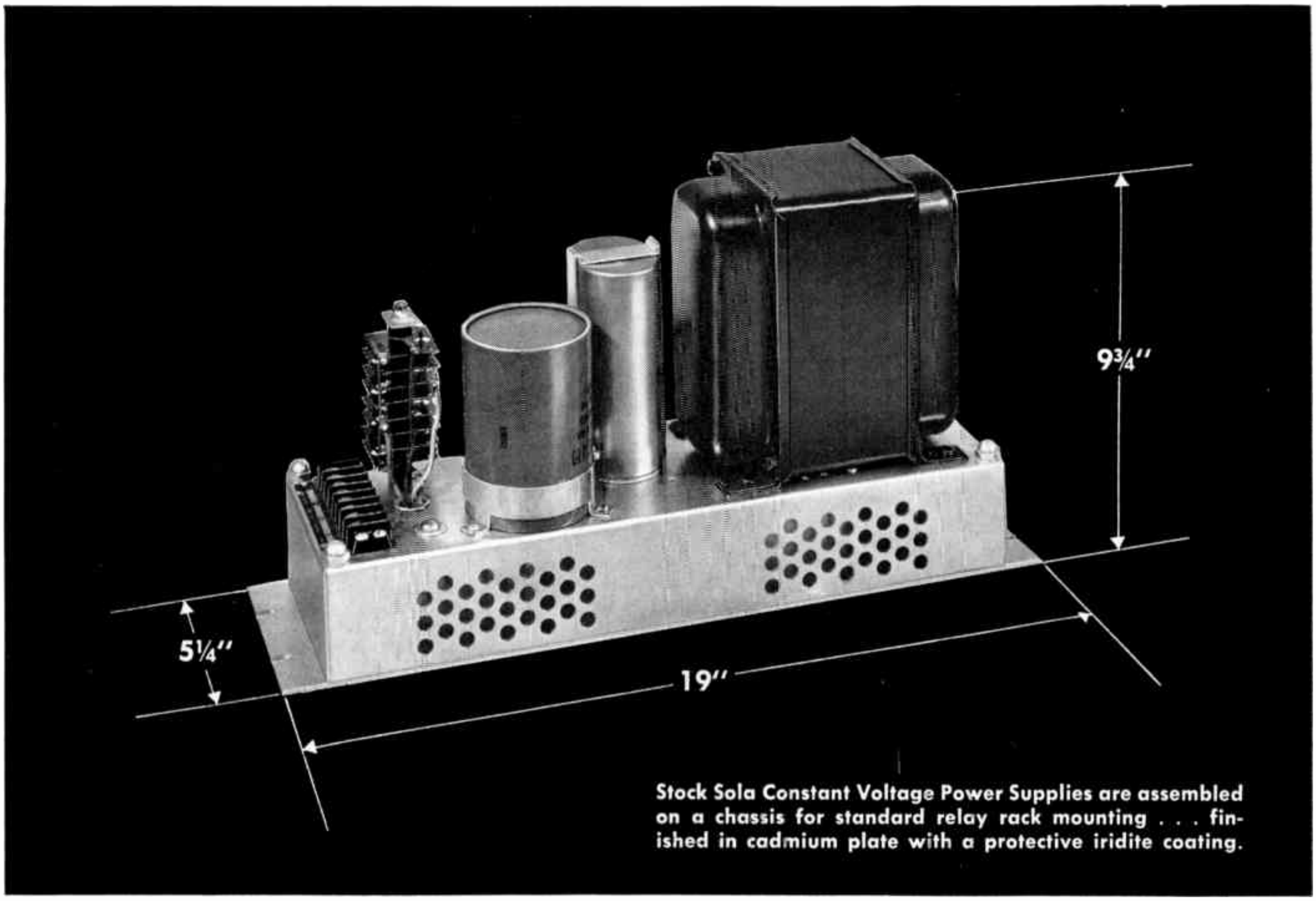

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Looking for a source of regulated dc power that fits into a little space? If so, then you'll find the Sola Constant Voltage DC Power Supply is just what you want.

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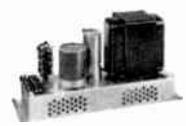
How's it done? Well, the answer is a unique combination of three components. A special Sola Constant Voltage Transformer is teamed up with a semiconductor rectifier and a high-capacitance filter. The Sola transformer's electrical characteristics maximize most of the advantages of the rectifier and filter, while virtually eliminating all their disadvantages. The resultant regulated dc power supply is simple, highly reliable, compact and moderately priced.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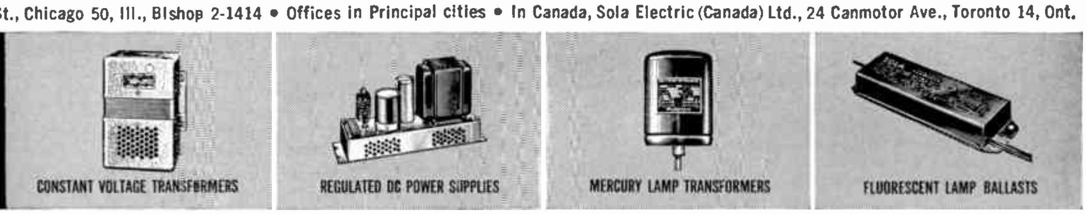


FREQUENCY STANDARDS

PRECISION FORK UNIT
TYPE 50

Size 1" dia. x 3¾" H. Wght., 4 oz.*

Frequencies: 240 to 1000 cycles
Accuracies:—
Type 50 (±.02% at —65° to 85°C)
Type R50 (±.002% at 15° to 35°C)
Double triode and 5 pigtail parts required
Input, Tube heater voltage and B voltage
Output, approx. 5V into 200,000 ohms

*3⅛" high
400 - 1000 cy.

FREQUENCY STANDARD
TYPE 50L

Size 3¾" x 4½" x 5½" High
Weight, 2 lbs.

Frequencies: 50, 60, 75 or 100 cycles
Accuracies:—
Type 50L (±.02% at —65° to 85°C)
Type R50L (±.002% at 15° to 35°C)
Output, 3V into 200,000 ohms
Input, 150 to 300V, B (6V at .6 amps.)

PRECISION FORK UNIT
TYPE 2003

Size 1½" dia. x 4½" H. Wght. 8 oz.*

Frequencies: 200 to 4000 cycles
Accuracies:—
Type 2003 (±.02% at —65° to 85°C)
Type R2003 (±.002% at 15° to 35°C)
Type W2003 (±.005% at —65° to 85°C)
Double triode and 5 pigtail parts required
Input and output same as Type 50, above

*3½" high
400 to 500 cy.
optional

FREQUENCY STANDARD
TYPE 2005

Size, 8" x 8" x 7¼" High
Weight, 14 lbs.

Frequencies: 50 to 400 cycles
(Specify)
Accuracy: ±.001% from 20° to 30°C
Output, 10 Watts at 115 Volts
Input, 115V. (50 to 400 cycles)

FREQUENCY STANDARD NEW
TYPE 2007-6
TRANSISTORIZED, Silicon Type

Size 1½" dia. x 3½" H. Wght. 7 ozs.

Frequencies: 400 — 500 or 1000 cycles
Accuracies:
2007-6 (± .02% at —50° to +85°C)
R2007-6 (±.002% at +15° to +35°C)
W2007-6 (±.005% at —65° to +125°C)
Input: 10 to 30 Volts, D. C., at 6 ma.
Output: Multitap, 75 to 100,000 ohms

FREQUENCY STANDARD
TYPE 2121A

Size
8¾" x 19" panel
Weight, 25 lbs.
Output: 115V
60 cycles, 10 Watt
Accuracy:
±.001% from 20° to 30°C
Input, 115V (50 to 400 cycles)

FREQUENCY STANDARD
TYPE 2001-2

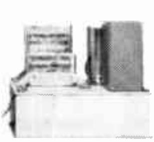
Size 3¾" x 4½" x 6" H., Wght. 26 oz.

Frequencies: 200 to 3000 cycles
Accuracy: ±.001% at 20° to 30°C
Output: 5V. at 250,000 ohms
Input: Heater voltage, 6.3 - 12 - 28
B voltage, 100 to 300 V., at 5 to 10 ma.

FREQUENCY STANDARD
TYPE 2111C

Size, with cover
10" x 17" x 9" H.
Panel model
10" x 19" x 8¾" H.
Weight, 25 lbs.
Frequencies: 50 to 1000 cycles
Accuracy: (±.002% at 15° to 35°C)
Output: 115V, 75W. Input: 115V, 50 to 75 cycles.

ACCESSORY UNITS
for TYPE 2001-2



L—For low frequencies multi-vibrator type, 40-200 cy.
D—For low frequencies counter type, 40-200 cy.
H—For high freqs, up to 20 KC.
M—Power Amplifier, 2W output.
P—Power supply.

This organization makes frequency standards within a range of 30 to 30,000 cycles. They are used extensively by aviation, industry, government departments, armed forces—where maximum accuracy and durability are required.

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DEW LINE'S VOICE

The Distant Early Warning Line needs more than advanced radar networks to do its job of guarding the northern reaches of the North American continent. It must be capable of sounding the alarm to continental defense centers unfailingly at any time should intruding aircraft be detected—even in the face of radio disturbances associated with the Arctic. A technological breakthrough in radio communication—Transhorizon scatter—provided the required reliability. Collins, a pioneer in Transhorizon, was given the assignment to supply the radio communication equipment by Western Electric Company, Inc., which directed the gigantic task of building the DEW Line.

UHF tropospheric scatter is used for lateral communication between DEW Line stations, tying together an advanced telephone system for coordinating the parts of the 3,000-mile electronic wall. VHF ionospheric scatter, with its longer range, provides rearward communication to defense centers. The equipment used on the DEW Line for conventional HF and ground-to-air VHF communication is also Collins.

DEW Line is another of the instances where Collins' engineering ability and experience have been called upon for advanced, reliable communication systems to serve in all parts of the world.

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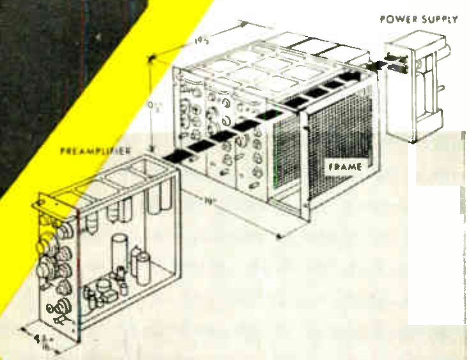
(Continued on page 104A)

greater **VERSATILITY . . . RELIABILITY . . . COMPACTNESS**

MODEL 450-1100 CARRIER PREAMPLIFIER

Carrier Freq. — 2400 cps (std.); 600, 1200, 4800 cps optional
Carrier Exc. — approx. 4.5–5 volts, depending on transducer imped.
Transducer Imped. — 100 ohms min. - 1000 ohms max.
Input Imped. — approx. 2500 ohms, incl. zero sup. ckt.
Sensitivity — 100 uv rms from transducer (output imped. 1000 ohms or less) gives 1 volt at output under max. output loading
Output — preferred circuit: between one active cathode and one reference cathode alternate circuit: between active cathode and ground
Output Voltage Capabilities — (a) ±3 volts into 2200 ohms min. load
(b) ±6 volts into 5000 ohms load
(c) ±7.5 volts open ckt
Output Linearity — better than 0.2% for (a) above
Output Impedance — approx. 1000 ohms, preferred ckt., 500 ohms alternate output ckt.
Freq. Response — 3 db at 20% of carrier freq.
Zero Suppression — can suppress 0 to 100% of transducer output (either sense via switch)
Power Req. — 115 volts, 50–400 cps, approx. 30 watts
All data subject to change without notice.



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. . . TAPE RECORDERS
. . . 'SCOPES

Here is amplifier design that gives you *wider* usefulness, *improved* performance and a *choice* of packaging, in reduced size and weight. Individually in portable cases, or as four-unit modules for standard 19" racks, with their own power supplies, Sanborn "450" Preamplifiers can be used with a variety of popular optical oscillographs, tape recorders, oscilloscopes or other indicating devices. The 450-1100 is a carrier amplifier-demodulator with calibrated zero suppression, for measuring strain, pressure, velocity, flow, temperature, displacement, etc. with a strain gage bridge, resistance or reactance transducer, or differential transformer. For differential input measurements, the 450-1800 provides the wide range of an AC amplifier and DC stability of a chopper, in a design featuring low noise, low drift and high gain. The 1800's usefulness is increased by a choice of models: *1800*, with high current output amplifier (±2.5 v, ±50 ma), position control, and zero suppression; *1800A*, same as 1800 but without zero suppression; *1800S*, with single-ended output amplifier, delivers ±2.5 volts at ±1 ma, linearity 0.1% full scale, 30 kc bandwidth, zero suppression and position controls; *1800SA*, same as 1800S except without zero suppression.

Investigate the many exclusive advantages these and other Sanborn "450" Preamplifiers offer your work. Call your local Sanborn Engineering Representative or write the Industrial Division in Waltham.

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Industrial Division
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MODEL 450-1800 TRUE DIFFERENTIAL DC PREAMPLIFIER

Input — Impedance: 200,000 ohms differentially between terminals (balanced) or 100,000 ohms each input lead to gnd. (single-ended)
Common mode rejection: at DC, 100 db; to 60 cps, 94 db; 400 cps, 80 db.
Equiv. input drift: ±2 uv for 24 hours
Equiv. input noise: 5 uv peak to peak (0–10 cps), 20 uv (0–1000 cps), 50 uv (0–30 kc)
Zero suppression: polarity pos. or neg. Ranges 0–100 mv and 0–1 volt. Accuracy 1% of full scale range.

Output — *Low Power Circuit:* (in all models)
Output appears between two cathodes as true push pull signal.
Common mode level of cathodes ±0.2 volts with respect to ground.
Output capability: ±3 volts into 5000 ohms
±10 volts open circuit
Zero position control not active for this output
Freq. response: 3 db down at 30 kc
Linearity: 0.1%
Gain: Fixed steps 1000, 500, 200, 100, 50, 20
Gain Accuracy 0.5% for D.C.
Smooth gain control covers range between fixed steps
High Power Circuit: (1800 and 1800A)
Output appears between two emitters as true push pull signal.
Common mode level of emitters ±2.5 volts with respect to ground.
Preferred load: 50 ohms
Output: ±2.5 volts, ±50 ma
Freq. response: 3 db down at 15 kc
Linearity: 0.5%
Zero position control is operative for high power output ckt.
Single-ended Output Circuit: (1800S & 1800SA)
Output appears as single-ended signal between emitter and amplifier chassis.
Load 2200 ohms, min.
Output ±2.5 volts
Freq. response: 3 db down at 30 kc
Linearity: 0.1%

Power Req. — 115 volts, 60 cps, approx. 50 watts

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(Continued from page 102A)

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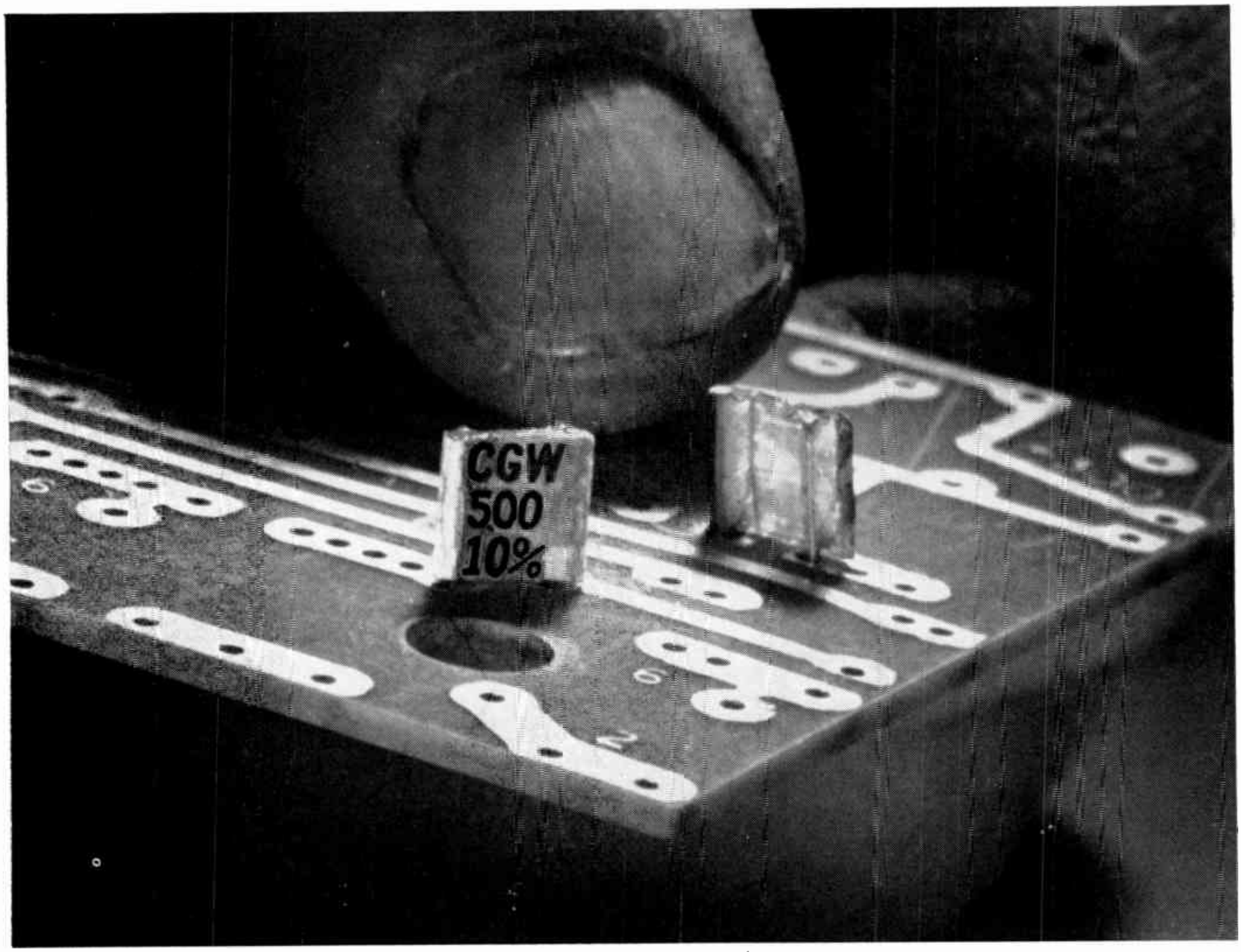
Schoenherr, H. W., Kingston, N. Y.

(Continued on page 106A)



Truly sub-miniature, these capacitors were devised especially for printed circuits and automatic assembly. Since they retain all the properties of larger, pig-tail capacitors, they are well suited to general circuitry as well.

Now—Corning Fixed Glass Capacitors in new sub-miniature size

Packing up to 1,000 uuf at 300 V. and 125°C. into 0.010 cubic inches, these new capacitors are designed for use on printed circuit boards and all applications requiring high-quality components. Advantages include fixed temperature coefficient, high insulation resistance, low dielectric absorption, the ability to operate under high humidity and high temperature conditions, plus the added advantage of increased miniaturization.

You can now up-grade your specs for miniature capacitors used on printed circuits.

These new capacitors measure only $\frac{9}{32}$ x $\frac{19}{64}$ x .115, yet have capacitances up to 1000 uuf at a full 300 V. rating at 125°C. Such exceptional thinness makes these capacitors particularly well suited for vertical mounting in small, high-rated units.

The capacitors have high temperature soldered leads which allow direct connection to circuit boards. The leads are .100 inches long, fitting most circuit board thicknesses and eliminating any trimming.

Reliable • Since the new construction is extremely simple, reliability is correspondingly high.

Rugged • These capacitors, when mounted, successfully withstand a standard five-hour vibration cycling test at 10 to 55 cycles, 15G Max.

Known as WL-4 capacitors, these units are in mass production. Your inquiries concerning data and prices are welcome.

FEATURES

1. to MIL C-11272A except smaller
2. 1 to 1,000 uuf
3. 300 volts
4. 125°C. full rating
5. .010 cubic inches

Corning means research in Glass

CORNING GLASS WORKS, 99-5 Crystal St., Corning, N.Y.

Electronic Components Department

Membership

(*Continued from page 104A*)

Schramm, C. T., Huntington Station, L. I., N. Y.
Schroeder, K. G., Alamogordo, N. M.
Schulze, G. H., Florissant, Mo.
Scott, D. P., Detroit, Mich.
Scroggins, E. B., Houston, Tex.
Selin, B. A., North Syracuse, N. Y.
Shadburn, T. H., Dayton, Ohio
Sharp, J. A., Medina, Ohio
Shashoua, F. E., Wilmington, Del.
Shaw, J. E., Oklahoma City, Okla.
Sherman, J. E., Palo Alto, Calif.
Shimizu, K., Los Angeles, Calif.
Shumway, J. L., San Francisco, Calif.
Simpson, J. R., Willowdale, Ont., Canada
Sinclaire, F. S., Brooklyn, N. Y.
Sivan, J. Rehovoth, Israel
Slear, L. S., Jr., Boalsburg, Pa.
Smith, D. S. J., Rochester, N. Y.
Smittenaar, G. H., Silver Spring, Md.
Snee, F. J., Torrance, Calif.
Sokol. S., New York, N. Y.
Spencer, L. V., Sarasota, Fla.
Spesock, G. J., Santa Clara, Calif.
Stevenson, K., Salisbury, Southern Rhodesia, Central African Federation
Stiltz, H. L., Marietta, Ga.
Stock, K. A., Harrisburg, Pa.
Stokes, L. C., Upland, Calif.
Strem, S. E., Claremont, Calif.
Sylvester, R. C., Holdrege, Nebr.
Taggart, H. E., Boulder, Colo.
Takiyama, K., Brooklyn, N. Y.
Tenney, J. E., Canoga Park, Calif.
Thomas, O. E., Dallas, Tex.
Thompson, A. R., Fort Davis, Tex.
Topping, J., Carlisle, Cumberland, England
Trost, H. K., Rochester, N. Y.
Tuchman, W. L., Kingston, N. Y.
Tucker, R. E., Columbus, Ohio
Turnbaugh, C. C., Indianapolis, Ind.
Vanous, P. J., North Hollywood, Calif.
Varekamp, J. F., Amsterdam, Holland
Vilipsons, J., Shipley Yorks, England
Walker, S. I., Oklahoma City, Okla.
Walsh, F. C., Fresh Meadows, L. I., N. Y.
Weaks, K. C., Pacoima, Calif.
Webb, H. R., Old Lyme, Conn.
Weber, S., Teaneck, N. J.
Weiland, E. H., Brunswick, Ga.
Whitney, C. A., Avon, Conn.
Wilkinson, S. A., Houston, Tex.
Willibald, G. E., Holloman AFB, N. Mex.
Winifield, R. J.. Gardena, Calif.
Winkler, T., Rochester, N. Y.
Winterstrom, R. L., Torrance, Calif.
Wittman, F. E., Los Alamos, N. Mex.
Wolf, J. E., Linthicum Heights, Md.
Wolfe, D. C., Elgin Air Force Base, Fla.
Wolfgang. W. W., Liberty, N. C.
Wolhowe, M. V., St. Charles, Ill.
Worch, A. B., San Pedro, Calif.
Worrall, J. W., Gardena, Calif.
Zalkin. S. P., New Hyde Park, L. I., N. Y.
Zaremba, J. G., Rolling Hills, Calif.

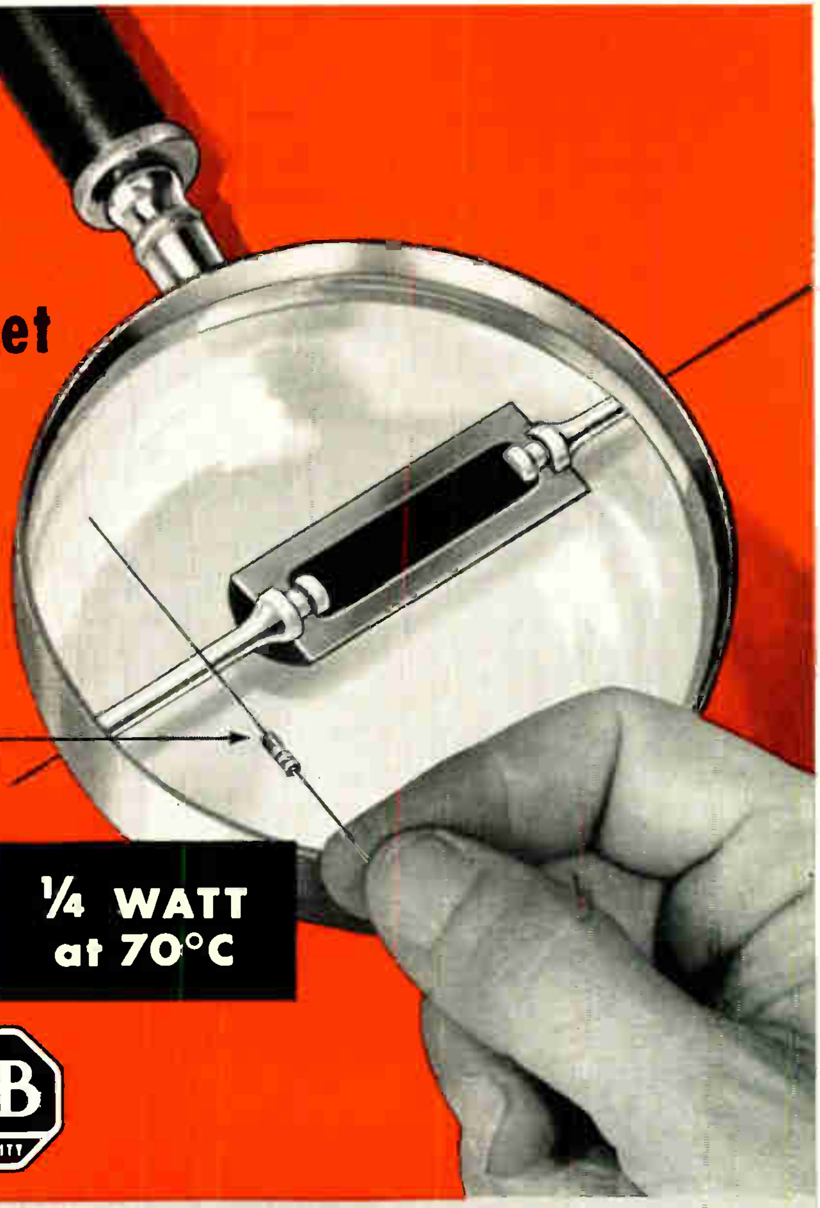
Admission to Associate

Allen, K. D., Dayton, Ohio
Baldwin, W. A., Sacramento, Calif.
Barcelo, P. O., Montreal, Que., Canada
Barlow, G. E., Washington, D. C.
Bascome, P. L., Tampa, Fla.
Becraft, C. E., Whippany, N. J.
Berenyi, S., Zarate, Argentina
Bowden, A. J., Amarillo, Tex.
Cohen, M. R., Paterson, N. Y.
Colbern, F. W., Houston, Tex.
Cole, P. J., Philadelphia, Pa.
Colligan, F. S., Washington, D. C.
Corvese, R. D., Anaheim, Calif.

(*Continued on page 108A*)

The same hot-molded insulating jacket as larger Allen-Bradley resistors

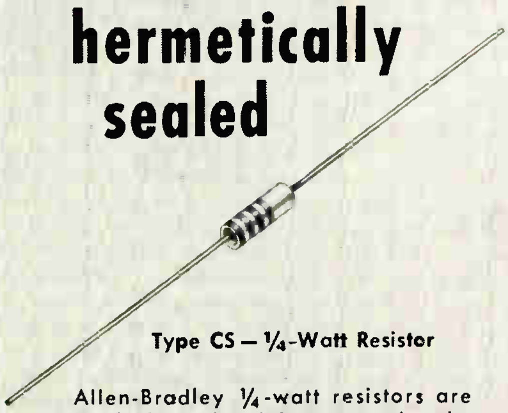
yet it's only ¼ inch long!



TYPE CB | **¼ WATT at 70°C**

A-B QUALITY

Also hermetically sealed



Type CS — ¼-Watt Resistor

Allen-Bradley ¼-watt resistors are available enclosed in a ceramic tube with high temperature end seals, making them impervious to humidity and moisture. Derated linearly from +70°C rating to 0 at +150°C. Available in 2% and 5% tolerances, and in resistance values from 47 ohms to 22 megohms.

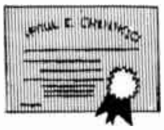
These ¼-watt composition resistors—ONLY ONE QUARTER OF AN INCH LONG—have the same hot-molded insulating jacket . . . the same reliability . . . the same physical uniformity . . . that have made the larger Allen-Bradley resistors the quality standard of the electronics industry for so many years!

Although exceptionally small, Allen-Bradley Type CB resistors are rated for continuous operation at 70°C ambient temperatures. The hot-molded insulating jacket of these resistors makes impregnation unnecessary . . . yet it provides the most reliable protection against extended periods of high humidity, as encountered in actual service. Available in all EIA resistance values from 47 ohms to 22 megohms. Tolerances: 5%, 10%, and 20%.

You can save space—with no sacrifice in performance or reliability—when you specify Allen-Bradley Type CB resistors. Write today for complete specifications.

Allen-Bradley Co., 1315 S. First St., Milwaukee 4, Wis.
In Canada: Allen-Bradley Canada Ltd., Galt, Ont.

ALLEN-BRADLEY
QUALITY
HOT-MOLDED COMPOSITION RESISTORS



Membership

(Continued from page 106A)

Crano, C. A., Akron, Ohio
Crosse, G. S., Fairborn, Ohio
Curtis, L. L., Dayton, Ohio
Davidson, J. R., St. Paul. Minn.
Des Lauriers, E. D., Hales Corners, Wis.
Dykes, W. E., Carrabelle, Fla.
Euwer, H. G., Van Nuys, Calif.
Floam, J. S., Baltimore, Md.
Friedmann, E., Saguenay, P. Q., Canada
Golder, R. L., West Chicago, Ill.
Goldstein, B. L., Baltimore, Md.
Gould, T. C., St. John's Newfoundland, Canada
Gregor, T. P., Williamsport, Pa.
Griffiths, J. D., Warren, Ohio
Grooms, W. G., Fairborn, Ohio
Hackenberger, R. B., Lexington, Mass.
Halland, L. J., Minneapolis, Minn.
Harding, P., Blawenburg, N. J.
Harrison, B. M., Montreal, Que., Canada
Henneke, J. T., New York, N. Y.
Hennessey, C. W., Norwood, Mass.
Housteau, C. M., Jr., Youngstown, Ohio
Hughes, R., Hounslow, Middlesex, England
Hullett, V. T., Hitchcock, Tex.
Jaeger, T. C., Zurich, Switzerland
Jaynes, F. H., Arlington, Va.
Johnson, A. B., Brentwood, Mo.
Kline, B. H., New York. N. Y.
Knowles, R. G., Alexandria, Va.
Kotze, J., Tampa, Fla.
Kreindel, B., Waltham, Mass.
Langstroth, R. W., Newton, N. J.
Langstroth, T. A., Newton, N. J.
Larson, J. E., Port Washington, L. I., N. Y.
Lazaro, S. A., Stockton, Calif.
Liiv, H. H., Utica, N. Y.
Malecki, A. S., Milford, Conn.
Mangels, L. A.. Hohokus, N. J.
Maranhao, S. M., Rio de Janeiro, Brazil
Marquess, R. L., Houston, Tex.
McArdle, O. L., Fort Tilden, L. I., N. Y.
McCreary, J. T., El Paso, Tex.
McDade, H. J., Waban, Mass.
McElyea, W. W., Muskegon, Mich.
Meehan, D. J., Lawrence, Mass.
Mehrwin, R. L., Fort Worth, Tex.
Meidell, P. I., Los Gatos, Calif.
Miller, E. C., Calgary, Alta.. Canada
Molnar, J. T., Boonton, N. J.
Moore, C. R., South Bend, Ind.
Murphy, J. L., Elko, Nev.
Nance, D. K., Flint, Mich.
Nelsom, M. W., Elverson, Pa.
Nolan, W., Needham Heights, Mass.
Norman, E. R., Vancouver, B. C., Canada
Ogawa, H. F., Levittown, Pa.
Olah, J., Bethpage, L. I., N. Y.
Opheim, L. E., Syracuse, N. Y.
O'Shea, J. P., Pittsburgh, Pa.
Osifchin, N., Clifton, N. J.
Ozolins. I., Baldwin, L. I., N. Y.
Palmer, E. F., Rhinebeck, N. Y.
Paradis, K. A., New York, N. Y.
Pauley, T. N., Mountain View, Calif.
Pernula, M. J., Minneapolis, Minn.
Peterson, K. F., Sunnyvale, Calif.
Polansky, P., New Rochelle, N. Y.
Pelton, E. A., St. Petersburg, Fla.
Pouliot, L. J., Fayville, Mass.
Powers, R. E., Lawrence, Mass.
Pratt, R. E., Chicago, Ill.
Price, E. M., Vancouver, B. C., Canada
Richardson, F. A., Florham Park, N. J.
Roberts. V., Port Angeles, Wash.
Roehrig, H. W., New York, N. Y.
Rose, L. L., Chicago, Ill.
Russell. R. A., Wappingers Falls, N. Y.
Schlachter, H. J., Detroit, Mich.
Shawda, F. E., Dayton, Ohio
Sibley, R. C., St. Petersburg, Fla.
Siekman, W. J., Jr., Batavia, Ill.
Silverman, B. S., Fort Huachuca, Ariz.
Skowron, A. J., Chicago, Ill.
Smith, R. M. Collegeville, Pa.
Stevens, A. W., Birchrunville, Pa.
Stevens, D. R., South Bend, Ind.
Stone, G. F., New York, N. Y.
Stonog, J. W., Syracuse, N. Y.
Stoy, K. E., Jr., Granada Hills, Calif.
Swaminathan, K., Chromepet, Madras, India
Tenney, F. T., Tulsa, Okla.
Tepper, C. A., Long Island City, L. I., N. Y.
Thesing, R. H., Ashland, Mass.
Tomovic, R. M., Belgrade, Yugoslavia
Upton, F. R., Bisbee, Ariz.
Urban, A., Pompton Plains, N. J.
Vecchia, E. F., Revere, Mass.
Warner, H. R., Salt Lake City, Utah
Watt, C. M., Willowdale, Ont., Canada
Wedel, J. O., Jr., Timonium, Md.
Williamson, J. V., Merchantville, N. J.
Wilson, G. M., Boston, Mass.
Yokomoto, R. K., Orlando, Fla.

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Ask your ARC dealer for a quotation on the Type 21A ADF and on other ARC equipment listed below.

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10-CHANNEL ISOLATION AMPLIFIERS • OMNIRANGE SIGNAL GENERATORS AND STANDARD COURSE CHECKERS • 900-2100 MC SIGNAL GENERATORS

Professional Group Meetings

(Continued from page 110A)

Cincinnati—January 21

"Development of Stereophonic Sound," Jack Drews, B. C. Carr, Ampex Audio, Inc.

Philadelphia—February 12

"Models of Auditory Processes," J. C. R. Lichlider, Bolt, Beranek and Newman.

Philadelphia—January 15

"The Westrex Sterodisk System," R. E. Warn and E. A. Dickinson, Westrex Co.

San Diego—January 31

Lecture and Tour sponsored jointly by the San Diego Section of Audio and the San Diego Section of the Acoustical Society of America on Broadcast Type Television Studio at San Diego State College, by Kenneth Jones, Assoc. Prof. of Speech Arts.

Washington, D. C.—December 17

"High Fidelity Reproduction of What," M. A. Kerr, Melpar, Inc.

Automatic Control

Baltimore—January 29

"Automatic Controls for an Unattended Radar," K. F. Molz, Bendix Radio.

Dallas—Fort Worth—February 4

"Problems and Progress in Computer Process Control," T. M. Stout, Ramo-Wooldridge Corp.

Broadcast & Television Receivers

Los Angeles—January 30

"Progress Review of New Developments in Commercial Transistors," J. W. Peterson, Pacific Semiconductors.

Broadcast Transmission Systems

Boston—January 30

"The Video Tape Recorder," Harold Bresson, Ampex Corp.

Florida West Coast—January 8

"New Facilities at WTVT," William Witt, WTVT.

Communications Systems

Florida West Coast—February 5

"Frequency Synthesis," Harold Cohen, Electronic Communications, Inc.

Philadelphia—January 30

"Electronic Design and the Operator," B. J. Smith, Gen. Elec. Co.

(Continued on page 115A)

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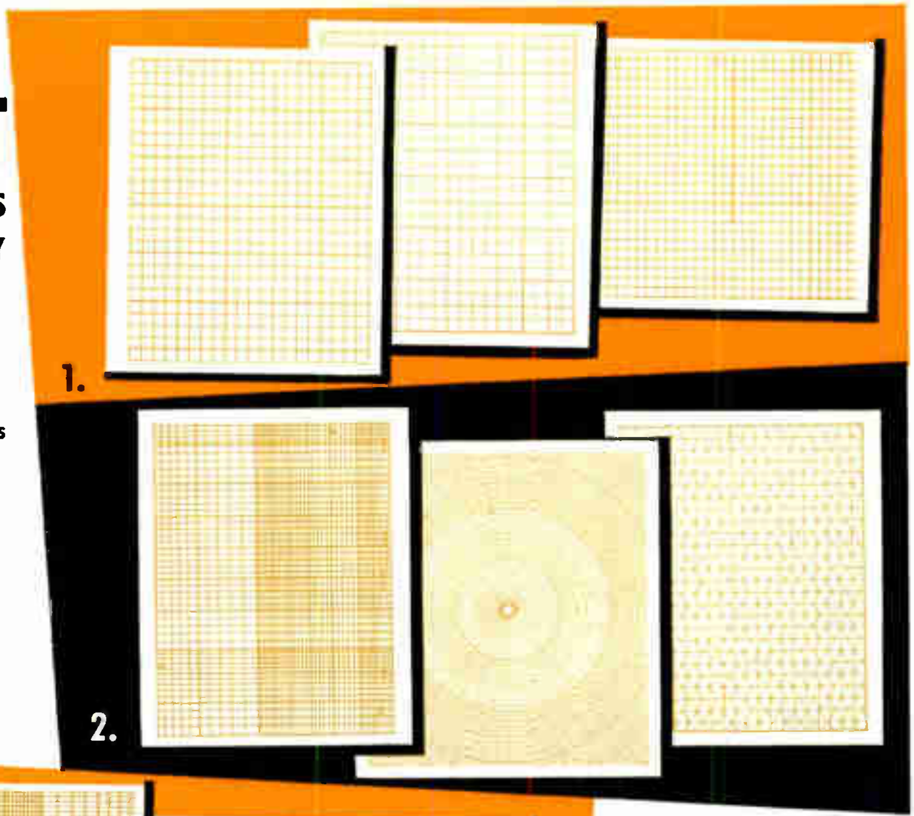

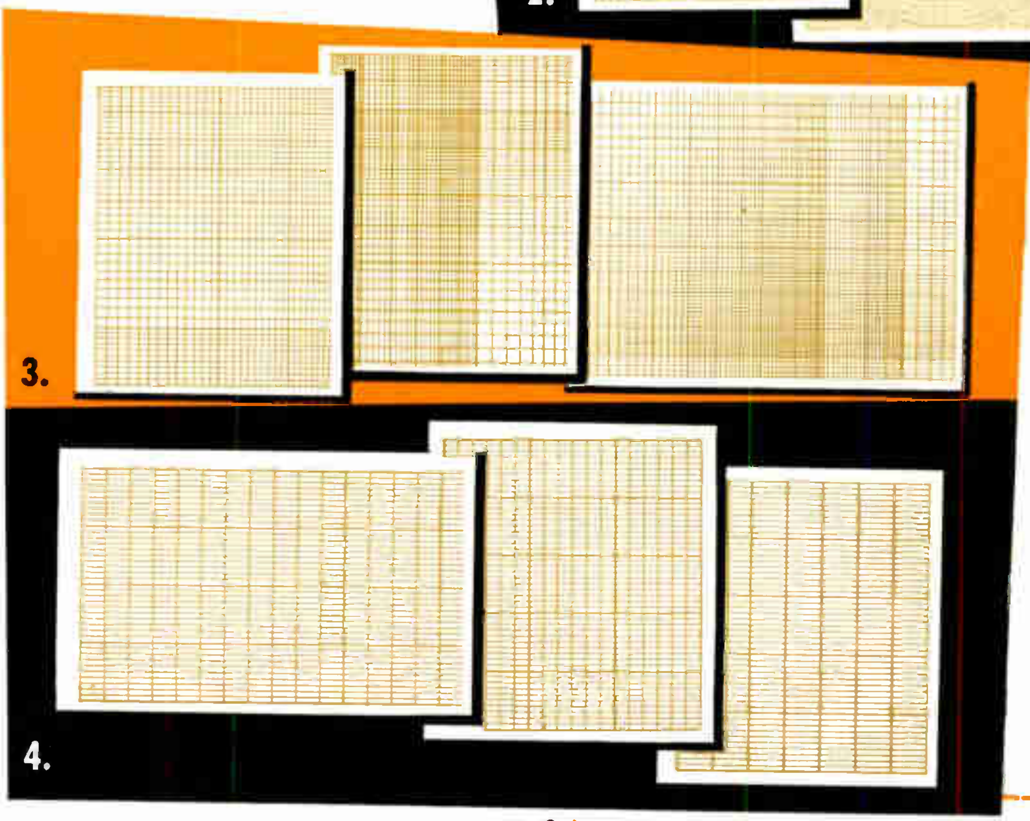

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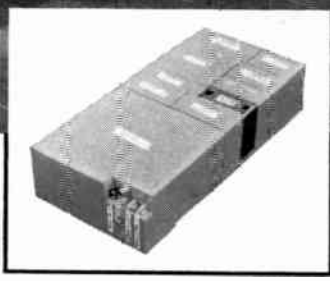

NAVTAC: "Pipeline" to a happy landing

The uniqueness of the new NAVTAC en route navigation and instrument landing system by Stromberg-Carlson is in its *combination* of functional modules.

The NAVTAC equipment is an assembly designed to provide high-performance aircraft with the TACAN navigational aid, plus marker beacon receiver, glide slope and runway localizer for instrument landing situations.

The entire system is packaged in a compact unit only 5" high, 10½" wide, 22" deep, and weighing only 47.5 lbs. Individual modules can be separated up to distances of several feet without any adverse effect on performance.

The equipment is designed to meet the rigorous environment of the high-performance aircraft of today and tomorrow. Its operating ambient temperature range is −60 to +125 degrees C. at altitudes up to 70,000 feet. Widespread use of semiconductors in the ILS receivers and TACAN circuitry means high reliability, small size and low power consumption.

Included in the design is the capability of performing complete preflight confidence tests with the use of a small auxiliary test set.

Complete technical details on the NAVTAC system are available on request.

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A DIVISION OF GENERAL DYNAMICS CORPORATION
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Electronic and communication products for home, industry and defense

Professional Group Meetings

(Continued from page 112A)

San Francisco—January 7

"The Moon(s) as a Passive Reflector for Long Distance Communications," A. M. Peterson, Stanford University; R. L. Leadabrand, Stanford Research Institute; R. B. Dyce, Stanford Research Institute.

Washington—February 3

"Meteor Burst Communication," A. M. Peterson, Stanford Research Institute.

Component Parts

Baltimore—February 25

"The Department of Defense Looks at Reliability," R. H. DeWitt, Office of Electronics, OASD.

Baltimore—January 28

"Evaporative Cooling Techniques Applied to Electronic Equipment," Peter Zukauskas, The Martin Co.

Baltimore—December 17

"The Paraballoon Antenna—A Revolutionary Air-Supported Antenna," Stan Saulson, Westinghouse Electronics Division; "Application of Choppers," Don Holdt, Airpax Products Corp.

Philadelphia—February 11

"The Astramatic Test System," Eugene Hoo, Electronic Control Systems, Inc.

Electron Devices

Los Angeles—February 17

"Microwave Interaction with Plasmas," Roy Gould, California Institute of Technology.

Los Angeles—January 20

"Voltage-Sensitive Capacitors Employing Semiconductors," Morgan McMahon, Pacific Semiconductors, Inc.

Washington—January 20

"Voltage Tunable Magnetrons," M. Weinstein, General Electric Co.

Electronic Computers

Akron—February 18

"Documentation Retrieval," J. W. Perry, Western Reserve University.

Baltimore—January 22

"Logical Design for the SAGE Input Monitor," Byron Bair, Bendix Radio Div.

Boston—December 3

"Characteristics of Airborne Digital Computers," Adolph Baker, RCA Airborne Systems Lab.

(Continued on page 118A)

18 Years of Progress at E-I

1940 . . . specialists in glass to metal seals for over 18 years. Research, development and manufacture devoted exclusively to the production of sealed terminations.

1951 . . . pioneered the production of compression type glass-to-metal seals. E-I developed and made a practical reality of the first compression seal, now the standard of the industry.

1957 . . . expanded research paces industry's varied needs. Universal acceptance of super-durable E-I hermetic seals made necessary expansion of research, engineering and production facilities.

1958

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GLASS-TO-METAL SEALS*

E-I glass-to-metal seals are available in hundreds of economical types of standardized designs. The complete line ranges from single lead terminals to miniature closures and color coded terminals. Specify E-I super-rugged compression type seals for critical military and commercial applications. Save time and eliminate custom sealing costs by using application-proven, standardized seals that meet practically every hermetic sealing requirement. Complete engineering data on the E-I Standard Line is contained in one helpful handbook. Write for your copy now.

Select your seals from the complete E-I Standard Line . . .

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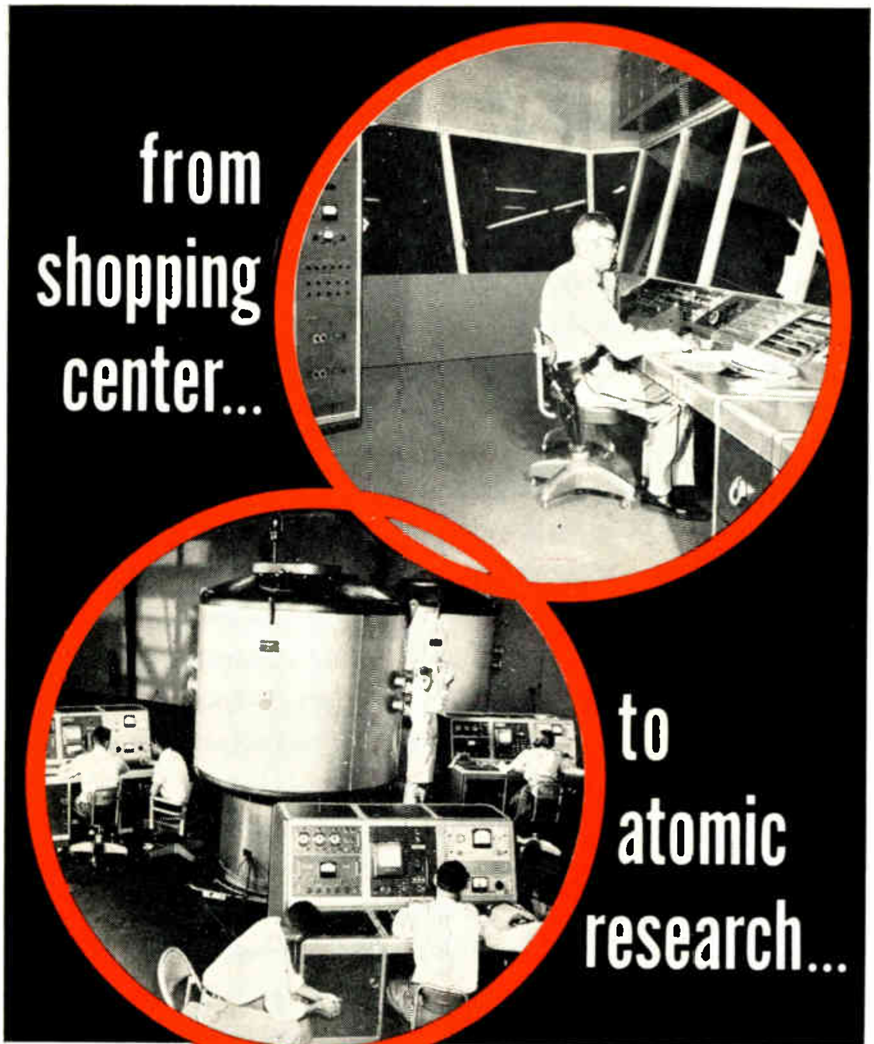


Professional Group Meetings

(Continued from page 115A)

Philadelphia—March 11

"Sampled Data Analog Computer," G. Stubbs & E. Greenberg, Franklin Inst.

Philadelphia—February 11

"A High Speed Page Printer," Epstein-Harris, Burroughs.

Philadelphia—January 14

"RCA BIZMAC," R. A. C. Lane, Radio Corp. of America.

Pittsburgh—March 5

"Bits and Pieces," A. L. Samuel, International Business Machines.

Pittsburgh—January 22

"The Techniques of Compiler Construction," (Panel Discussion), Harold Stern, IBM Programming Research; and A. J. Perlis, Carnegie Institute of Technology.

San Francisco—December 3

"SAGE System," Brig. Gen. S. T. Wray, USAF.

Twin Cities—February 26

"Reliability of Digital Computer Process Control Systems," W. S. Aiken, Ramo-Wooldridge Corp.

Washington, D. C.—February 5

"Logical Design of the Datatron 220," T. Withington, Burroughs Corp.

Washington, D. C.—January 14

"Using Computers to Extract Information Content," S. N. Alexander and R. B. Thomas, U. S. Govt., Natl. Bureau of Standards.

Engineering Management

Boston—February 13

"The Engineering—Administration—Management Triangle," R. W. Couch, Sylvania.

Los Angeles—February 18

"Russian Education," W. P. Lear, Lear, Inc.

Los Angeles—January 17

"Relationships between Prime and Sub-Contractors and their Suppliers," Adm. C. F. Horne, Convair, Pomona Div.

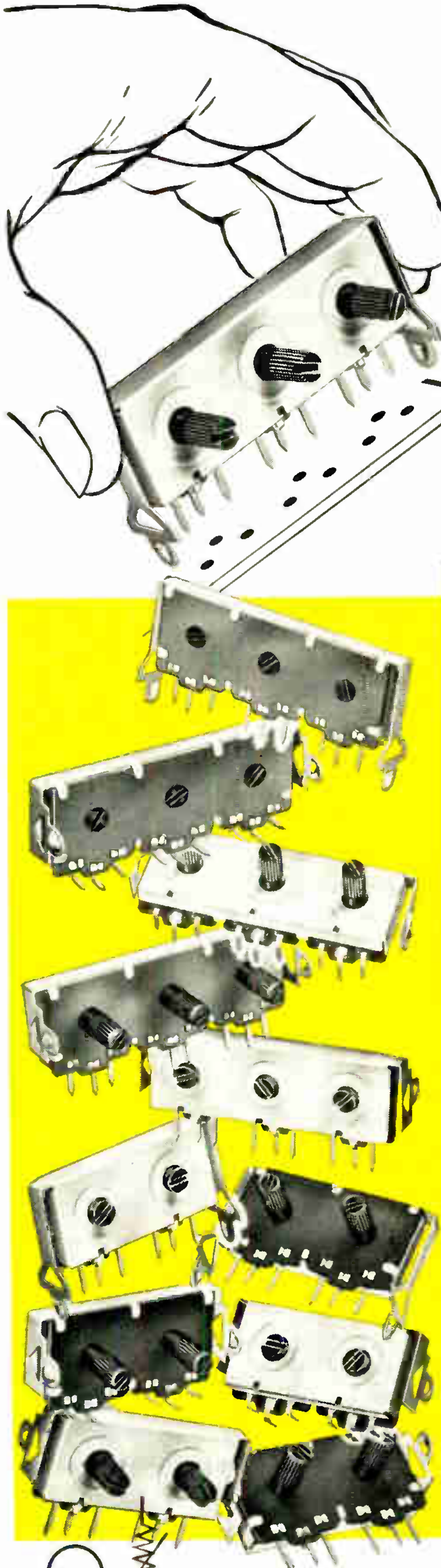
Metropolitan New York—February 13

"Use of Computers for Engineering Development and Design," W. H. McWilliams, Bell Tel. Labs.

Philadelphia—January 16

"Administration of a Patent Program," John Sims, Remington Rand Univac.

(Continued on page 120A)



12 New
COST-SAVING
Side-By-Side Controls

Proportioned for today's compact chassis.

Designed for simplified production assembly.

SIDE-BY-SIDE MOUNTING in twelve dual- and triple-unit models.

POSITIVE-LOCKING, PLUG-IN BRACKET in styles for chassis and panel mounting.

SLOTTED PHENOLIC SHAFTS in two styles, both adjustable from either side of control.

17 STANDARD SHAFT LENGTHS to meet many mounting requirements.

SPACE-SAVING DESIGN allows wires and cables to run underneath controls.

PRINTED CIRCUIT TERMINALS parallel or perpendicular to shafts as needed.

Here's the new Stackpole multiple-control design that lets you mount a single unit equivalent to 2 or 3 control sections in less space . . . in far less time—and with perfect alignment between chassis and panel.

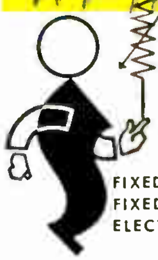
Proportioned for today's compact TV receivers, and other miniaturized equipment, Stackpole Side-by-Side Controls are now available in production quantities in 12 dual- and triple-unit styles rated at 0.75-watt in values up to 10K ohms, 0.5-watt above 10K.

Write for details. Samples gladly sent to quantity users.

STACKPOLE

VARIABLE composition **RESISTORS**

Electronic Components Division



STACKPOLE CARBON COMPANY • St. Marys, Pa.

In Canada: Canadian Stackpole Ltd., 550 Evans Ave., Etobicoke, Toronto 14, Ont.

FIXED AND VARIABLE COMPOSITION RESISTORS • SNAP AND SLIDE SWITCHES • CERAMAG® FERROMAGNETIC CORES FIXED COMPOSITION CAPACITORS • IRON CORES • CERAMAGNET® CERAMIC MAGNETS • BRUSHES FOR ALL ROTATING ELECTRICAL EQUIPMENT • ELECTRICAL CONTACTS • AND HUNDREDS OF RELATED CARBON, GRAPHITE AND METAL POWDER PRODUCTS.

Professional Group Meetings

(*Continued from page 118A*)

Rome–Utica—March 4

"The Responsibility of Design Engineers," M. C. Batsel, Radio Corp. of America.

Rome–Utica—January 21

"Communications in an Engineering Organization, Up, Down and Laterally," M. V. Ratynski, Moderator (panel discussion).

San Francisco—January 8

"Engineering Organization at Lockheed," K. T. Larkin, Moderator (panel discussion).

San Francisco—November 14

"Starting a New Electronics Firm—Some Pitfalls and Rewards," J. V. N. Granger, Granger Associates.

Syracuse—January 16

"Experiences in Establishing a Southern Plant Operation," A. F. Persons, General Electric Co.

Industrial Electronics

Chicago—December 13

"Electronics & Banking," J. M. Deterding, Armour Research Foundation.

Omaha-Lincoln—February 14

"Closed Circuit Television for Education and Industry," John Almen, Radio Corp. of America.

Information Theory

Washington—January 14

"Using Computers to Extract Information Content," S. N. Alexander, National Bureau of Standards.

Instrumentation

Long Island—January 21

"Infra-Red Instrumentation," Frank Willey, Servo Corp. of America.

Washington, D. C.—January 6

"Satellite Instrumentation of the Future," J. P. Hagen, Naval Res. Lab.

Medical Electronics

Chicago—January 10

"Internal Sample Liquid Scintillation Counting," S. M. Bristol, Packard Instrument Co.

Los Angeles—February 20

"Studies in Fetal Electrocardiography," Kanak Dasgupta, U.C.L.A.; "An Integrated Electronic System for Multiple Physiologic Recordings," H. M. Hanish, Mt. Sinai Hospital.

(*Continued on page 122A*)

How BUSS fuses can help you solve electrical protection problems

BUSS offers the widest possible selection of fuses to meet your exact electrical protection requirements. The complete BUSS line of fuses includes: single-element, quick-blowing type; single-element for normal circuit protection type; dual-element, slow blowing type and signal or visual indicating types . . . ranging in sizes from 1/500 ampere up — plus a companion line of fuse clips, blocks and holders.

ELECTRONIC TESTING ASSURES DEPENDABILITY: With BUSS fuses, dependable electrical protection is not left to chance. Every BUSS fuse is tested in a sensitive electronic device that automatically rejects any fuse not correctly calibrated, properly constructed and right in all physical dimensions.

By operating as intended, BUSS fuses provide maximum protection against damage due to electrical fault and—BUSS fuses eliminate shutdowns caused by needless blows. Hence, BUSS fuses help the good name of your product for service and reliability.

FUSE ENGINEERING HELP: If you have a special problem in electrical protection, BUSS engineers are at your service — and in many cases can save you engineering time by helping you choose the right fuse for the job. Whenever possible, the fuse selected will be available in local wholesalers' stocks that your device can be serviced easily.

For more information on the complete line of BUSS and FUSETRON Small Dimension Fuses and Fuseholders, write for bulletin SFB.

Bussmann Mfg. Division McGraw-Edison Co., University at Jefferson, St. Louis 7, Mo.

BUSS fuses are made to protect . . . not to blow, needlessly

FUSETRON
TRUSTWORTHY NAMES IN ELECTRICAL PROTECTION
BUSS

A complete line of fuses for home, farm, commercial, electronic, automotive and industrial use. 558

Professional Group Meetings

(Continued from page 120A)

Los Angeles—January 16

"Electronics in Radiology," Arthur Fisher, Mt. Sinai Hospital; "Electronics in Medical Use of Radioisotopes," W. H. Blahd, Veterans Adm. Hosp.

Washington, D. C.—February 6

"Instrumentation for the Anaesthetist," John Severinghaus, Natl. Inst. of Health.

Microwave Theory & Techniques

Chicago—January 10

"Cooking With Microwaves," E. L. Macoicz, Hotpoint Co.

Long Island—February 18

"Survey of Recent Millimeter Wave Techniques," A. H. Nethercot, Jr., I.B.M. Watson Lab.

Los Angeles—February 13

"Recent Advancements in Microwave Ferrites," T. N. Anderson, Airtron, Inc.; "Radio Astronomy," J. G. Bolton, California Institute of Technology.

Los Angeles—December 12

"The Paramagnetic Amplifier—A New Solid-State Microwave Device," Hubert Heffner, Stanford University; "Characteristics of Microwave Power Tubes and Systems," J. E. Gerling, Litton, Ind.

New York—January 21

"Ferrites and Microwaves," Howard Boyet, Radio Corp. of America.

New York—January 14

"Fundamentals of Ferromagnetism," J. H. Rowen, Bell Tel. Labs.

New York—January 8

"New Frontiers in Radio Astronomy," Harold Ewen, Ewen Knight Corp.

Northern New Jersey—February 19

"Recent Advances in Microwave Ferrites," T. N. Anderson, Airtron, Inc.

Northern New Jersey—January 15

"Will Success Spoil the Traveling Wave Tube?" W. R. Beam, Radio Corp. of America.

Northern New Jersey—December 11

"Ferrite MAVAR," M. T. Weiss, Bell Telephone Labs.

Syracuse—January 9

"Construction of Large Antennas," C. W. Creaser, D. S. Kennedy & Co.

(Continued on page 136A)

Three voltage ranges:
0-200, 125-325, 325-525 VDC

1.5 AMPERE MODELS NEED ONLY 8¾" OF PANEL HEIGHT!

(metered)	(unmetered)
MODEL C-1580M: 0-200 VDC, 0-1500 MA.....580.00	MODEL C-1580: 0-200 VDC, 0-1500 MA.....550.00
MODEL C-1581M: 125-325 VDC, 0-1500 MA....605.00	MODEL C-1581: 125-325 VDC, 0-1500 MA.....575.00
MODEL C-1582M: 325-525 VDC, 0-1500 MA....680.00	MODEL C-1582: 325-525 VDC, 0-1500 MA.....650.00

800 MA MODELS NEED ONLY 7" OF PANEL HEIGHT!

(metered)	(unmetered)
MODEL C-880M: 0-200 VDC, 0-800 MA.....370.00	MODEL C-880: 0-200 VDC, 0-800 MA.........340.00
MODEL C-881M: 125-325 VDC, 0-800 MA.....345.00	MODEL C-881: 125-325 VDC, 0-800 MA.........315.00
MODEL C-882M: 325-525 VDC, 0-800 MA.....390.00	MODEL C-882: 325-525 VDC, 0-800 MA.........360.00

400 MA MODELS NEED ONLY 5¼" OF PANEL HEIGHT!

(metered)	(unmetered)
MODEL C-480M: 0-200 VDC, 0-400 MA......289.50	MODEL C-480: 0-200 VDC, 0-400 MA.........259.50
MODEL C-481M: 125-325 VDC, 0-400 MA......274.50	MODEL C-481: 125-325 VDC, 0-400 MA........244.50
MODEL C-482M: 325-525 VDC, 0-400 MA......289.50	MODEL C-482: 325-525 VDC, 0-400 MA.........259.50

200 MA MODELS NEED ONLY 5¼" OF PANEL HEIGHT!

(metered)	(unmetered)
MODEL C-280M: 0-200 VDC, 0-200 MA......214.50	MODEL C-280: 0-200 VDC, 0-200 MA.........184.50
MODEL C-281M: 125-325 VDC, 0-200 MA......189.50	MODEL C-281: 125-325 VDC, 0-200 MA.........159.50
MODEL C-282M: 325-525 VDC, 0-200 MA......199.50	MODEL C-282: 325-525 VDC, 0-200 MA.........169.50

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POWER SUPPLIES

Less space! Improved performance!

Long, trouble-free service!

Transient free output!

Fills the need for compact, regulated DC power supplies. Economy of panel space, functional simplicity, new quick-service features.

Wiring, tubes and other components readily accessible. You can reach them easily, service them fast.

400 MA, 800 MA, and 1.5 ampere models include new, high-efficiency, long-life, hermetically-sealed semi-conductor rectifiers. All Com-Pak models are constructed with hermetically-sealed transformers, chokes and capacitors.

Condensed Data

LINE REGULATION . . . . Better than 0.15% or 0.3 Volt, whichever is greater.

LOAD REGULATION . . . . . . .Better than 0.25% or 0.5 Volt, whichever is greater.

INTERNAL IMPEDANCE

C- 200 Series . . . . .Less than 6 ohms.
C- 400 Series . . . . .Less than 3 ohms.
C- 800 Series . . . . .Less than 1.5 ohms.
C-1500 Series . . . . .Less than 0.75 ohms.

RIPPLE AND NOISE. . . . . . . .Less than 3 millivolts rms.

POLARITY . . . . .Either positive or negative may be grounded.

AMBIENT TEMPERATURE. . . .Continuous duty at full load up to 50°C (122°F) ambient.

AC OUTPUT (unregulated) . . . 6.5 VAC (at 115 VAC Input).

C- 200 Series . . . . . 10 AMP
C- 400 Series . . . . . 15 AMP
C- 800 Series . . . . 20 AMP
C-1500 Series . . . . .30 AMP

AC INPUT . . . . 105-125 VAC, 50-400 CPS

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11-11 131st Street, College Point 56, New York

By return mail, send complete specifications on Lambda Com-Pak Power Supplies.

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Company__________

Address__________

City__________ Zone____ State__________

28 Fields of Special Interest-

The 28 Professional Groups are listed below, together with a brief definition of each, the name of

Aeronautical and Navigational Electronics

Annual publications fee: $2.

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

Mr. Joseph General, 6019 Highgate Drive, Baltimore 15, Md.

25 Transactions, 5 Newsletters, *5, *6, *8, & *9, and *Vol. ANE-1, Nos. 1, 2, 3 and 4. Vol. 2, No. 1-4; Vol 3, No. 1-2-3-4. Vol. 4, No. 1, 2, 3, 4.

Antennas and Propagation

Annual publications fee: $4.

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

Dr. John I. Bohnert, Chairman, Naval Research Lab., Washington 25, D.C.

20 Transactions, 1 Newsletter, *4, *Vol. AP-1, Nos. 1, 2; *Vol. AP-2, Nos. 1-4, AP-3, No. 1-3; AP-4, No. 1-2-3. AP-5-1-4. AP-6-1.

Audio

Annual publications fee: $2.

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

Mr. F. H. Slaymaker, Chairman, Stromberg-Carlson, Rochester, N.Y.

39 Transactions, 4 Newsletters, *5, *7, *10. *Vol. AU-1, Nos. 1-6; *Vol. AU-2, Nos. 1-6; Vol. AU-3, Nos. 1-6; Vol. AU-4, No. 1-2-3-4-5-6. Vol. AU-5, No. 1-5.

Automatic Control

Annual publications fee: $2.

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

Mr. E. M. Grabbe, Chairman, Ramo-Wooldridge Corp., Los Angeles 45, Calif.

4 Transactions, PGAC-1-2-3-4.

Broadcast & Television Receivers

Annual publications fee: $2.

The design and manufacture of broadcast and television receivers and components and activities related thereto.

Mr. Harland A. Bass, Chairman, Avco Mfg. Corp., Arlington, Cincinnati, Ohio.

19 Transactions, *1, *2, *3, *5, *6, *7, 8; BTR-1, No. 1-4. BTR-2, No. 1-2-3. BTR-3, No. 1-2. BTR-4, No. 1-2.

Broadcasting Transmission Systems

Annual publications fee: $2.

Broadcast transmission systems engineering, including the design and utilization of broadcast equipment.

Mr. Clure H. Owen, Chairman, American Broadcasting Co., 7 West 66th St., New York 23, N.Y.

9 Transactions, No. 1-9.

Circuit Theory

Annual publications fee: $3.

Design and theory of operation of circuits for use in radio and electronic equipment.

Dr. W. H. Huggins, Chairman, the Johns Hopkins Univ., Baltimore 18, Md.

18 Transactions, *1, *2, *Vol. CT-1, Nos. 1-4; CT-2, No. 1-4; CT-3, Nos. 1-4; CT-4, No. 1-4.

Communications Systems

Annual publications fee: $2.

Radio and wire telephone, telegraph and facsimile in marine, aeronautical, radio-relay, coaxial cable and fixed station services.

Mr. J. W. Worthington, Jr., Chairman, Rome Air Force Depot, Rome, N.Y.

10 Transactions, 5 Newsletters, *Vol. CS-1, No. 1; *Vol. CS-2, No. 1-2; CS-3, No. 1; CS-4, No. 1-2-3; CS-5, No. 1, 2, 3.

Component Parts

Annual publications fee: $3.

The characteristics, limitation, applications, development, performance and reliability of component parts.

Dr. R. M. Soria, Chairman, American Phenolic Corp., Chicago, Ill.

11 Transactions, *PGCP-1-2-3-4. Vol. CP-3, No. 1-3; CP-4, No. 1-2, 3; CP-5, No. 1.

Education

Annual publications fee: $3.

To foster improved relations between the electronic and affiliated industries and schools, colleges, and universities.

Dr. John D. Ryder, Chairman, College of Engineering, Michigan State Univ., East Lansing, Mich.

1 Transactions, Vol. E-1, No. 1.

Electron Devices

Annual publications fee: $3.

Electron devices, including particularly electron tubes and solid state devices.

T. M. Liimatainen, Chairman, Diamond Ordnance Fuze Lab., Washington, D.C.

21 Transactions, 3 Newsletters, 2 Technical Bulletins. *1, *2, *4, *Vol. ED-1, No. 1-4; ED-2, No. 1-4, ED-3, No. 1-2-3-4. ED-4, No. 1-2,3,4, ED-5, No. 1.

Electronic Computers

Annual publications fee: $2.

Design and operation of electronic computers.

Dr. Willis H. Ware, Chairman, Rand Corp., Santa Monica, Calif.

24 Transactions, 5 Newsletters, *Vol. EC-2, No. 2-4; *Vol. EC-3, No. 1-4, EC-4, No. 1-4; EC-5, No. 1-2-3-4. EC-6, No. 1-2, 3, 4.

Engineering Management

Annual publications fee: $3.

Engineering management and administration as applied to technical, industrial and educational activities in the field of electronics.

Dr. C. R. Burrows, Chairman, Ford Instrument Co., Long Island City, N.Y.

11 Transactions, 8 Newsletters, *1, *2-3. EM-3, No. 1-2-3. EM-4, No. 1-2, 3, 4. EM-5, No. 1.

Engineering Writing and Speech

Annual publications fee: $2.

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.

Mr. D. J. McNamara, Sperry Gyroscope Co., Great Neck, L. I., N. Y.

1 Transactions, Vol. EWS-1, No. 1.

Human Factors in Electronics

Annual publications fee: $2.

Development and application of human factors and knowledge germane to the design of electronic equipments.

Mr. Henry P. Birmingham, Chairman, U. S. Naval Research Lab., Washington 25, D.C.

THE INSTITUTE OF RADIO

Professional Group Meetings

(Continued from page 122A)

Washington—February 5

"Some Applications of Microwaves at Very Low Temperatures," W. B. Gager, Natl. Bureau of Standards.

Washington, D. C. Section—January 8

"Some Applications of Microwaves at Very Low Temperatures," W. B. Gager, National Bureau of Standards.

Military Electronics

Fort Walton—March 4

"Microwave Amplification by Stimulated Electronic Radiation," R. F. Simons, Airborne Instruments Lab.

Fort Walton—February 19

"High Gain Telemetry Antennas," C. H. Hoeppner, Radiation, Inc.

Long Island—January 7

"Engineering Management for Electronic Sub-systems," C. W. Thompson, WADC, WPAFB.

Los Angeles—February 18

"Russian Electronics," F. S. Atchison, U. S. Naval Ordnance Lab.; "The NOTS Precision Velocity Measuring System," Benjamin Glatt, U. S. Naval Ordnance Test Sta.

Philadelphia—January 22

"Trends in Modular Design," George King, Radio Corp. of America; "Module Environment—Its Effect on Design," J. C. McElroy, Collins Radio Co.

San Francisco—December 12

"The Electronics Engineer on the Weapons Systems Team," J. J. Dover, Edwards Air Force Base.

Washington, D. C.—February 10

"The ABC's of Missile Guidance," B. Van Dusen, Navy Bureau of Ordnance.

Nuclear Science Medical Electronics

Boston—February 4

Panel Discussion—"Problems of Radiation Health Monitoring," Sam Levin, MIT; Gorden Brownell, Massachusetts General Hospital; Constantine Maletskos, MIT; John Leborveau, Yankee Atomic Electric Co.

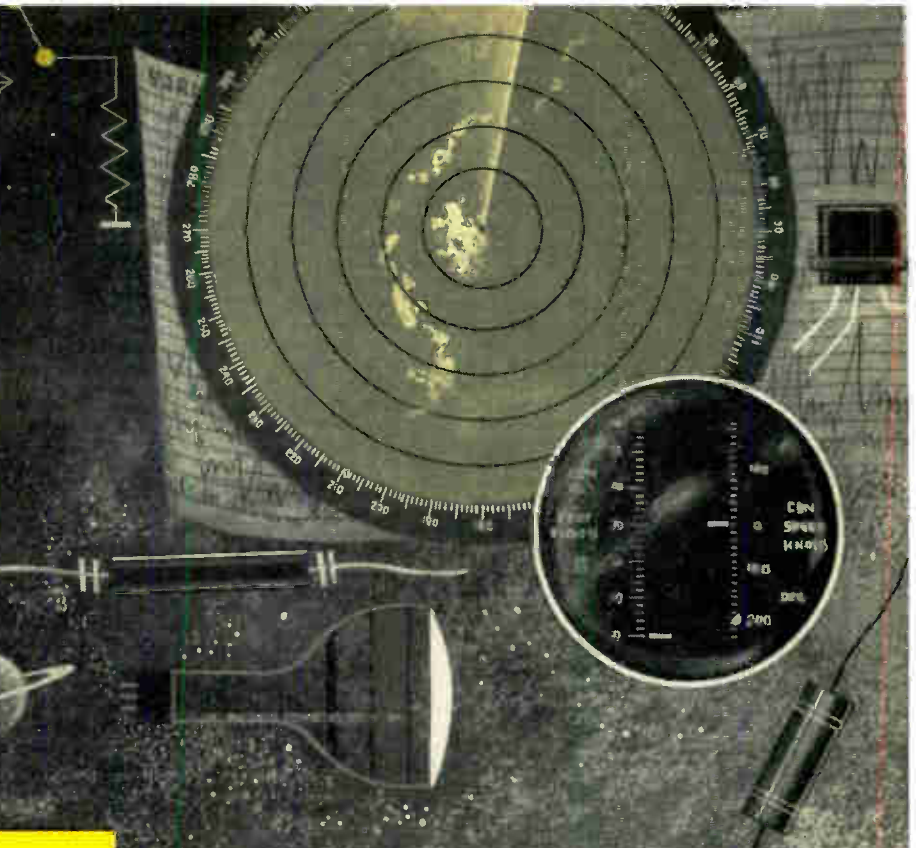
Production Techniques

San Francisco—February 25

"Case for Printed Circuits," Calvin Larson, Printronics Corp.; "Case Against Printed Circuits," John Coffin, Lenkurt Electric Co., J. R. Morison, Lenkurt Electric Co.

(Continued on page 128A)

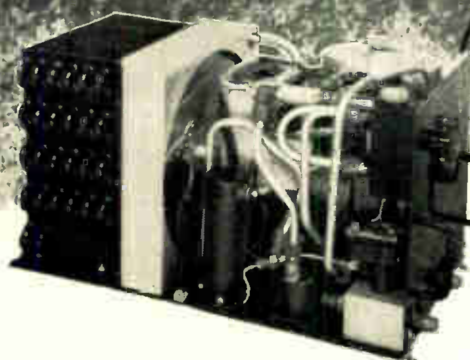
cooling avionic systems



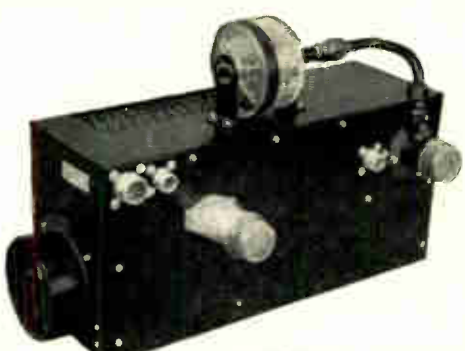
During World War II, Eastern Industries pioneered cooling systems for aircraft electronic systems. Now, thousands of installations later, and as the leader in this challenging field, Eastern is still pioneering.

Experience has been a springboard to new developments . . . compactness, simplification, refrigeration cycles. Research and development continue to play their vital parts in perfecting systems to overcome the new problems as expanded aircraft performance produces fantastic rises in temperatures.

If you have a challenging problem, come to the leader in the field for complete and creative engineering help.



COOLING UNIT



REFRIGERATION-TYPE

ELECTRONIC TUBE COOLING UNITS

Custom-made units, with or without refrigeration cycles, provide a method of maintaining safe operating temperature limits in electronic equipment. Standard sub-assemblies and components normally are used to create a custom-made design to fit your exact needs. Costs are minimized for these completely self-contained units by combining heat exchangers, fans or blowers, liquid pumps, reservoirs, flow switch, thermostat, and other common components.

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West Coast Office: 1608 Centinela Avenue • Inglewood 3, California

Professional Group Meetings

(Continued from page 126A)

San Francisco—January 28

Plant Tour through Turbine Condenser and Gear-Assembly Mfg. Facilities.

Washington, D. C.—December 10

"The Effect of Mechanized Production Techniques Upon Reliability," R. C. Marder, Stavid Engineering, Inc.; "Achieving Reliability Through Improved Production Techniques," S. Levine, Melpar, Inc.

Reliability & Quality Control

Florida West Coast—January 15

"The Sandia Laboratory Quality Assurance Program," L. J. Paddison, Sandia Corp.

Los Angeles—January 20

"Reliable Power Supply Design Using Semiconductor Rectifiers," Don Waldman, Int'l. Rectifier Corp.; and R. F. Edwards, Int'l. Rectifier Corp.

Philadelphia—January 23

"Reliability Prediction Techniques," E. R. Jervis, Aeronautical Radio Inc.; J. H. Hershey, Bell Tel. Labs.; M. P. Feyerherm, RCA (panel discussion).

Telemetry & Remote Control

Philadelphia—February 19

"Radio Web Navigation," Laurence Fels & R. D. Laughead, Jr., Stavid Engrg., Inc.

Philadelphia—January 15

"The Historical Development of Remote Control," E. S. Purington, Hammond Research Corp.

Washington, D. C.—January 30

"Radio Tracking of Earth Satellites," F. J. Friel, U. S. Naval Res. Lab.

Vehicular Communications

Chicago—December 13

"Public Air-Ground Telephone Service Trial," L. M. Augustus, Michigan Bell Telephone Co.

Detroit—January 22

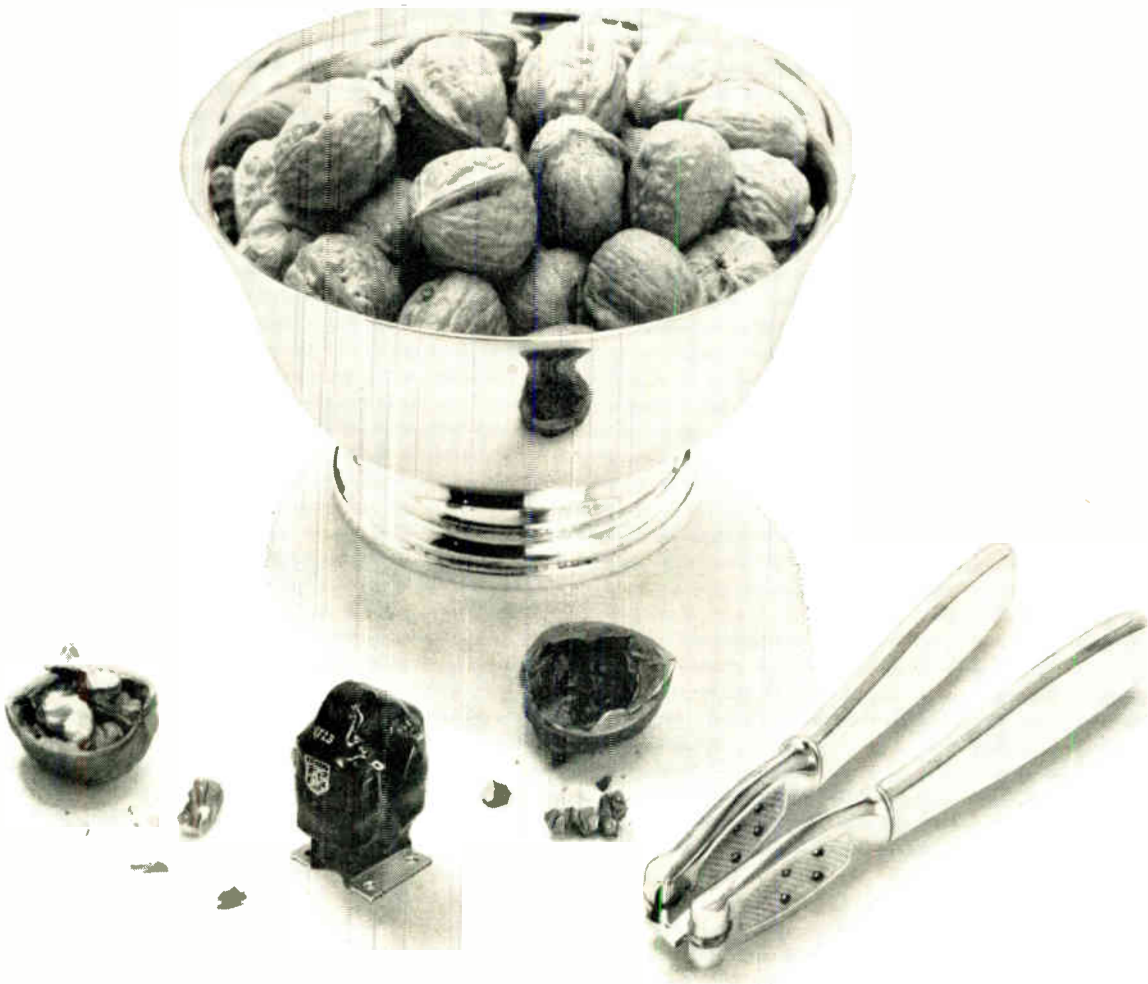
"Calibration of Mobile Radio Test Equipment," L. Elias, S. Sterling Co.

Detroit—November 20

"Tone and Coded Squelch Applications," S. Meyer, Allen B. DuMont Labs.

(Continued on page 130A)

Use your IRE DIRECTORY! It's valuable!

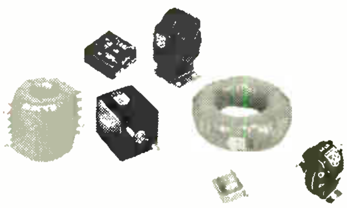


NO PULSE TRANSFORMER PROBLEM TOO TOUGH TO CRACK!

This is our guarantee: we'll solve your toughest transformer problem or it can't be cracked anywhere! From our complete line of prototype pulse transformers, medium and low-power transformers and recently-developed wide band video transformers (range of 50 CPS to 8.0 MC) we'll come up with the right custom-built or stock unit...in the size and shape to meet your specific needs.

When the prototype transformer is developed, it reaches your desk accompanied by a comprehensive laboratory report, which includes submitted electrical requirements, photo-oscillograms (which indicate input and output pulse shape and output rise-time), the test equipment used and an evaluation of the electrical characteristics of the prototype. Write for our complete new catalog-folder today.

Pulse transformers • Medium and low-power transformers • Filters of all types • Pulse-forming networks • Miniature plug-in encapsulated circuit assemblies



electronic components division

ESC®

exceptional employment opportunities for engineers experienced in pulse techniques

CORPORATION · 534 BERGEN BOULEVARD · PALISADES PARK, NEW JERSEY

Professional Group Meetings

(Continued from page 128A)

Florida West Coast—January 16

"History, Aims and Advantages of IRE Membership, and Development of Professional Groups and Chapters in the IRE," Capt. E. N. Dingley, Jr., Electronics Communications, Inc.; "Report on National Convention of Vehicular Communications in Washington, D. C.," D. C. Bailey, Dir. of Communications, City of Tampa.

Philadelphia—January 30

"Electronic Design and the Operator," B. J. Smith, General Electric Co.

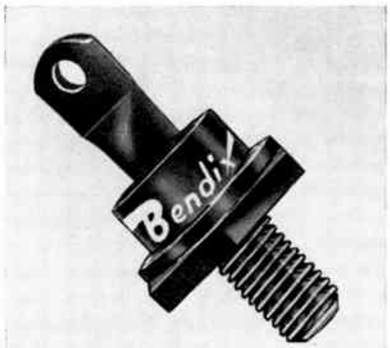
NEWS New Products

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

(Continued from page 44A)

Diffused Junction Silicon Rectifiers

A series of new silicon rectifiers have just been released for production by the **Bendix Aviation Corp., Red Bank Division,** 201 Westwood Ave., Long Branch, N.J.



The rectifiers have peak inverse voltage ratings ranging from 50 to 600 volts and can deliver 30 amperes of rectified current. The operating temperature extends from −65°C to +175°C. The rectifier package is in conformance with the latest JETEC proposed standards.

The rectifiers are of the diffused junction type for lower forward drop and lower reverse leakage current. EIA has reserved the JETEC designations 1N1434–1N1438 for this series of 5 rectifiers.

Besides application to power rectification, these units are useful in magnetic amplifier and dc blocking circuits.

For further information please contact Marketing Dept., Semi-conductor Prods., Bendix Aviation Corp., Red Bank Div., Long Branch, N.J.

(Continued on page 132A)

NOW . . . a complete AM-FM Service

unavailable from any other source

SUBSCRIBING TO OUR AM-FM SERVICE, you first receive our 212-page 1958 AM-FM Station Directory which lists all U. S., Canadian, Mexican, Cuban and other North American stations (1) by States and Cities, (2) by Frequencies; also, all AM & FM applications pending as of Dec. 31, 1957 (1) by States, (2) by Frequencies. There is a special listing of all Stations by Call Letters, and a listing of Type-Accepted Transmitters and Monitors.

The 1958 AM-FM Directory is punch-holed, and contained in a sturdy binder big enough also to include the Weekly Addenda you will receive. These run 4-6 pages, and fully report each week's FCC actions—grants, additions, changes, applications, etc. Also reported are latest radio station sales, first as announced and then as acted upon by the FCC.

The AM-FM Addenda pages contain all data available through each Friday, are mailed Saturday, usually will reach you Monday. They are designed, in a word, to keep the AM-FM Station Directory up-to-the-minute for ready reference.

New subscribers will, of course, receive all 1958 Addenda to date, along with the Directory, in the embossed binder.

- **Full year of AM-FM Service . . $50.00**
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Company____________________

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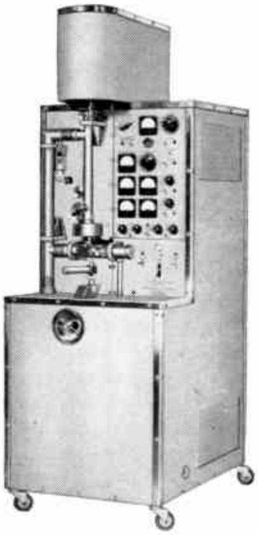
City & State____________________

☐ Bill me ☐ Bill company

high-speed microoscillograph

flat response to ten thousand megacycles



Central Research Laboratories' High-Speed Microoscillograph makes single-sweep oscillograms of three simultaneous phenomena at frequencies up to 10,000 megacycles. With a high sensitivity of 0.2 volt per trace width, this is the instrument of choice for recording phenomena occurring in time intervals of 10^{-8} to 10^{-10} second.

Electromagnetically focussed beams, 0.01 millimeter in diameter, write directly on a photographic plate inserted into the vacuum chamber through a vacuum lock. One plate holds 27 oscillograms with no overlapping. A complete cycle of photographic plate changing and reestablishing operating vacuum takes less than 5 minutes.

6 individually-shielded deflecting systems are provided: 3 signal, 3 time. Signal deflecting systems are of traveling-wave type with a nominal impedance of 50 ohms.

The instrument complete with all necessary pumps, gages, and power supply circuits weighs 700 pounds on a caster-mounted chassis of 26″ x 36″ x 76″

For complete information, write to

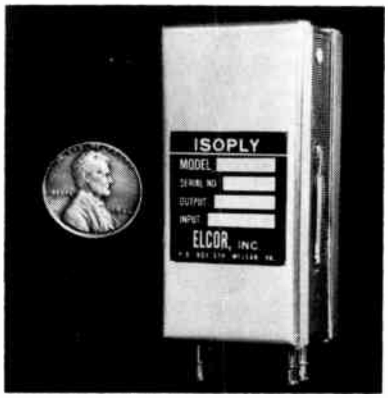
Central Research laboratories, inc.
Dept. 204 Red Wing, Minn.

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)

Subminiature Power Supply

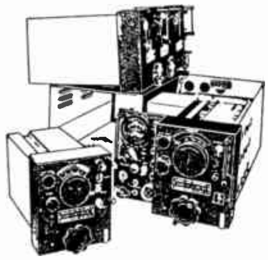


Small size and weight are outstanding features of this new isolated power supply designed for transistor circuits by **Elcor, Inc.**, P.O. Box 354, McLean, Va. Measuring only $\frac{7}{8}\times1\frac{1}{8}\times2\frac{1}{2}$ inches and weighing less than three ounces, the supply furnishes enough regulated and filtered dc power for the collector circuit of a transistor. Alternatively, it may be used as a source of bias for transistor or vacuum-tube circuits.

(Continued on page 134A)

AN/APT-5 TRANSMITTERS

Make excellent power signal generators for the range 300-1625MC., rated at 58 watts CW RF at 500MC. Contains blower-cooled 3C22 in re-entrant cavity with precision cathode, plate and loading controls, plus 6 tube AM modulator and amplifier flat from 50KC to 3MC. (easily converted to audio) with phototube noise generator. 115 volt 60 cycle filament supply. New, in export packing, with matching special plugs, lecher line, alternate feedback assembly, manual, audio conversion instructions and technical data, at $250.00. Limited stock.



AN/APR-4 RECEIVERS

with all five Tuning Units covering 38 to 4,000MC.; versatile, accurate and compact, the aristocrat of lab receivers in this range. Complete with wideband discone antenna, wavetraps, 100 page manual, plugs, cables and mobile accessories as required.

The AN/APR 4 has been our specialty for over ten years. Over 40 hours of laboratory time is invested in each complete set immediately prior to shipment, so that we can not only guarantee it to pass A-N specifications, but to have the inevitable mass-production irregularities corrected and the latest improvements added. We maintain a complete stock of spare parts and expect to service our customers' sets indefinitely. Write for data sheet and quotation. Also, prior models SCR-587A, RDO and AN/APR-1.

TEST EQUIPMENT

of highest quality, new and reconditioned, military surplus and top commercial brands in stock at money-saving prices. Let us quote to your requirements.

ENGINEERING ASSOCIATES
434 PATTERSON ROAD, DAYTON 9, OHIO

"Is seat of pants Amerikan spacemen is flying by. Is not knowing of Reeves-Hoffman's . . .

NEW HIGH PRECISION CRYSTAL FOR FREQUENCY MEASUREMENT

WRITE FOR BULLETIN RH-5MC.

REEVES-HOFFMAN

Designed for use as frequency standards, Reeves-Hoffman's new 5mc, high precision crystals offer exceptionally long term frequency stability, ±.0001%, with aging of less than one part per 10^8 a week! These units are available in hermetically sealed glass T5 1/2 enclosures with pigtail leads or 9-pin Bakelite bases. They are manufactured to meet the most exacting military and commercial standards for frequency measurement.

DIVISION OF DYNAMICS CORPORATION OF AMERICA
CARLISLE, PENNSYLVANIA

NEWS New Products

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

(Continued from page 132A)

Various models of these inexpensive Zener-diode regulated Isoplys are available with input for 60–400 cps ac and with dc output voltages ranging from 4 volts to 26 volts and current ratings ranging from 9 ma at 4 volts to $1\frac{1}{2}$ ma at 26 volts. Shunt capacitance from output to ground is 20 $\mu\mu f$, making the supply useful as a means of direct coupling in high-speed circuits, and in many bridge circuits in which a signal voltage appears between the power supply output and ground. Leakage resistance to ground exceeds 10,000 megohms. Special shielding reduces the noise in ungrounded applications to the negligible value of 1 microvolt per kilohm impedance to ground. No-load to full-load regulation is 2 per cent and regulation for an input voltage change of 10 per cent is 1 per cent. Output ripple is less than 5 mv rms.

Reliable Transistors

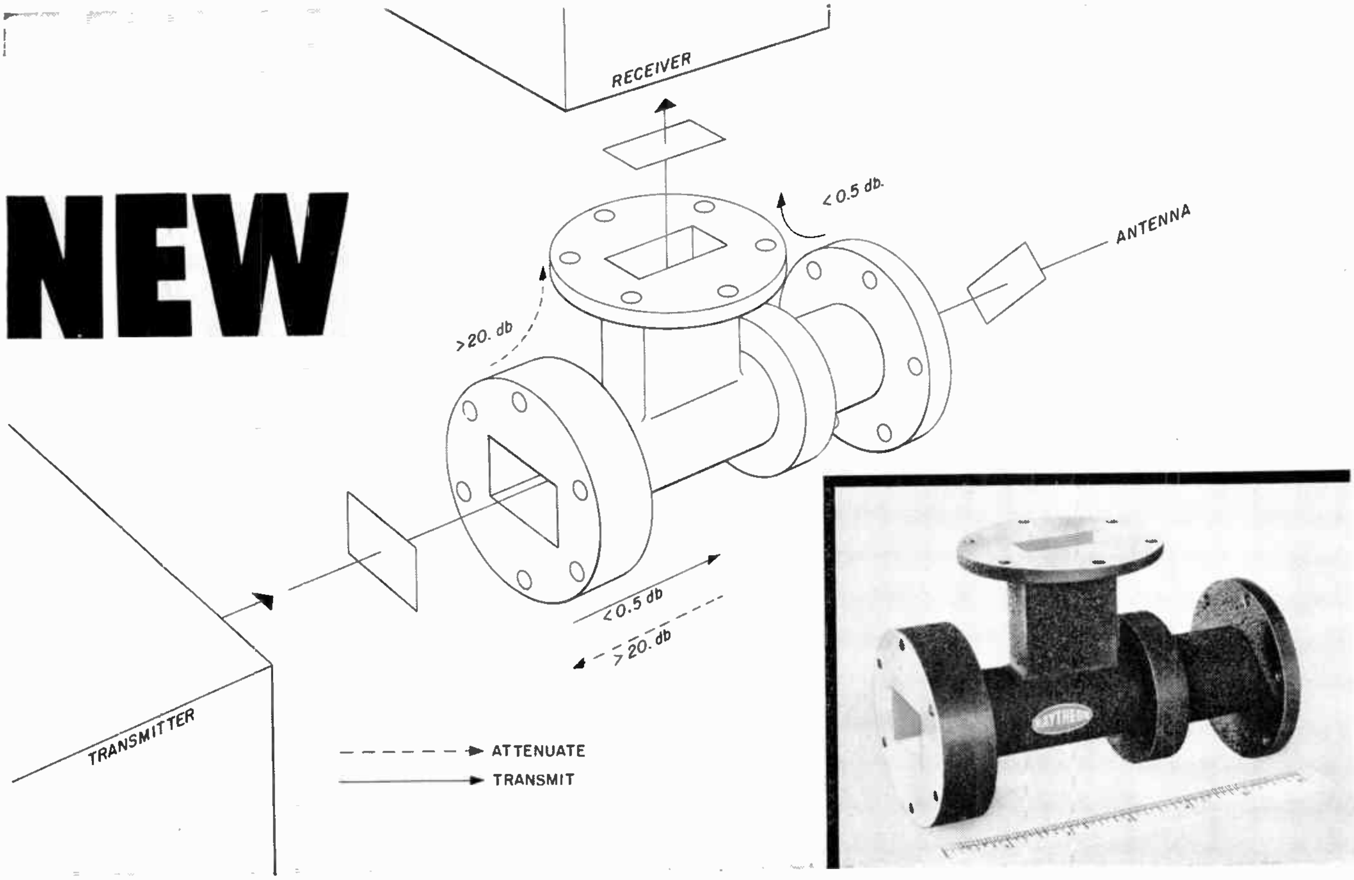
The first complete line of Reliable Subminiature Transistors has been announced by **Raytheon Manufacturing Co.**, 55 Chapel St., Newton 58, Mass. These are PNP Germanium units made by the Raytheon-perfected fusion-alloy process which ensures reliability. 13 Types are available in quantity production.

The new Subminiature types have a volume only one-fourteenth that of the Jetec-30 package so are suited where space is a consideration. Four types CK25, CK26, CK27 and CK28 duplicate the electrical characteristics of the Raytheon computer types previously announced in the larger package. Four more types, CK13, CK14, CK16 and CK17 are for general purpose rf use; four types, CK64, CK65, CK66 and CK67 are for general purpose audio use. CK22 is a low noise audio amplifier.

Further information about these 13 new transistors is available from Technical Information Service, Raytheon Mfg. Co.

(Continued on page 156A)

NEW



MICROWAVE FERRITE CIRCULATOR . . .

Compact C-band unit replaces gas-tube duplexer; needs no external power.

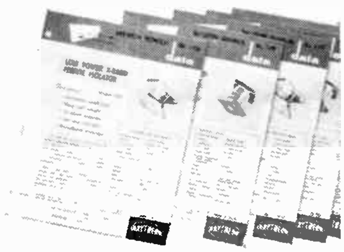
System designers: This new circulator is lighter and more compact than the differential phase-shift type unit and readily replaces typical TR or ATR gas tubes in C-band microwave transmission systems.

The Raytheon Model CCM1 weighs less than 5 lbs. and is less than 6 inches long. Its permanent magnet design eliminates the need for external drive power. The CCM1 reduces requirements for filters and klystron isolation common to systems using T-junction duplexers.

With Raytheon's advanced microwave component designs like this new C-band circulator, systems designers now have more freedom than ever before to design compact lightweight packages. Other devices now available and in advanced stages of development include isolators, both high and low power, ranging from L-band to Ku-band; ferrite switches; modulators; and side-band generators.

RAYTHEON MINIATURIZED X-BAND ISOLATORS weigh as little as 2.2 oz. For somewhat different requirements in the lower frequency L-band, Raytheon recently introduced the first high-power L-band isolator commercially available.

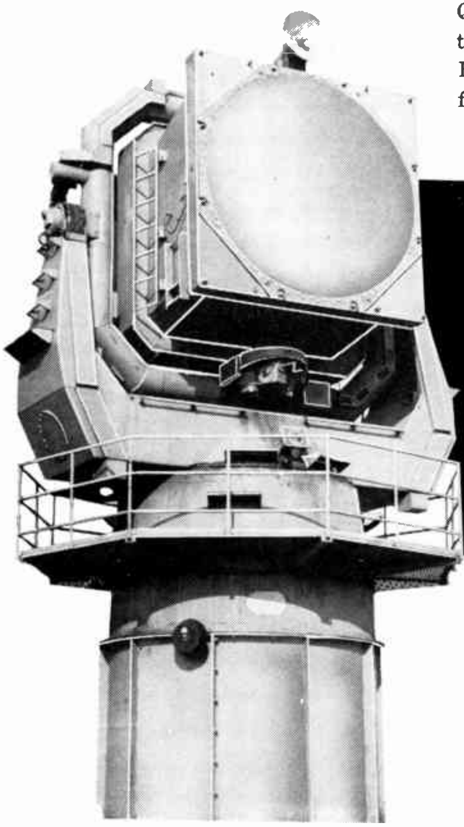
FOR COMPLETE FACTS or assistance in solving your microwave ferrite component problems, simply write to the address below, outlining your requirements.



Excellence in Electronics

RAYTHEON

RAYTHEON MANUFACTURING COMPANY
Special Microwave Device Group
100 River Street, Waltham 54, Massachusetts



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.

DEAN OF ENGINEERING

Dean of Engineering or Department Chairman, Electrical Engineering or Physics. Ph.D. Under 40. Listed "Who's Who in Engineering," "American Men of Science," etc. 12 years academic experience (full time and part time) teaching graduate and undergraduate engineering, and thesis supervision. Many publications IRE, etc. Applied mathematics, UHF antennas and propagation, magnetic alloys, missile systems. Industry last 6 years. Box 1029 W.

ELECTRICAL ENGINEER

BSEE. University of Notre Dame, 1956. Age 23; married. Experience: 2 years as a repair technician in radio and television. Presently serving in the Navy as Communications Superintendent with 1 year experience as a Ship Superintendent. Just released from active duty. Desires engineering position with future supervisory opportunities in the Chicago area. Box 1030 W.

ENGINEER

BSEE. Age 27. Private pilot with 4 place plane wants sales or field engineering position. Prefer overseas or South American location (speaks 4 languages). R&D experience electronics, industrial instrumentation; field engineer training and experience avionics, airborne radar, electronic and electromechanical computers and systems. Box 1031 W.

ELECTRICAL ENGINEER

BSEE., MEE. Presently instructor in a large metropolitan university in New York City. Experienced in propagation and microwaves, theoretical and experimental. Desires position on a one day per week basis in the New York City vicinity. Box 1032 W.

PROFESSOR

Ph.D. engineering physicist, designer of digital computers, and theoretical engineer. Teaching experience. Ready to join university, especially small university with high hopes of future growth. Box 1033 W.

INSTRUCTOR

Expecting MSEE. in August 1958. 2 years experience in electronic research plus limited experience as teaching assistant. Bachelor. Age 29. Any location considered. Box 1034 W.

(Continued on page 140A)

ABOUT MEN AND WEAPONS

For centuries men have tried to develop new and more powerful weapons to achieve victory in war.

Lately these have been weapons of unprecedented power.

Now war can become race suicide, and victory thus gained is a delusion.

Yet we keep on trying to develop new and more powerful weapons, because we must.

Not because we seek victory through a nuclear war, but because through strength we may prevent one.

For as long as there are powerful forces with a record of cynical duplicity and oppression, the free world must have weapons capable of neutralizing them.

At least until men learn that the only alternate to peace is oblivion.

At Sandia, we play an important part in providing this protective strength. Our scientists and engineers are responsible for research, design, and development of nuclear weapons for the Atomic Energy Commission. This makes these men exceptionally valuable assets in our nation's efforts to secure the future.

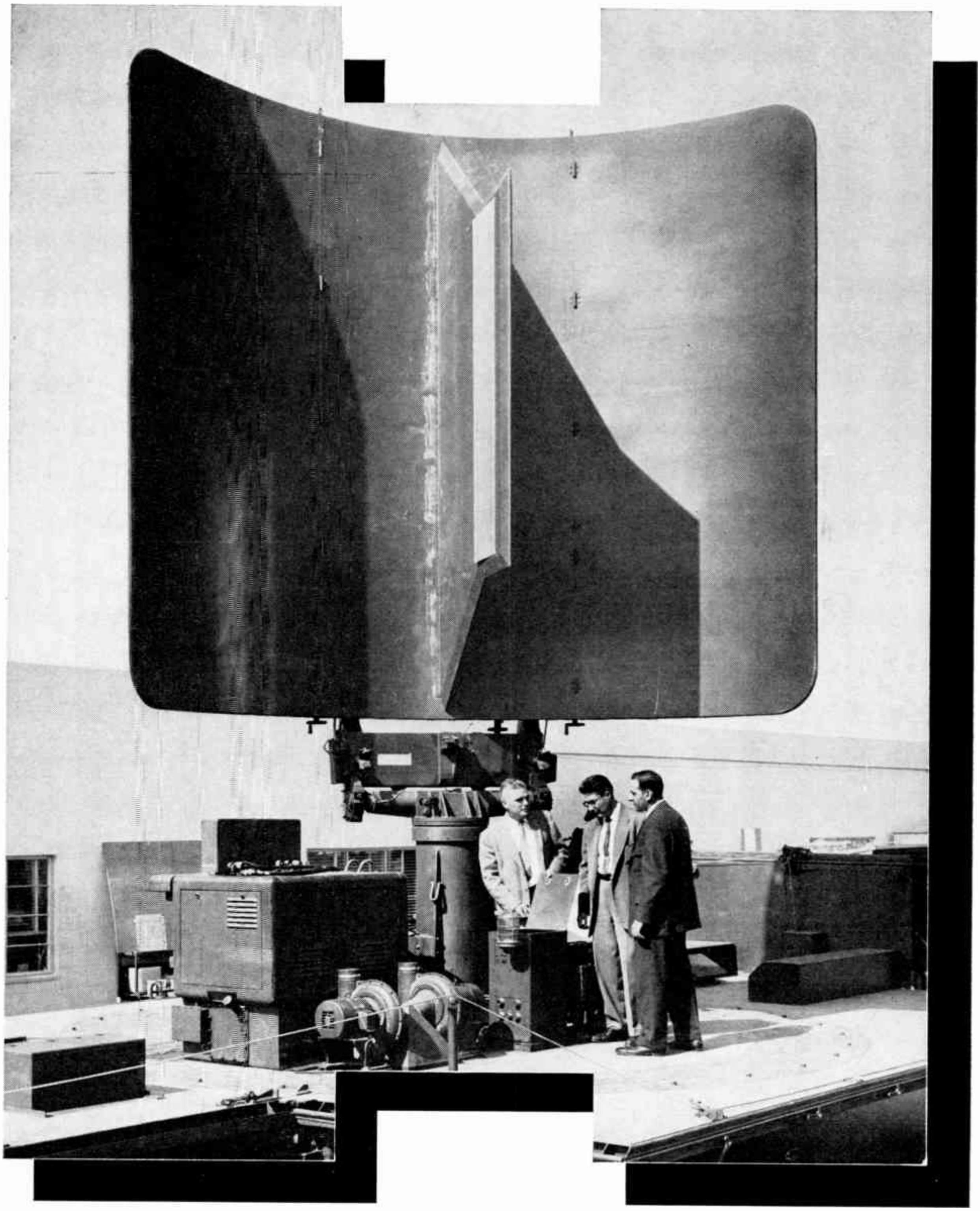
We need more such men — outstanding engineers and scientists in many fields, especially at the highest academic and experience levels. At Sandia in Albuquerque and at our branch laboratory in Livermore, California, we need their knowledge, skill, and perseverance.

If you can help us meet this need, or if you know anyone who can, write Staff Employment Section 554.

SANDIA CORPORATION

ALBUQUERQUE, NEW MEXICO

HOW TO SEE IN



ALL DIRECTIONS AT ONCE

They add new dimension to defense

Three dimensional radar . . . it is a positioning of radar beams in space by electronic rather than mechanical means. It provides three-dimensional target data from a single antenna, transmitter, and receiving channel. It is a radical new weapon for national defense.

Engineers at the Hughes Ground Systems Division in Fullerton are responsible for pioneering this advancement (see antenna at left). But even more importantly, these same engineers are working on an elaborate radar warning system which will not only provide this complete radar data, but also translate it into meaningful information and relay it to central communications centers.

Other Hughes activities offer similar engineering challenge. The Research and Development Laboratories in Culver City, for example, are probing into the effects of nuclear radiation on electronics equipment, studying advanced microwave theory and applications, examining communications on a global scale, and developing new methods for insuring product reliability.

The Hughes Products engineering team makes electronics useful in solving industrial problems. For example, this group has just unveiled an industrial electronics system which will automate a complete and integrated line of machine tools.

The diversity of Hughes activity offers prospective employees opportunity to build a rewarding career in a highly progressive and expanding environment.

New commercial and military contracts have created an immediate need for engineers in the following areas:

Communications	Microwaves
Reliability	Crystal Filters
Circuit Design	Computer Engineering
Systems Analysis	Field Engineering
Vacuum Tubes	Semiconductors

Write, briefly outlining your experience, to Mr. Phil N. Scheid, Hughes General Offices, Bldg. 17F-1, Culver City, California.

Advanced research on the Maser (Microwave Amplification by Simulated Emission of Radiation) performed by the R&D Laboratories is directed towards applications of a portable, airborne Maser for missiles and aircraft.

Falcon missiles have been an important factor in establishing Hughes as a leader in advanced airborne electronics. Manufactured in Tucson, Arizona, the Falcon missiles have both infrared and radar guidance systems.

Creating a new world with ELECTRONICS

HUGHES

HUGHES AIRCRAFT COMPANY
Culver City, El Segundo,
Fullerton and Los Angeles, California
Tucson, Arizona

Positions Wanted

By Armed Forces Veterans

(Continued from page 136A)

ELECTRICAL ENGINEER

BSEE. Eta Kappa Nu. 1 year Navy electronics technician; 1 year digital computer sales; 2 years fire control systems field service engineering; 2 years nuclear instrumentation sales including contract negotiation, preparation of proposals and advertising. Age 30. Desires work in suburban New England area. Box 1041 W.

TELEVISION BROADCAST TECHNICIAN

Desires position in any phase of broadcasting. Age 23; married. 2½ years in AM xmtr and TV studio equipment operation and maintenance. Will relocate. Resume upon request. Box 1042 W.

ELECTRONIC ENGINEER

BEE. plus graduate work. 6 years experience R&D electronic circuits, including feedback amplifiers, analogue to digital conversion, pulse generating and gating methods, logic circuit design, oscillators, and regulated power supplies. Desires instrumentation work in computers/automatic control. Box 1043 W.

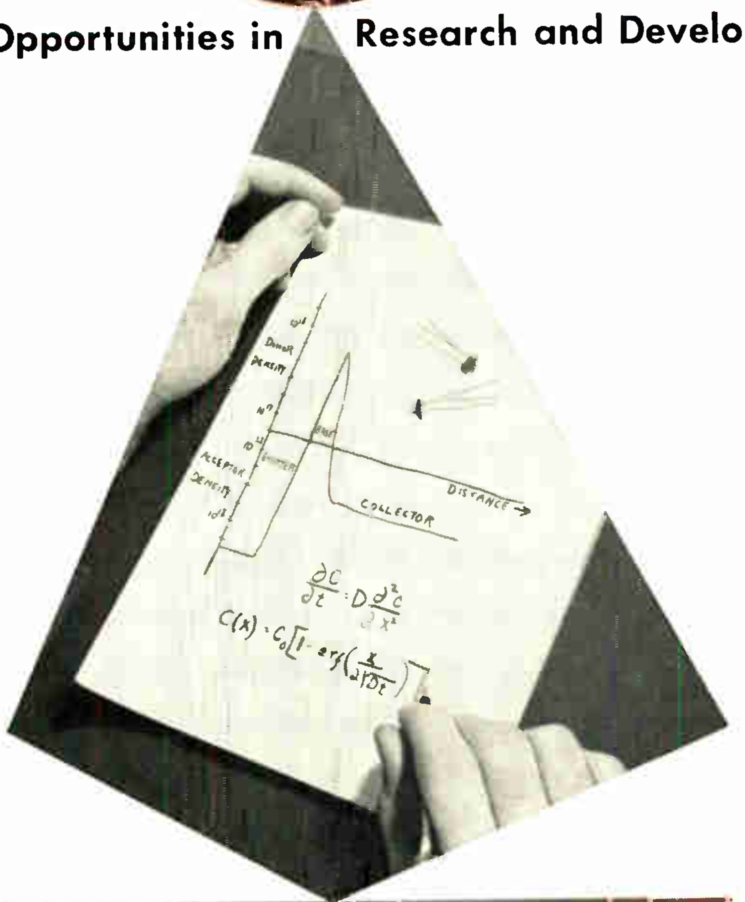
ENGINEER

BS. USMA. 1948; MSEE. University of Illinois 1954; age 32; married. 3 years Signal Corps in Germany; 4 years in Special Weapons Training Group; Prefer southeast United States location and missile or special weapons work. Box 1044 W.

(Continued on page 142A)

Today's emphasis on advanced transistor technology and solid-state devices and materials has opened up new opportunities in research and development at IBM.

Opportunities in Research and Development at IBM



So broad are IBM's activities in the electronic computer field that a properly qualified applicant finds the widest variety of research and development assignments available. Which of these areas, now open at IBM laboratories at Endicott, Poughkeepsie, and Yorktown, New York, offers the kind of career you want?

- Solid-state physics
- Semi-conductor devices
- Analytical mechanics
- Mathematics, topology
- Physical chemistry
- Magnetic devices
- Circuitry application
- Reliability studies

Qualifications

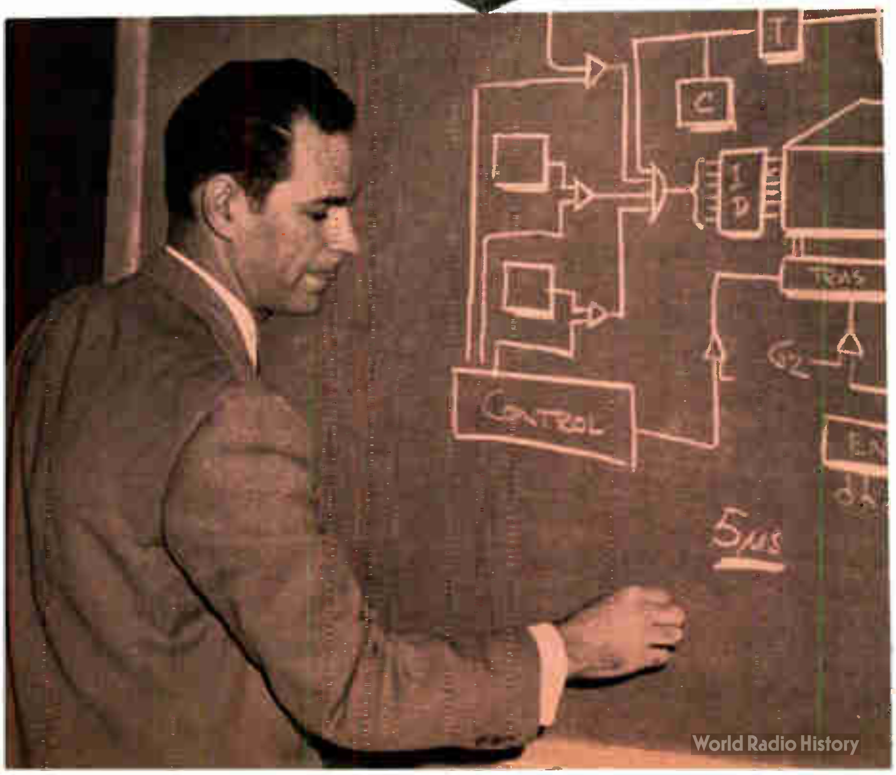
Degree in electrical, mechanical or chemical engineering, physics, engineering physics, mathematics, metallurgy or chemistry, and

Proven ability to assume a high degree of technical responsibility in your sphere of interest.

FOR DETAILS, just write, outlining background and interests, to:

Mr. R. A. Whitehorne, Dept. 645Q
Mgr. of Technical Recruitment
International Business Machines Corp.
590 Madison Ave., New York 22, N. Y.

IBM is a recognized leader in the rapidly expanding electronic computer field with a stable balance of military and commercial work. You will find exceptional opportunities for technical achievement and growth at IBM. Original developments and publications will undoubtedly stem from your work. Salaries are excellent. Liberal company benefits set standards for industry.



IBM INTERNATIONAL BUSINESS MACHINES CORPORATION

DATA PROCESSING
ELECTRIC TYPEWRITERS
MILITARY PRODUCTS
SPECIAL ENG'G. PRODUCTS
SUPPLIES
TIME EQUIPMENT

Plants and laboratories: Endicott, Kingston, Owego, Poughkeepsie, Yorktown, N. Y.; Lexington, Ky.; Rochester, Minn.; San Jose, Calif.

Positions Wanted

By Armed Forces Veterans

(Continued from page 140A)

ELECTRONIC TECHNICIAN

Four years communications experience with U. S. Air Force. A graduate of RCA Institute. Holder of 1st class radio telephone license. Box 1045 W.

SEMICONDUCTOR ENGINEER

Senior engineer with 6 years varied experience in manufacture and development of transistors and diodes. Desires position providing diversified engineering and management opportunities. Graduate engineer, married, age 31. Box 1046 W.

ENGINEER

BSEE., BS. math., MSEE.; desires teaching or research position where part-time graduate work may be taken toward a Ph.D. Also considers schools having no Ph.D. program if any two of the following three languages are taught: German, Russian or French. Available June 1958. Box 1047 W.

ENGINEER

Desires to travel—or field service. RCA graduate with 3 years radio and TV engineering experience; 1 year computer installation and maintenance. Presently employed in this field. Equipped to travel. Box 1054 W.

(Continued on page 146A)

LINK YOUR CAREER TO TOMORROW

climate:

For career men it is almost impossible to tell where mental climate leaves off and physical climate begins. Like the Moebius strip, properly joined, they present a single-sided situation that is most benign. At Link Aviation the climate is ideal . . . Pioneers of flight simulation, Link's modern electronic laboratories now bracket our continent. The facilities in Binghamton, New York, have been supplemented by the energetic Research & Development Laboratories at Palo Alto, California. At both places, primary consideration was given to the importance of climate to you . . . the very engineer we now seek.

LINK

GENERAL PRECISION EQUIPMENT CORPORATION GPE

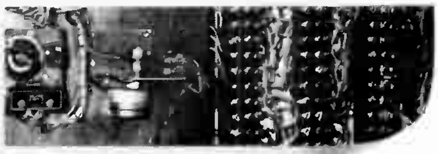
The mental climate at Link Aviation, Inc., is ideal for your growth. Its facilities, its attitude, its personnel provide a congenial, purposeful atmosphere.

The physical climate furnishes you with the maximum in comfortable suburban living. And you have your choice between the full-seasoned stimulation of Binghamton in upstate New York, or the temperate California living in Palo Alto.

The climate of learning is here, too. If you are the man we want, you will never claim that your education is finished. Link Aviation does more than give lip service to this concept. As a participant in the Honors Cooperative Programs at outstanding universities, you can advance your education by attending regular classes with time and tuition provided by Link.

In addition to providing you with the ideal *climates,* Link also assures you that your material benefits will be commensurate with today's most enlightened practices; good pay, ample vacations, hospital, health and retirement benefits.

If this entire work-live-study setup at Link sums up the ideal *you* have established in your mind, we urge you to respond immediately.



Write:
Mr. Ken Viall
Link Aviation, Inc., Dept. P-5
Binghamton, New York
Or:
Mr. Joe Larko
Link Aviation, Inc., Dept. P-5
P.O. Box 1313
Palo Alto, California

Link Aviation, Inc. *Binghamton, New York*

Positions Wanted

By Armed Forces Veterans

(Continued from page 142A)

PRODUCTION OF SALES ENGINEERING

BSEE. MIT., 1952; MBA. Boston University. 1958. 2 years project administration experience Signal Corps engineering laboratories. 4 years equipment design experience on instruments and servomechanisms. Desires position with instrument manufacturer in production administration and control or application engineering. Age 27; married. Will relocate. Box 1055 W.

ELECTRONIC ENGINEER

Electronic engineer desires employment in Japan. 8 years experience, including radar, ECM, communications system engineering for Signal Corps, operations research. Interested in position with responsibility, utilizing technical ability. Box 1056 W.

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

ENGINEERS—PHYSICISTS

Positions are open on the research and development staff of Paul Rosenberg Associates for an electronic engineer with experience in applied research, development and design, and for a physicist with experience in electronic development and design. PRA is a 12 year old firm of consulting physicists (not manufacturers) engaged in unusually interesting and challenging R&D projects for Government and industry. Ideal working conditions in a suburb 25 minutes from the center of New York City. Applications kept in strict confidence. Paul Rosenberg Associates, 100 Stevens Ave., Mt. Vernon, New York.

ENGINEERS

Electrical, civil, mechanical and drawing faculty members required for rapidly growing degree granting engineering programs. Colleges located in pleasant small city at foot of Sierra Nevada recreation region within 3 hours driving time to San Francisco Bay area. Write, President Glenn Kendall, Chico State College, Chico, California.

ELECTRONIC ENGINEERS

One engineer in design, applications, testing and evaluation of tape wound metallic cores utilizing molypermalloy tapes and foils for pulse and magnetic amplifiers and memory cores for computers AND one designer wound plastic capacitors in close tolerance for practical application. Opportunity and salary attractive. All replies strictly confidential. Write, Mr. W. Aston, Witty Consultants, 415 Lexington Ave., New York 17, N.Y.

(Continued on page 148A)

• R-F RECEIVER DESIGN
• INERTIAL NAVIGATION

Two of many areas in Avionics in which Bell Aircraft has openings for qualified electronics engineers

Particularly good opportunities are now available for engineers with radio frequency experience in the 100 kilocycle to 35,000 megacycle range with emphasis on transistorizing of circuits . . . and for those with experience in inertial instrumentation design and evaluation.

Present openings include assignments in:

- Pulse and Digital Coding
- Identification Systems
- Electronic Counter Measures
- Landing Systems
- Digital Computers
- Precise Instrumentation Development

These assignments embrace a wide range of high level design and development problems which will afford full scope to your creative ingenuity with unusual opportunities for rapid advancement and professional recognition. Salaries commensurate with your background, good living and working conditions, and liberal benefits. Please write: Supervisor of Engineering Employment, Dept. O-24, **BELL AIRCRAFT CORPORATION, P.O. Box 1, Buffalo 5, N.Y.**

BELL
Aircraft CORP.

Positions Open

(*Continued from page 146A*)

AUDIO FILTER DESIGN ENGINEER

A director of research and development AND a design engineer of toroids, filters and related networks. Experienced in mechanical and/or crystal audio filter design. Excellent laboratory facilities, fast growing company. New York metropolitan area. Write, Mr. W. Aston, Special Projects Director, Witty Consultants, 415 Lexington Ave., New York 17, N.Y.

ENGINEERS

Scholarships to the value of $1500.00-$2100.00 per annum are offered for advanced study and research leading to the degree of M.Sc. and D.Sc. Additional payments up to $24.00 per week may be made for part-time teaching assistance. Application and inquiries to the Director, Dept. of Electrical Engineering, Laval University, Quebec, P.Q. Canada.

ASSISTANT PROFESSOR—INSTRUCTOR

Assistant professorship in electrical engineering; graduate degree required. Instructorships available for young BS. graduates who intend to pursue graduate work. Excellent salary scale; ideal San Francisco Bay location. Write to Head, Div. of Engineering, San Jose State College, San Jose 14, Calif.

(*Continued on page 152A*)

OUTSTANDING OPPORTUNITY FOR A SENIOR ELECTRONICS SCIENTIST

Unusually attractive high-level research opening for graduate engineer or scientist, aged about 35 to 45, with broad comprehensive background in electronics and electrical engineering. Must be capable of supplying technical leadership in coordination of research projects involving such technical areas as electronic circuit theory, digital and analog systems, solid-state devices, electromechanical systems, and related fields. Should have insight into economics of electronics and electrical manufacturing industry. Salary open, ranging $12,000 to $16,000 to start.

In reply, please state education, experience, age, marital situation, present salary, and starting salary required. Please address reply to Box No. 1073, INSTITUTE OF RADIO ENGINEERS, 1 East 79th St., New York 21, N. Y.

ENGINEERS: *Electronic & Mechanical* Physicists

Research · Development · Pilot Production

$$ds^2 = dx^2 + dy^2 + dz^2 - c^2dt^2$$

Is today the same as yesterday?

Every now and then a man stops and takes stock of himself and his career. He sizes up what he has accomplished. Where he is heading.

If you are doing just that and find that you are ready for a long step forward—for increased responsibility and stature—it may pay you to consider Melpar.

These forces will be working for your advancement when you join our organization: diversified and stimulating programs with an opportunity to follow projects from inception to prototype completion or production; an atmosphere of professionalism and regard for your individual ideas and contributions; a promotion policy based solely on your ability; a steady program of expansion which continually creates new positions.

Our well equipped laboratories and manufacturing facilities are located near suburbs that promise gracious living for your family and easy commuting for you.

For details about career opportunities at Melpar, write: Technical Personnel Representative.

MELPAR *Incorporated*

A Subsidiary of Westinghouse Air Brake Company

3038 Arlington Boulevard, Falls Church, Virginia

10 miles from Washington, D. C.

Openings are also available at our laboratories in Boston and Watertown, Massachusetts

Positions Open

(Continued from page 148A)

SEMICONDUCTOR DEVICE DEVELOPMENT

Senior and Junior positions in semiconductor device research. Senior position requires graduate degree and at least 2 years experience. Sponsored research in academic environment with unsurpassed technical facilities. Excellent salaries and all benefits including opportunity for graduate study. Our present staff knows of this announcement. Box 1070.

PROFESSOR OF ELECTRICAL ENGINEERING

Primarily to participate in graduate and research program. Opportunity to cooperate in development of well equipped new laboratory, space and facilities. Should have Ph.D. Location: east. Box 1071.

ELECTRONIC ENGINEER

Electronic engineer to teach lecture and laboratory courses. Up-to-date knowledge of the field required. Working and living conditions excellent; salary and opportunity very attractive. Write to Dean of Engineering, California State Polytechnic College, San Luis Obispo, California.

(Continued on page 156A)

SCIENTISTS
ENGINEERS

CALIFORNIA

offers you and your Family

- A world center of the electronic industry for CAREER ADVANCEMENT
- The High Sierras and the Pacific Ocean for RECREATION
- Some of the nation's finest public schools for your CHILDREN
- World Famous Universities for ADVANCED STUDY
- MAJOR CULTURAL CENTERS

while living in such places as
Exciting San Francisco
Fabulous Southern California
Cultural Palo Alto

companies pay interview, relocation and agency expenses

submit resume in confidence to:

PROFESSIONAL & TECHNICAL

Recruiting Div. of
Permanent Employment Agency
Brewster-Warren Building
Redwood City, Calif.

Special Opportunity for MICROWAVE SCIENTISTS

at the NEW

Westinghouse MICROWAVE TUBE CENTER at ITHACA, N.Y.

(Adjacent to Cornell University)

Our new Microwave Tube Center at Ithaca, N.Y. offers competent microwave engineers and physicists unusual opportunities. They can penetrate new and unexplored channels of investigation with the highest types of professional and academic associations, providing a stimulating environment for individual achievement.

Qualified candidates are encouraged to work for advanced degrees.

Working and living conditions are enjoyable in this delightful Finger Lakes region.

Salary commensurate with training and experience.

The following positions are now available:

PROJECT DIRECTORS

For those capable of initiating, directing and participating in theoretical and applied research and long range major development in the microwave tube field. Ph.D. or M.S. degree, plus industrial experience required.

MICROWAVE TUBE SPECIALISTS

To perform applied research and long range major development of a highly creative nature. Ph.D. or M.S. degree, plus industrial experience required.

Openings

also available in ELMIRA (30 miles from Ithaca, N.Y.) for Microwave Tube Development & Design Engineers.

Write or send resume to Mr. William Kacala, Technical Recruiting, P.O. Box 284, Dept. M1A2, Elmira, N.Y., or phone collect Elmira 9-3611. Weekends, 2-2139.

Westinghouse

ELECTRONIC TUBE DIVISION ELMIRA, NEW YORK

ENGINEERS... *Let's Trade Ideas!*

SPERRY PHOENIX

wants your ideas on advanced flight control systems . . .

and here's an idea → YOUR FAMILY will like!

IF data sensors, gyroscopics and/or advanced electronics are your specialty, your ideas are needed by this new. important division of Sperry Rand Corporation.

And we know you and your family will like the idea of wholesome, sun-filled. fun-filled life in Phoenix! You'll like the year-around picnic weather . . . clean, dry air . . . spectacular scenery . . . remarkably low housing costs . . . the wide, pleasant, uncrowded streets that carry you to work. You'll like Arizona's well-rated grade and high schools . . . and the fact that nearby colleges offer you advanced courses.

You and your family will like the idea of living in friendly, easygoing Phoenix!

We're building a basic staff at Sperry Phoenix Company to spearhead research, development and applications. Starting salaries and opportunities for increases are right at the top. There's a great future for top-notch idea men . . . and their families . . . in Phoenix.

Here's another good idea: mail the coupon today!

If you believe you are qualified for advanced flight control systems and associated aircraft equipment engineering, return this coupon without delay (no need to send resume) while these rare basic staff positions are still open. By return mail you will receive an application. Mail the coupon now . . . this is an unusual opportunity for you and your family!

SPERRY PHOENIX COMPANY

DIVISION OF SPERRY RAND CORPORATION

PHOENIX, ARIZONA

"Win" Ames
Sperry Phoenix Company • P. O. Box 2529, Phoenix, Ariz.
I'm interested! Please send me a confidential application.

NAME__________

ADDRESS__________

CITY__________ ZONE______ STATE______

I-48

RCA—leader in military electronics—now needs imaginative engineers for

Scientists and engineers have probed, and the mysterious world of electronics has steadily yielded rich rewards. But the massive door to electronic wonders is only slightly ajar. The greatest discoveries still lie ahead!

In the field of national defense, Project "X" takes many forms at RCA. It might be found in airborne electronics, surface communications, missile weapons systems, radar, ballistic missile detection systems. Or, it might be something entirely new in concept—something vital to the defense of the free world.

We are interested in talking with all experienced engineers and engineering managers who can contribute to this important work.

Attractive salaries . . . Desirable locations . . . Liberal benefits

TO ARRANGE INTERVIEW WITH ENGINEERING MANAGEMENT, SIMPLY CALL COLLECT OR SEND RESUME

AT CAMDEN, N.J.
—in Greater Philadelphia near many suburban communities

- AVIATION ELECTRONICS
- COMMUNICATIONS AND MILITARY TV
- SPECIAL DEVELOPMENT STUDIES

AT TUCSON, ARIZ.
—healthful, western climate

- COMMUNICATIONS SYSTEMS APPLICATIONS

TO: MR. C. B. GORDON, RCA, DEPT. X-1E
BLDG. 10-1, CAMDEN 2, N.J.
WOODLAWN 6-3300

AT WEST LOS ANGELES, CALIF.
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AT NEW YORK, N.Y.
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TO: MR. W. J. HENRY, RCA, DEPT. X-1E
MOORESTOWN, N.J.—BELMONT 5-5000

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Pay is excellent, advancement is rapid, the attitude is friendly and helpful. Other advantages: no commuting problems; colleges that offer advanced study; remarkably low home costs.

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If you are qualified for one of these positions, write, wire or phone Motorola in Phoenix today!

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System Analysis, Design and Test

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Navigation	Missile Guidance

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Circuit Design, Development and Packaging

Microwave	Pulse and Video
Antenna	Digital and Analog
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Servos

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Write:
Mr. Kel Rowan
Western Military Electronics Center
Motorola, Inc., Dept. C-5
8201 E. McDowell Road
Phoenix, Arizona

Engineering positions also available at Motorola, Inc. in Chicago, Illinois, and Riverside, California.

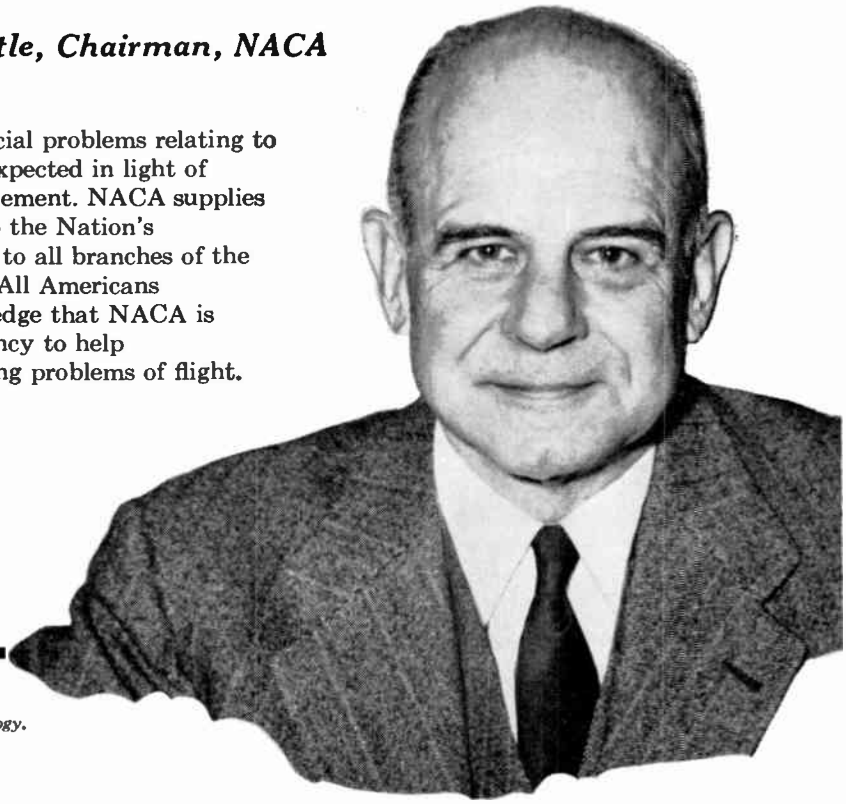
MOTOROLA, INC.

A message for young physical scientists & engineers

from James H. Doolittle, Chairman, NACA

Future breakthroughs on crucial problems relating to aircraft and missiles can be expected in light of NACA's long record of achievement. NACA supplies advanced research findings to the Nation's aircraft and missile industry, to all branches of the military, and to the airlines. All Americans can be assured by the knowledge that NACA is working with a spirit of urgency to help solve the current most pressing problems of flight.

James H. Doolittle



James H. Doolittle, Chairman, NACA;
Sc.D., Massachusetts Institute of Technology.

NACA has a staff of 7,750 research scientists and supporting personnel spread among centers on both Coasts and in Ohio. NACA staff members in pursuit of new knowledge have available the finest research facilities in the world, including several of the largest and fastest supersonic and hypersonic wind tunnels, hot jets, a fleet of full scale research airplanes, which will include the X-15, hypersonic ballistics ranges, shock tubes, a nuclear reactor establishment, rocket facilities, a research missile launching site, tracking devices, and the most advanced mechanical and electronic computers.

NACA Fields of Research Include: Aerodynamics, Aircraft and Missile Structures, Materials for Aircraft and Missiles, Automatic Stabilization, Propulsion Systems, Propulsion Systems Structures, Rocket Systems, Solid State Physics, Fuels, Instrumentation.

A number of staff openings are becoming available. You are invited to address an inquiry to the Personnel Director at any one or all four of the NACA research centers:

Langley Aeronautical Laboratory, Hampton, Virginia
Ames Aeronautical Laboratory, Mountain View, California
Lewis Flight Propulsion Laboratory, Cleveland, Ohio
High-Speed Flight Station, Edwards, California

(Positions are filled in accordance with the Aeronautical Research Scientist Announcement 1B)

NACA ***The Nation's Aeronautical Research Establishment***

Key Openings for Electronics Engineers

Electronics activities are broad and fast-growing at Chance Vought. Projects involve advanced guidance and control and fire control systems for missiles and high-performance manned aircraft. They begin with investigations and theory and progress through systemization and packaging to detailed hardware design. Key responsibilities await additional men who are qualified in these areas. Advanced degrees are preferred.

Stability and Control Engineer. E.E., M.E., or A.E. with emphasis on flight stability and control problems or dynamics. (Special consideration given graduate study or extensive experience in transients or closed loop stability analysis.) To assist in design of autopilot and control systems for high-performance missiles and aircraft.

Antenna Design Engineer. E.E. or Physics Degree with demonstrated aptitude for antenna design. To join active projects involving design of flush-mounted, recessed and external antennas at all frequencies for very high-performance aircraft and missiles.

Fire Control and Microwave Systems Engineer. Requires E.E. or Physics Degree; at least 2 years experience in radar, data link, or fire control systems; and strong ability in this work.

Test Equipment Engineer. Requires E.E. or Physics Degree and at least 2 years experience in this or related field. (Desirable: broad background in electronics design with emphasis on digital computers or microwave systems.) To join in the design of complete checkout systems for missiles and associated subsystems.

Guidance Design Engineer. E. E. or Physics Degree, plus 2 or more years experience. To design various active and self-contained missile guidance systems, and to design and develop radar beacons.

Reliability Analyst. Requires M.E., Physics, E.E., or Math Degree; broad knowledge of electronic and mechanical systems; experience in operations research or reliability. Helpful: statistical methods experience.

Electronic Packaging Engineer. M.E., E.E., or equivalent packaging design experience. To help design ground, airborne and shipboard electronic equipment for use in severe environments. Involves consideration of heat transfer, shock, vibration and other factors.

To arrange for a personal interview, or for more information on these or other current openings, return coupon to:

C. A. Besio
Supervisor, Engineering Personnel
CHANCE VOUGHT AIRCRAFT, Dept. V-1
Dallas, Texas

I am a ______ Engineer,
interested in the opening for ______
Name ______
Address ______
City ______ State ______

Positions Open

(*Continued from page 152A*)

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor with M.Sc. or Ph.D. degree. To teach communications or electronic courses and direct advanced degree candidates in communications or electronics beginning September 1, 1958. Opportunities for research. Salary depends upon qualifications. Write Chairman, Dept. of Electrical Engineering, University of Nebraska, Lincoln 8, Nebraska.

INSTRUCTOR OR ASSISTANT PROFESSOR

Instructor or Assistant Professor in electrical engineering beginning September, 1958. One qualified and interested in teaching fundamental undergraduate courses plus one graduate course. Ph.D. or MS. required with salary dependent on qualifications. City location with many industrial contacts. Apply Dept. of Electrical Engineering, St. Louis University, St. Louis 8, Missouri.

TECHNICAL CORRESPONDENT

Technical correspondent wanted by a very important radio electronics periodical for inquiries about latest American news. Necessary to write in French. Write, S.A.R.P., 81 rue de la Pompe, Paris (16eme) France.

SYSTEMS ENGINEERS

Good opportunity in systems integration for competent engineers with broad backgrounds in several of these fields: antenna design, propagation, tracking, data links, data transmission, radio relay links, telemetry, television, computers, missile instrumentation. Must work smoothly and effectively with people, think effectively and rapidly, speak fluently, and write clearly. Engineering effort includes generating new ideas, evaluating technical concepts, participating in technical meetings, coordination and monitoring, preparation of specifications and reports. Send resume to Lyman Nickel, Philco Western Development Labs., 3875 Fabian Way, Palo Alto, Calif.

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor of electrical engineering at state supported college located on Texas Gulf Coast. Excellent opportunity for man with MS. or Ph.D.; salary depending on background and qualifications. Permanent position in expanding department. Address reply to Director, Div. of Engineering, Texas College of Arts and Industries, Kingsville, Texas.

NEWS New Products

(*Continued from page 134A*)

Printed Circuit Design Brochure

Arthur Ansley Mfg. Co., New Hope, Pa., has prepared a new booklet ***Some Suggestions on Printed Circuit Layout and Design.*** It includes quite a lot of information not previously published, and is based on data compiled through improved manufacturing techniques achieved by this firm. It is available at $1.00 per copy.

(*Continued on page 158A*)

PAIR OF UNIVACS SOLVE TOUGH MISSILE DESIGN PROBLEMS

*A report to
Engineers and Scientists
from
Lockheed Missile Systems
where expanding
missile programs insure more
promising careers*

Two Univac Scientifics today aid preliminary design work for Lockheed missiles by solving tough flight simulation problems at the Division's research and development laboratories in Palo Alto, California.

These high speed digital computers aid in the study of missile characteristics, performing scientific and engineering calculations and data reduction by means of the most advanced techniques.

The $1½ million computers are part of an installation which is one of the largest and most complete in the West.

Advanced facilities such as this are the result of constant refining and improvement as our missile programs continue to expand. Projects like the Polaris, X-7 ramjet and Q-5 target missiles – and our Earth Satellite for example – have already placed Lockheed in the forefront of U.S. missile developers. Moreover, by joining our young Division at this stage in its development, you increase your opportunities to grow rapidly with us.

If you are a qualified engineer or scientist, you are invited to inquire today about positions open in the following fields:

Information Processing, Telecommunications, Reliability-Producibility, Ground Support, Guidance, Flight Controls, Aerodynamics, Thermodynamics.

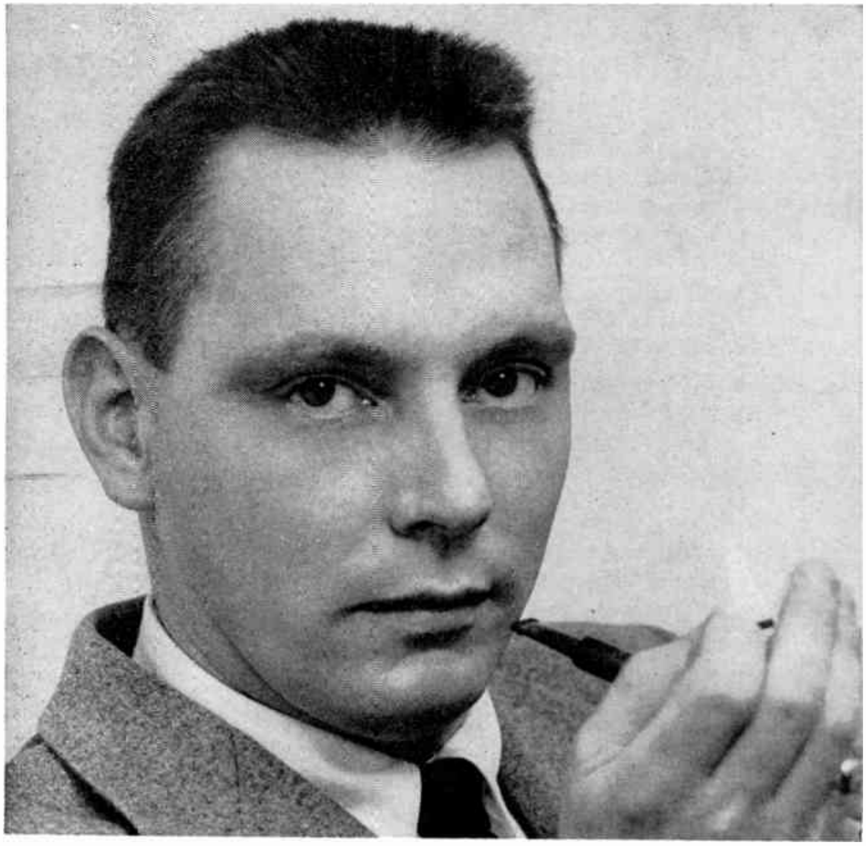
Write Research and Development Staff, Palo Alto 12, California

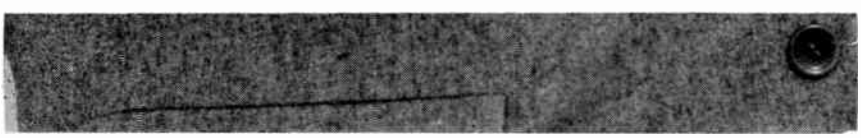
Lockheed MISSILE SYSTEMS

A DIVISION OF LOCKHEED AIRCRAFT CORPORATION

PALO ALTO • SUNNYVALE • VAN NUYS • SANTA CRUZ • CALIF.
CAPE CANAVERAL, FLORIDA • ALAMOGORDO, NEW MEXICO

Shown with the Univac is Dr. J. P. Nash, Information Processing Division Manager, second from right, examining results of a trajectory study with J. P. Ryan, Jr. Left rear, J. B. Potter; right, Byron Zurcher, liaison engineer.





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 156A)

Range Marker Generator

An unusual Range Generator is being manufactured by **Missouri Research Laboratories,** 2109 Locust St., St. Louis, Mo. It is designed to ease the task of calibrating radars and other time based equipment.

The task of laboriously counting time marks is made unnecessary by producing at one output terminal three separately identifiable sets of time marks at increments of 500; 2,000; and 10,000 yards. Although normally used mixed, any of the three sets is separately available at another terminal to operate auxiliary equipment.

The Range Marker Generator also features a Marker Phase Control to advance or recede the marks up to 500 yards for ease in indexing.

A trigger generator is included to provide PRF from 200 to 2,000 cps; pulse width of 0.5 microseconds for dependable triggering; and an amplitude of 0–50 volts positive or negative.

Marker phases have a width of 0.2 microseconds and are variable in amplitude up to 30 volts plus or minus. The temperature controlled crystal gives a range accuracy of 0.1 per cent. Both Marker Generator and Trigger Generator have a 100 ohm output.

The unit is available either in its own box, or for rack mounting.

Bi-Directional Couplers

The Microwave Electronics Div., Sperry Gyroscope Co., Great Neck, N.Y. has developed a new series of bi-directional couplers. Each consists of two opposing directional couplers mounted in a single waveguide unit. The three types cover all frequencies from 2.6 to 12.4 kmc. They were developed as a precision component for continuous measurement of impedance and power comparisons in microwave systems. Each has application in radar and communication monitor units, also in

(Continued on page 160A)

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 158A)

system test equipment and in laboratory measurement setups. Each operates in combination with ratiometers, comparators, barretter mounts and other detectors.

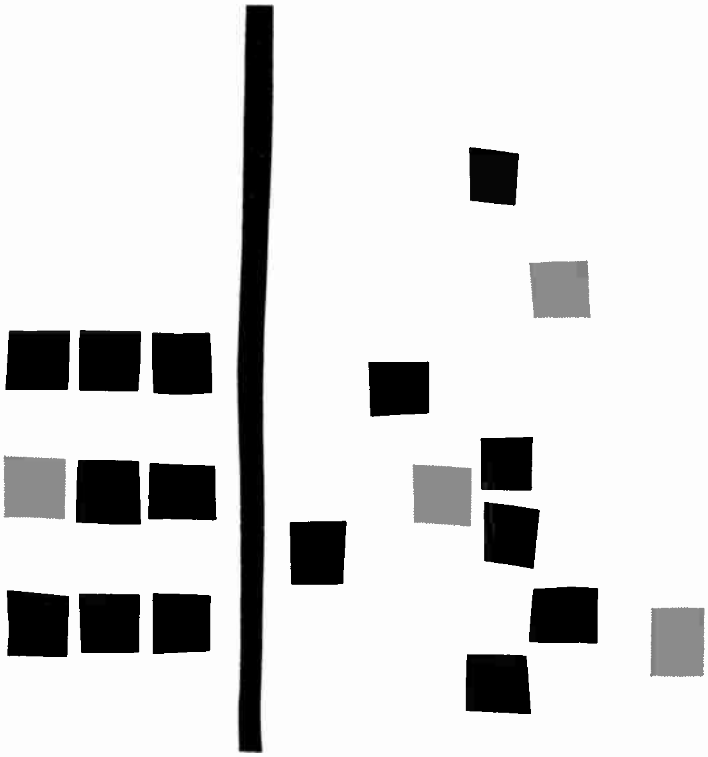
The frequency range of the Type 604, "S-Band" is 2.6–3.95 kmc; the 605, "C-Band" is 3.95–6.0 kmc; and the 606 "X-Band" is 8.2–12.4 kmc. Each has a nominal coupling ratio of 10.0 db; a coupling variation of ±0.5 db; and a minimum effective directivity of 40 db.

Transistorized FM Telemetering Transmitter

The Model TR-10, a hybrid unit incorporating the best features of both transistors and vacuum tubes is a new transistorized FM telemetering transmitter, designed by **United Electrodynamics,** Dept. S, 1200 S. Marengo Ave., Pasadena, Calif. All frequency-determining and low-level stages are completely transistorized to eliminate incidental FM noise produced by tube elements under vibration. Rugged subminiature tubes are used in the output stages.

(Continued on page 164A)

11-19



THE ELEMENT OF FREEDOM

and the System Designer

Freedom is doing what you like. Some System Designers like best to match their wits and skills against difficult problems. This characteristic (or idiosyncrasy) of liking complicated technical problems is one of the chief qualifications of the engineer we need.

The advanced nature of the assignments requires at least four years' experience, including circuit design or development, equipment construction and a knowledge of logical design for computers. Moreover, the position also calls for a sound background in computer programming, the ability to write programs and familiarity with the IBM 700 series or similar single address parallel machines.

You are invited to write for more information or phone collect. Address R. W. Frost, System Development Corporation, 2418 Colorado Avenue, Santa Monica, Calif.; phone EXbrook 3-9411.

SYSTEM DEVELOPMENT CORPORATION

An independent nonprofit organization, formerly a division of the Rand Corporation

ELECTRONIC ENGINEERS

needed at

MARTIN

New long-term developments at Martin in the field of electronics have created exceptional opportunities for top electronic engineers. At least 5 years experience required. Salaries from $9,000 to $12,000.

Openings in these areas:

- Circuit Design
- Systems
- Inertial Guidance
- Countermeasures
- Digital Computers
- Test Equipment Design

WRITE TO:

William Spangler, Manager
Professional Employment
The Martin Company
Baltimore 3, Md.

MARTIN
BALTIMORE

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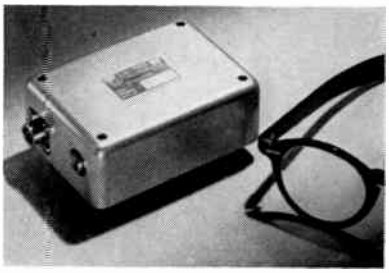
(Continued from page 160A)



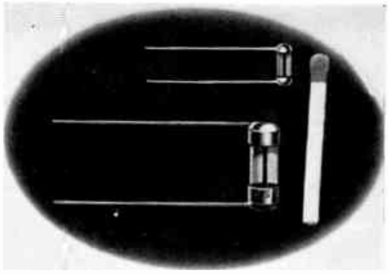
The unit accepts inputs from standard IRIG subcarrier oscillators (channels 1–18 and A–E). The power output is 2 watts in the 215–245 mc telemetering band. It is crystal stabilized and designed for operation in accordance with IRIG standards.

The Model TR-10 Transmitter measures only 1.5×2.87×4.25 inches and weighs less than 16 ounces. The unit is designed for missile environments and features light weight, low power consumption and extremely rugged construction.

For information, write to United Electrodynamics, 1200 South Marengo Ave., Pasadena, Calif., or call SY 9-7161, Dept. S.

Sub-Miniature Hermetically Sealed Fuses

Fuses of minute physical dimensions for use with miniaturized circuits, controls, electronic devices, and electrical equipment are currently available from **Bussmann Mfg. Division, McGraw-Edison Co.,** University at Jefferson, St. Louis 7, Mo. Made of hermetically sealed glass tubes with lead-ins, these fuses meet requirements for potting and encapsulating. The hermetic seal prevents potting material from seeping into the fuse case and interfering with dependable operation of the fuse. The fuses are designed to withstand heavy shocks and vibrations



They are available in two sizes: 0.140×0.300 inch, in an amperage range of 1/20 to 1/2; rated at 125 volts. 1/4×5/8 inch, in an amperage range of 1 to 5; rated at 32 volts.

The larger size has ⅛ inch ferrules, and can be furnished with or without lead-ins. For complete information, write to Bussmann.

(Continued on page 166A)

Engineers and scientists . . . grow with Stromberg-Carlson

Fast-growing division of General Dynamics

If you qualify, we can offer you challenging, important assignments in some of today's most fascinating areas of electronics.

Many of these openings are senior positions, including opportunities for group leaders and section heads. We're particularly interested in men with substantial experience in solid state circuitry for applications in the most advanced state of the art.

Our business is well balanced between commercial and military products. To you this means *stability* as well as rapid growth.

These are some of the areas of work where you may find your greatest challenge:

- Missile Electronics
- Automatic Test Equipment
- Air Traffic Control
- Countermeasures
- Telecommunications
- Nucleonics
- Electroacoustics

You would be located in Rochester, N. Y., a beautiful, progressive city in the heart of the Upstate vacationland. Invigorating four-season climate; variety of educational, recreational and cultural facilities unsurpassed anywhere in the U. S.

To arrange for a personal interview, send complete résumé to: FRED E. LEE, *Manager of Technical Personnel.*

S-C STROMBERG-CARLSON

STROMBERG-CARLSON

A DIVISION OF GENERAL DYNAMICS CORPORATION

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Electronic and communication products for home, industry and defense

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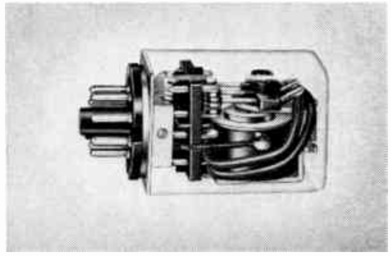
NEWS New Products

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

(Continued from page 164A)

Sealed Relay



The Series 1200 relay is now available in a new version called the 1210 Series Relay from **Guardian Electric Mfg. Co., 1621** W. Walnut St., Chicago 12, Ill. It is encased in a unique, dust-tight, transparent plastic enclosure that withstands heavy impacts. The unit is unaffected by weather and resists temperatures up to +200°F. Contact rating is: 8 amperes; 115 volts, 60 cps, noninductive. The series 1210 relay is available in any ac voltage 6 to 230 and in any dc voltage 6 to 110. Corrosion proof and sturdy, this dependable relay will give long-lasting, efficient service. For complete details write to the company.

(Continued on page 168A)

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 166A)

Transistor Tester

The Model 960 portable transistor and crystal diode tester for industrial and service maintenance use, was announced by the **Precision Apparatus Company, Inc.,** Glendale, L.I., New York.

Designed in accordance with the recommendations of leading transistor manufacturers, the Model 960 is the only semiconductor tester in its price class that gives comprehensive test for I_{cbo}, gain, leakage, shorts, etc., on low, medium and high-power transistors of both the *p-n-p* and the *n-p-n* types, as well as the new tetrode transistors.

Operating specifications of the Model 960 are: Direct I_{cbo} Readings, in terms of true collector current, on wide-angle 5½ inch, 100 microampere PACE meter.

Five I_{cbo} Ranges, cover all types of transistors—low, medium and high power.

Wide Range of Collector Potentials from 0.5 volts dc to 100 volts dc in 17 selected steps.

Direct Reading Gain Ranges with Five separate injection currents for low, medium and high power types.

Leakage Tests check emitter to collector current at fixed collector bias, for maximum accuracy.

All transistor test settings on high speed roller chart. Patchcord Element Selector System provides for accommodation of future semiconductor releases.

Net priced at $89.00, the Model 960 is a complete, self-contained, ac operated unit. No batteries are used or required. The case measures 18×10½×6¼ inches.

DC-AC Chopper

A revolutionary new concept in dc-ac choppers, permitting micro-miniaturization, without sacrifice to contact rating, ruggedness, reliability, or long life, is announced by **Rawco Instruments, Inc.,** 3527 W. Rosedale, Fort Worth 7, Texas.

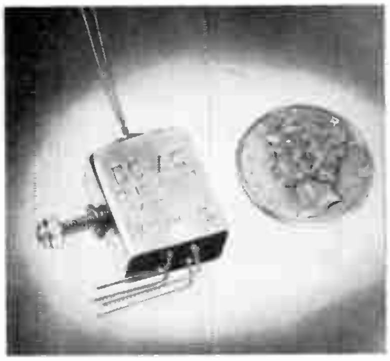
Weighing less than ¼ ounce with hardware, it has a versatile new encasement (35×0.53×0.65 inches) that can be mounted in any position, to either circuit board, or metal chassis, providing leads accessible for ease of installation and proper inspection.

(Continued on page 170A)

NEWS New Products

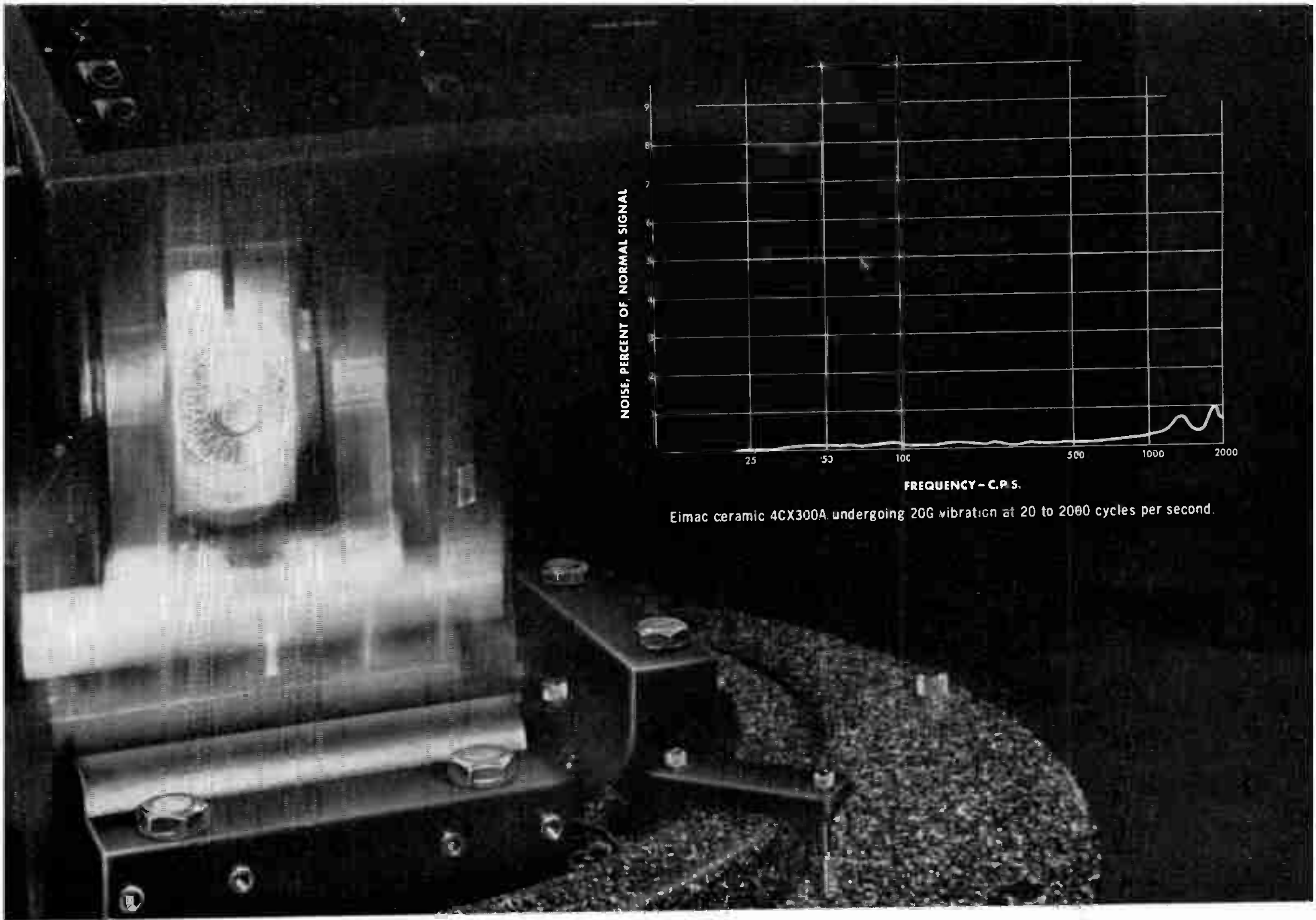
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 168A)



Designed for Military Application, with a temperature range of $-65°C$ to $+125°C$, its true hermetic seal, and balanced shock mounted armature, leave it virtually unaffected by shock, vibration, and environmental conditions. An extremely low noise factor (less than 10 micro volts (rms) across 1 megohm impedance) is accomplished by exceptional shielding and low ac requirements. A unique make-after-break or make-before-break, single-pole-double-throw switch configuration eliminates thermal EMF's,

(Continued on page 172A)



Eimac ceramic 4CX300A undergoing 20G vibration at 20 to 2000 cycles per second.

Third in a series describing the advantages of ceramics in electron tubes. Previously discussed: impact and heat.

Surviving Vibration is an Eimac Ceramic Tube Extra

High reliability in severe environments is an important vacuum tube requirement in many aeronautical applications. An important aspect of this reliability is a tube's ability to operate under extreme vibration without envelope damage, introducing noise or developing interelectrode short circuits. Eimac ceramic design incorporates many advanced features that improve tube performance under these conditions.

In the illustration an Eimac ceramic 4CX300A, 300 watt tetrode, is being operated in a circuit while undergoing 20G vibration at 20 to 2000 cycles per second. The exceptionally low noise level, produced under these conditions shown in the graph above, remains less than 1% of normal signal over the entire test range.

Other advantages of Eimac ceramic tubes are: resistance to damage by shock or high temperature; compactness without sacrifice of power; ability to withstand rigorous processing techniques that lead to high tube reliability, uniformity and longevity.

In its new line of ceramic tubes, Eimac has the answer for the aeronautical engineer who needs a tube that will deliver full output under extreme environment.

Write our Application Engineering Department for a copy of the new explanatory booklet "Advantages of Ceramics in Electron Tubes"

EITEL-McCULLOUGH, INC.
SAN BRUNO · CALIFORNIA

Eimac First **with Ceramic Tubes that can take it**

Eimac®

PRODUCTS DESIGNED AND MANUFACTURED BY EIMAC

- Negative Grid Tubes
- Reflex and Amplifier Klystrons
- Ceramic Receiving Tubes
- Vacuum Tube Accessories
- Vacuum Switches
- Vacuum Pumps

Includes the most extensive line of ceramic electron tubes

PROFESSIONAL CONSULTING SERVICES

NEWS New Products

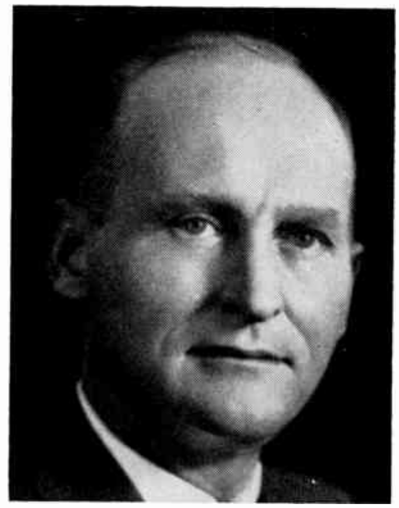
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 170A)

generated between contact material and leads into casement.

With a driving range of 0 to 1800 cps, this chopper is available in three series: R-20, with a null switching configuration; R-70, with a normally closed or open switching configuration; R-80, with snap action switching. All the above series can be supplied with a coil center tap for biasing and contact material for low or high current switch requirements, (up to 10 volts, 1 ma).

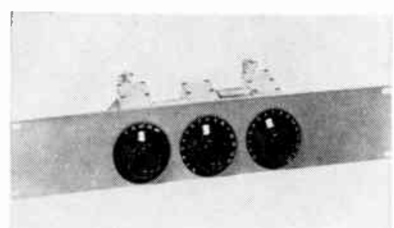
Pool VP For Cinch



E. J. Pool was elected Executive Vice-President at the recent Directors meeting of the **Cinch Manufacturing Corp.**, 1024 S. Homan St., Chicago, Ill., manufacturers of electronic components. He continues as General Manager, a position he has held for the past four years.

RF Attenuators

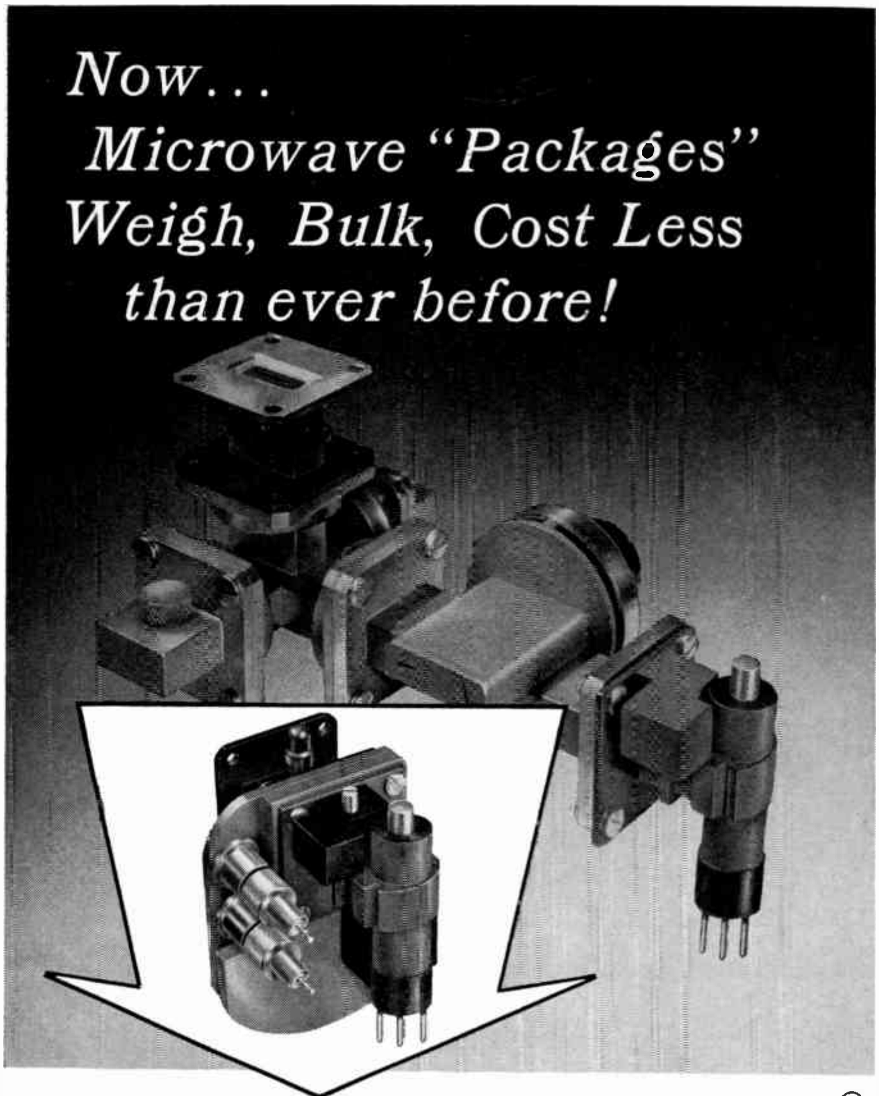
The series VA attenuators developed by **Ortho Filter Corp.**, Paterson, N.J., are extremely accurate variable rf Attenuators which operate from dc to above 250 mc. Standing wave ratio at 100 mc is less than 1.01.



Single rotating units are available at 0.1, 1 or 10 db per step. Multiple unit assemblies are available in various combinations and are supplied mounted on a standard $3\frac{1}{2} \times 19$ inch panel with front or rear access connections. The maximum change in loss from 0 to 100 mc is 0.1 db.

Advertising Index

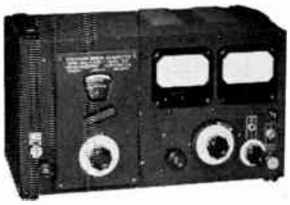
DISPLAY ADVERTISERS



Advertising Index

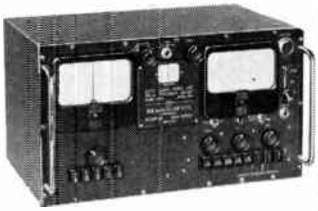
Advertising Index

MODEL 65-B Standard Signal Generator



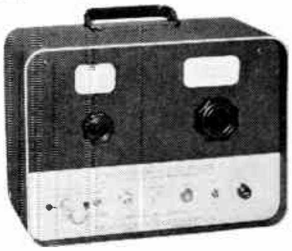
MODELS 80 and 80-R Standard Signal Generators

MODEL 210 Series FM Standard Signal Generators



MODEL 58-AS VHF Radio Noise and Field Strength Meter

MODEL 72 Square Wave Generator



MODEL 202-C Standard Barretter Bridge

MODEL 59 Megacycle Meter

MODEL 179 Standard Pulse Generator

MODEL 162 Vacuum Tube Voltmeter

MEASUREMENTS' Electronic Instruments for Research, Development and Design

STANDARD SIGNAL GENERATORS

MODEL	FREQUENCY RANGE	PRICE
65-B	75 Kc to 30 Mc	$875.00
80	2 Mc to 400 Mc	590.00
80-R	5 Mc to 475 Mc	625.00
82	20 cps to 200 Kc 80 Kc to 50 Mc	660.00
84-TVR	400 Mc to 1000 Mc	785.00
95	50 Mc to 400 Mc	1800.00
210 Series	25 Mc to 480 Mc	450.00 to 475.00

VHF RADIO NOISE AND FIELD STRENGTH METER

MODEL	FREQUENCY RANGE	PRICE
58-AS	15 Mc to 150 Mc	$925.00

SQUARE WAVE GENERATORS

MODEL	FREQUENCY RANGE	PRICE
71	6 cps to 100,000 cps	$195.00
72	5 cps to 5 Mc	248.00

MEGACYCLE "GRID-DIP" METERS

MODEL	FREQUENCY RANGE	PRICE
59-LF	100 Kc to 4.5 Mc	$168.00
59	2.2 Mc to 420 Mc	168.00
59-UHF	420 Mc to 940 Mc	198.00

VACUUM TUBE VOLTMETERS

MODEL	FREQUENCY RANGE	PRICE
67	5 to 100,000 sine-wave cps	$235.00
162	27 cps to over 150 Mc	210.00

CRYSTAL CALIBRATORS

MODEL	FREQUENCY RANGE	PRICE
111	250 Kc to 1000 Mc	$ 97.50
111-B	100 Kc to 1000 Mc	110.00

STANDARD BARRETTER BRIDGE

MODEL	FREQUENCY RANGE	PRICE
202-C	2 Mc to 1000 Mc	$375.00

STANDARD PULSE GENERATOR

MODEL	FREQUENCY RANGE	PRICE
179	60 cps to 100,000 cps	$365.00

This is only a partial listing of Measurements' products. Write for our complete Catalog.

Laboratory Standards

MEASUREMENTS
A McGraw-Edison Division
BOONTON, NEW JERSEY

Advertising Index

Our reputation as the world's most Consistently Dependable producer of capacitors has been maintained for over 46 years. But *any* reputation can be lost overnight. That's why we resist the temptation to gain temporary advantage through methods that risk *our* reputation or yours. C-D's Consistently Dependable products can mean PLUS dollars to you.

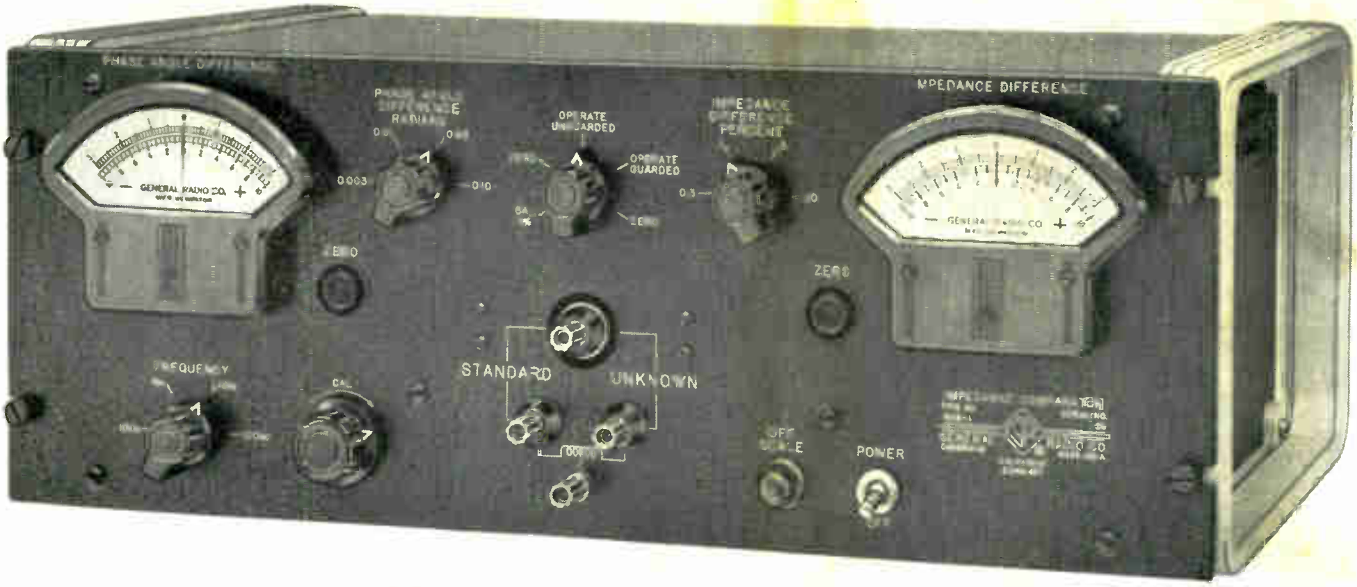
Widest Choice of Impregnants and Dielectrics to meet your needs: More than a score of liquid and solid impregnating media and dielectrics, including Polystyrene, Mylar*, Teflon, metallized paper and metallized Mylar, are readily available to meet your temperature, size and other circuit requirements. Operating temperature ranges from −40°C to +85°C and −60°C to +200°C. Whatever your capacitor problems, depend on Cornell-Dubilier to fulfill your needs most promptly, most economically and most satisfactorily.

Write for catalog to Cornell-Dubilier Electric Corporation, South Plainfield, New Jersey.

(*DUPONT TM)

CONSISTENT HI-DEPENDABILITY
CORNELL-DUBILIER CAPACITORS

SOUTH PLAINFIELD, N. J.; NEW BEDFORD, WORCESTER & CAMBRIDGE, MASS.; PROVIDENCE & HOPE VALLEY, R. I.; INDIANAPOLIS, IND.; SANFORD, FUQUAY SPRINGS & VARINA, N. C.; VENICE, CALIF.; & SUB.: THE RADIART CORP., CLEVELAND, OHIO; CORNELL-DUBILIER ELECTRIC INTERNATIONAL, N. Y.



A Measurement Once Every Second with the GR® IMPEDANCE COMPARATOR

The G-R Type 1605-A Impedance Comparator serves as the "brain" in a Semi-Automatic Tester designed for checking components on printed-circuit subassemblies.

The Comparator indicates directly on two panel meters the differences in magnitude and phase angle between unknown and standard impedances. No manual balancing is required.

As used with the Semi-Automatic Tester, the Comparator's panel meters are disconnected. Comparator metering voltages, proportional to impedance-magnitude difference (in per cent) and phase-angle difference (in radians), are amplified by the Tester and compared against d-c reference voltages which correspond to allowable tolerances. Printed-circuit components which produce voltages in excess of pre-set tolerances are automatically rejected.

Relays have been added in the Tester so that Comparator impedance ranges can be switched automatically by a remote punched-card programmer. Since programs are switched every second, three continuously running oscillators are used as fixed-frequency sources for the Comparator.

The Unit was constructed by Bendix Radio Division under sub-contract for IBM and the United States Air Force.

Why The GR Impedance Comparator Was Selected For Automatic Sorting:

★ Indicates *both* impedance magnitude and phase angle without knob manipulation.

★ D-C voltages proportional to percentage deviation from standard are provided.

★ No excess switches or complex controls.

★ Excellent guard circuitry permits long cable runs, which are usually necessary in automatic equipment.

★ Wide impedance range, high measurement accuracy.

★ Constructed for long, reliable service.

★ Practical size and weight.

COMPARATOR SPECIFICATIONS

Wide Range of Internal-Test Frequencies
100, 1000, 10,000, and 100,000 cps.

Impedance Ranges
Resistance: 2 ohms to 20 megohms
Capacitance: 40 $\mu\mu f$ to 500 μf
Inductance: 20 μh to 10,000 μh

Direct-Reading Meter Ranges
Impedance-Magnitude Differences:
0.3%, 1%, 3%, and 10% of full scale.
Phase-Angle Differences:
0.003, 0.01, 0.03, 0.1 radians full scale.

Over-All Accuracy
3% of full scale (0.01% over-all accuracy on 0.3% impedance magnitude range).

Price: $790

WE SELL DIRECT. Our District Sales Offices are staffed by engineers especially trained to help you in the selection of instruments and measuring systems best suited to your needs. We welcome your inquiries — will help solve your problems.

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