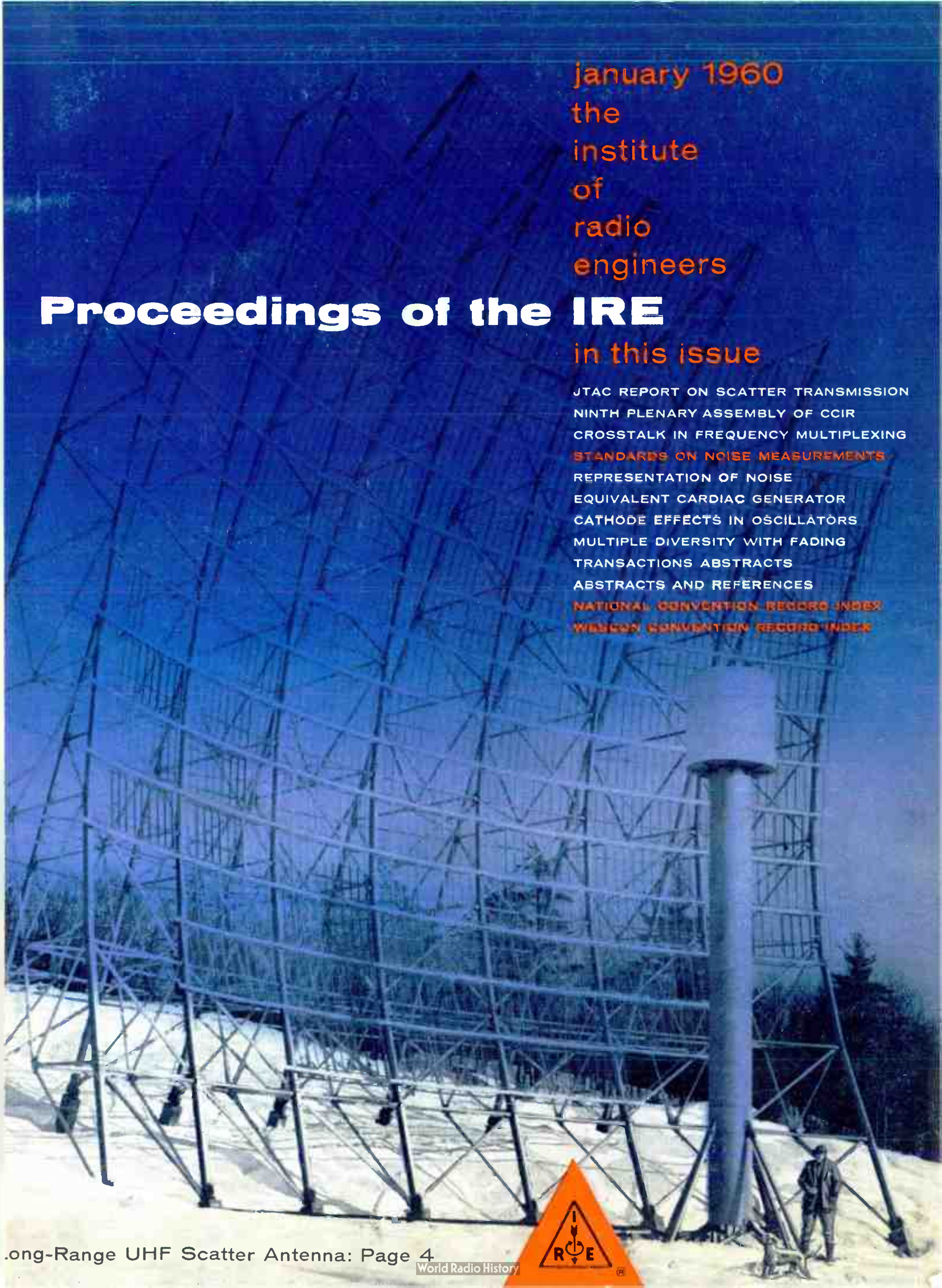
january 1960

the institute of radio engineers

Proceedings of the IRE

in this issue



Long-Range UHF Scatter Antenna: Page 4

For Your Special Applications

The bulk of UTC production is on special units designed to specific customers' needs. Illustrated below are some typical units and some unusual units as manufactured for special applications. We would be pleased to advise and quote to your special requirements.

FILTERS

All types for frequencies from .1 cycle to 400 MC.

400 ~ telemetering, 3 db at ± 7.5%, 40 db at 230 and 700 ~, 5/8 x 1 1/4 x 2".

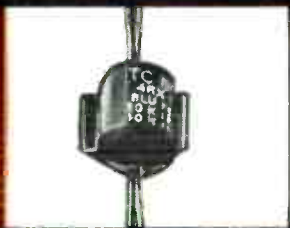
15 ~ BP filter, 20 db at 30 ~, 45 db at 100 ~, phase angle at CF less than 3° from —40 to + 100° C.

LP filter within 1 db to 49 KC, stable to .1 db from 0 to 85° C., 45 db at 55 KC.

LP filter less than .1 db 0 to 2.5 KC, 50 db beyond 3 KC.

HIGH Q COILS

Toroid, laminated, and cup structures from .1 cycle to 400 MC.



Tuned DO-T servo amplifier transformer, 400 ~ .5% distortion.

Toroid for printed circuit, Q of 90 at 15 KC.

Dual toroid, Q of 75 at 10 KC, and Q of 120 at 5 KC.

HVC tapped variable inductor for 3 KC oscillator.

SPECIALTIES

Saturable reactors, reference transformers, magnetic amplifiers, combined units.

RF saturable inductor for sweep from 17 MC to 21 MC.

Voltage reference transformer .05% accuracy.

Multi control magnetic amplifier for airborne servo.

Input, output, two tuned interstages, peaking network, and BP filter, all in one case.

PULSE TRANSFORMERS

From miniature blocking oscillator to 10 megawatt.

Wound core unit .01 microsecond rise time.

Pulse current tansformer 100 Amp.

Pulse output to magnatron, bifilar filament.

Precise wave shape pulse output, 2500 V 3 Amps.

POWER COMPONENTS

Standard and high temperature . . . hermetic, molded, and encapsulated.

Multi-winding 140 VA, 6 KC power transformer 1 1/4 x 1 1/4 x 1"

200° C. power transformer, 400 ~, 150 VA.

400 ~ scope transformer, 20 KV output.

60 ~ current limiting filament transformer, Sec. 25 Mmfd., 30 KV hipot.

UNITED TRANSFORMER CORPORATION

150 Varick Street, New York 13, N. Y. • EXPORT DIVISION: 13 E. 40th St., New York 16, N. Y.,
CABLES: "ARLAB" PACIFIC MFG. DIVISION, 4008 W. Jefferson Blvd., Los Angeles, Cal.

VOLUME 48, NUMBER 1

January, 1960

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE

contents

COVER

This giant 120-foot parabolic antenna in Westford, Mass., is part of a long-range tropospheric scatter system developed for the Air Force by the M.I.T. Lincoln Laboratory, making possible UHF communication over a distance of 640 miles. A JTAC report on scatter communications appears on page 4.

JANUARY

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE®

continued

PROCEEDINGS OF THE IRE, published monthly by The Institute of Radio Engineers, Inc. at 1 East 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.
Thirty days advance notice is required for change of address. Price per copy: members of the Institute of Radio Engineers, one additional copy $1.25; non-members $2.25. Yearly subscription price: to members $9.00, one additional subscription $13.50; to non-members in United States, Canada, and U. S. Possessions $18.00; to non-members in foreign countries $19.00. Second-class postage paid at Menasha, Wisconsin under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927. Printed in U.S.A.

Bernard R. Weisfeld of our Department of Special Systems and Components discusses application of a well known phase-shift circuit to achieve image rejection in microwave receivers.

Preselection Without Filters

Though it offers many advantages, the superheterodyne receiver suffers from "image reception." The image is a received signal occupying a position on the spectrum on the side of the local-oscillator frequency opposite to that of the desired signal, and symmetrical thereto. In high-frequency (microwave) receivers where the intermediate frequency is very much lower than the incoming frequency, the design engineer faces the problems of double reception and false frequency identification. Double reception is the response of the receiver to the same signal at two different tuning positions of the local oscillator; false frequency identification results because the receiver will respond with as much sensitivity to the image frequency as it does to the desired signal. These difficulties are generally overcome by means of a preselection filter, which passes the desired signal frequency without attenuation while rejecting the image frequency sufficiently to avoid confusion at the output.

In practice this preselection filter presents problems of tracking with the local oscillator, avoiding attenuation of the desired signal, and keeping all the elements of the filter tuned properly with respect to one another.

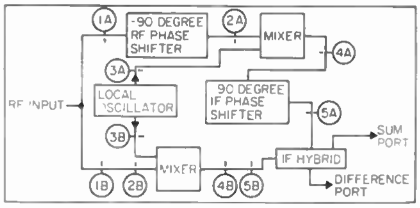


Figure 1. Image-Rejector Receiver

At AIL, we have developed a receiver that eliminates image reception without the use of a preselection filter. The circuit is shown in Figure 1. The circuit is essentially the same as the phase-shift circuit for single-sideband reception. Conceived in the 1930's, this circuit has been in common use at low frequencies. It has now been successfully applied to achieve image injection in the microwave regions.

The theory of operation can be seen in the following equations. Let the input signal be $E \sin \omega_s t$. The signal is equally divided, producing

at 1A: $k_1 E \sin w_s t$

at 1B: $k_1 E \sin \omega_s t$

at 2A: $k_1 E \sin (\omega_s t - 90°) = k_1 E \cos \omega_s t$

at 2B: $k_1 E \sin \omega_s t$

The local oscillator feeds both mixer signals in phase. Thus the voltages at 3A and 3B are equal to $k_2 \sin \omega_{LO} t$. Considering only the difference signal output of the mixer, we find

at 4A: $k_3 E \sin |\omega_{LO} - \omega_s| t$ when $\omega_{LO} > \omega_s$

and $-k_3 E \sin |\omega_{LO} - \omega_s| t$ when $\omega_{LO} < \omega_s$

and at 4B: $k_3 E \cos (\omega_{LO} - \omega_s)$

at 5A:

when $\omega_{LO} > \omega_s$,

$$+k_3 E \sin [|\omega_{LO} - \omega_s| t - 90°] = k_3 E \cos (\omega_{LO} - \omega_s) t$$

and when $\omega_{LO} < \omega_s$,

$$-k_3 E \sin [|\omega_{LO} - \omega_s| t - 90°] = -k_3 E \cos (\omega_{LO} - \omega_s) t$$

at 5B: $k_3 E \cos (\omega_{LO} - \omega_s)$

When $\omega_{LO} > \omega_s$, the voltages feeding the IF hybrid are equal and in phase; thus, they appear at the sum port. When $\omega_{LO} < \omega_s$, the voltages are equal and out of phase; thus, they appear at the difference port. Consequently, the desired signal is present at one output port of the hybrid and the image is present at the other port.

In theory perfect separation of these signals is possible, but in practice the rejection of the image is limited by the quality of the components. The imperfection of power split at the input, deviation of the phase shifters from 90 degrees, crystal dissimilarities, and the IF hybrid isolation all contribute to imperfect suppression of the image. An image-rejection receiver was constructed to operate over a 12-percent region of X-band. This receiver had better than 20-db rejection of the image. Over narrow-band regions, rejections as great as 40 db were achieved.

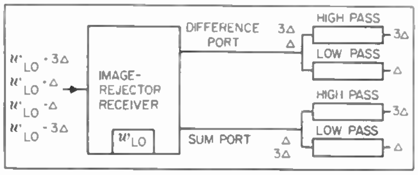


Figure 2. Multiplexer Receiver

This circuit can also be used in multiplexing applications where the signal frequencies to be separated are relatively close to each other. Figure 2 illustrates the separation of four signals equally spaced from each other. The two signals at the difference port result from the inputs above ω_{LO}; those at the sum port result from inputs below the ω_{LO}. Each of these pairs of signals are easily separated at the IF level with high-pass and low-pass filters. If desired, the resulting signals can be heterodyned again so that identical-frequency IF amplifiers can be used. This technique can be further extended to achieve a higher degree of multiplexing.

The engineer can utilize this circuit in many applications to eliminate difficult receiver filter problems. Where the image-rejection requirements are more stringent than can be met by this technique, the circuit can still be advantageously used to relax the selectivity specification of the preselection filter.

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

Airborne Instruments Laboratory

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Now . . . *for the first time* . . . an extended-life electrolytic in miniature tubular case styles. Sprague's New Type 40D Extended-Life Electrolytics are designed to give more than 10 years of service under normal operating conditions in actual circuit applications.

○ Broader Application

Though similar in many respects to Sprague's famous extended-life electrolytics for telephone and communications systems, these capacitors have the added advantage of low temperature characteristics previously unavailable in an aluminum electrolytic. As a result, Type 40D offers much broader industrial and military application.

○ Special Construction

Type 40D capacitors are specially constructed to assure freedom from open circuits even after extended periods of operation in the millivolt signal range. Ultra-low leakage currents are the result of special design and processing techniques based on the use of the highest purity anode and cathode foils.

○ Hand or Machine Assembly

For applications which require an insulated case, Sprague furnishes an outer insulation of either flexible plastic for hand assembly or rigid phenolic for machine insertion on printed wiring boards.

Get details on Type 40D Extended-Life Electrolytics by writing for Bulletin 3205 to Technical Literature Section, Sprague Electric Co., 235 Marshall St., North Adams, Mass.

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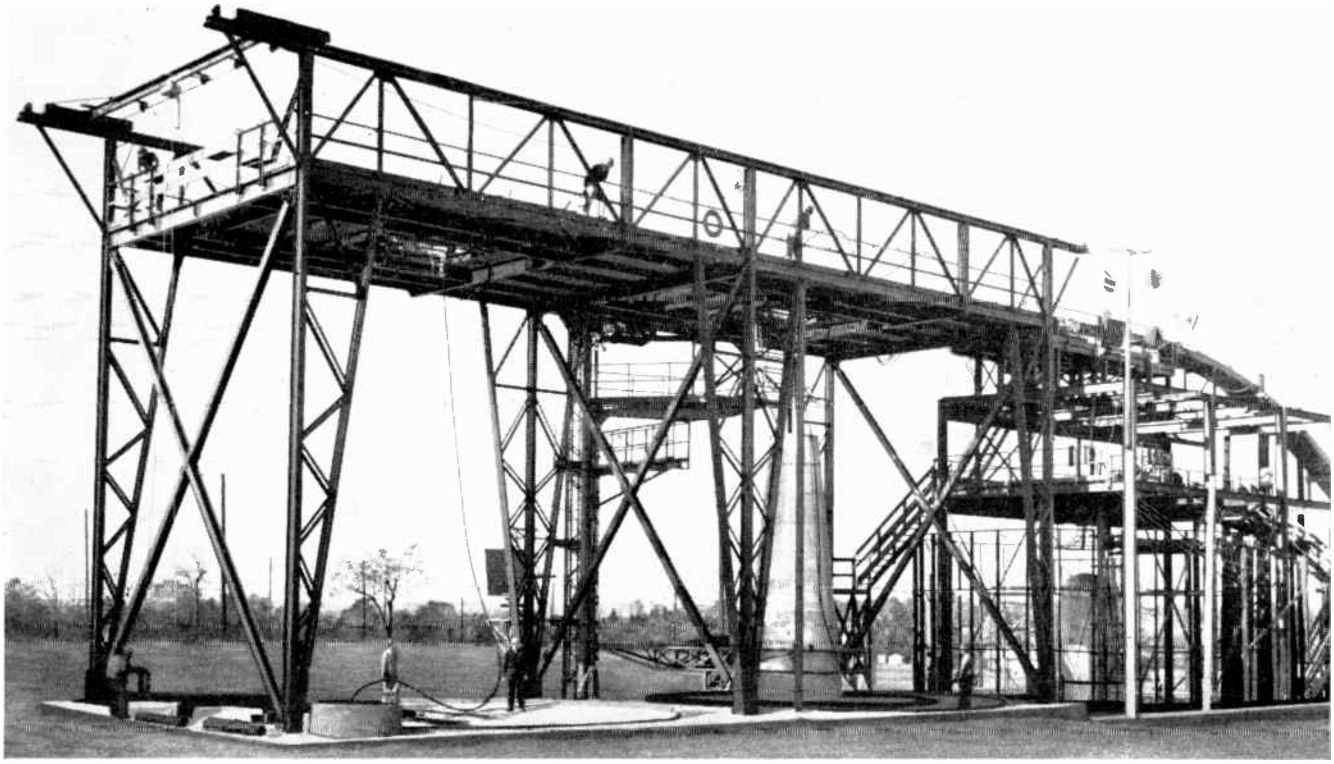


SHIP WITHOUT AN OCEAN

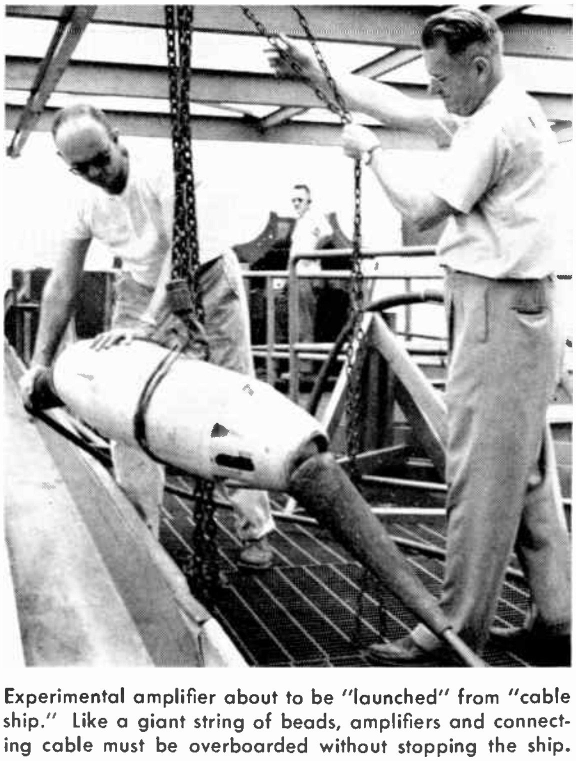
How do you lay a cable on the ocean floor—a cable that is connected to scores of large, heavy amplifiers? How do you "overboard" such a system in a continuous operation, without once halting the cable ship?

Bell Telephone Laboratories engineers must answer these questions in order to lay a new deep-sea telephone system designed to carry many more simultaneous conversations. They're experimenting on dry land because it is easier and more economical than on a ship. Ideas that couldn't even be attempted at sea are safely tested and evaluated.

In one experiment, they use a mock-up of the storage tank area of a cable ship (above). Here, they learn how amplifiers (see photo right), too rigid and heavy to be stored with the cable coils *below* decks, must be positioned *on* deck for trouble-free handling and overboarding.

Elsewhere in the Laboratories, engineers learn how best to grip the cable and control its speed, what happens as the cable with its amplifiers falls through the sea, and how fast it must be payed out to snugly fit the ocean floor. Oceanographic studies reveal the hills and valleys which will be encountered. Studies with naval architects show how the findings can be best put to work in actual cable ships.

This work is typical of the research and development effort that goes on at Bell Laboratories to bring you more and better communications services.



Experimental amplifier about to be "launched" from "cable ship." Like a giant string of beads, amplifiers and connecting cable must be overboarded without stopping the ship.

BELL TELEPHONE LABORATORIES

WORLD CENTER OF COMMUNICATIONS RESEARCH AND DEVELOPMENT

SPRAGUE® RELIABILITY
in these two dependable wirewound resistors

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VITREOUS-ENAMEL POWER RESISTORS

Sprague's new improved construction gives even greater reliability and higher wattage ratings to famous Blue Jacket miniature axial lead resistors.

A look at the small *actual sizes* illustrated, emphasizes how ideal they are for use in miniature electronic equipment with either conventional wiring or printed wiring boards.

Get complete data on these dependable minified resistors, **write for Engineering Bulletin 7410.**

TAB-TYPE BLUE JACKETS: For industrial applications, a wide selection of wattage ratings from 5 to 218 watts are available in Sprague's famous Tab-Type Blue Jacket close-tolerance, power-type wirewound resistors. Ideal for use in radio transmitters, electronic and industrial equipment, etc. **For complete data, send for Engineering Bulletin 7400A.**

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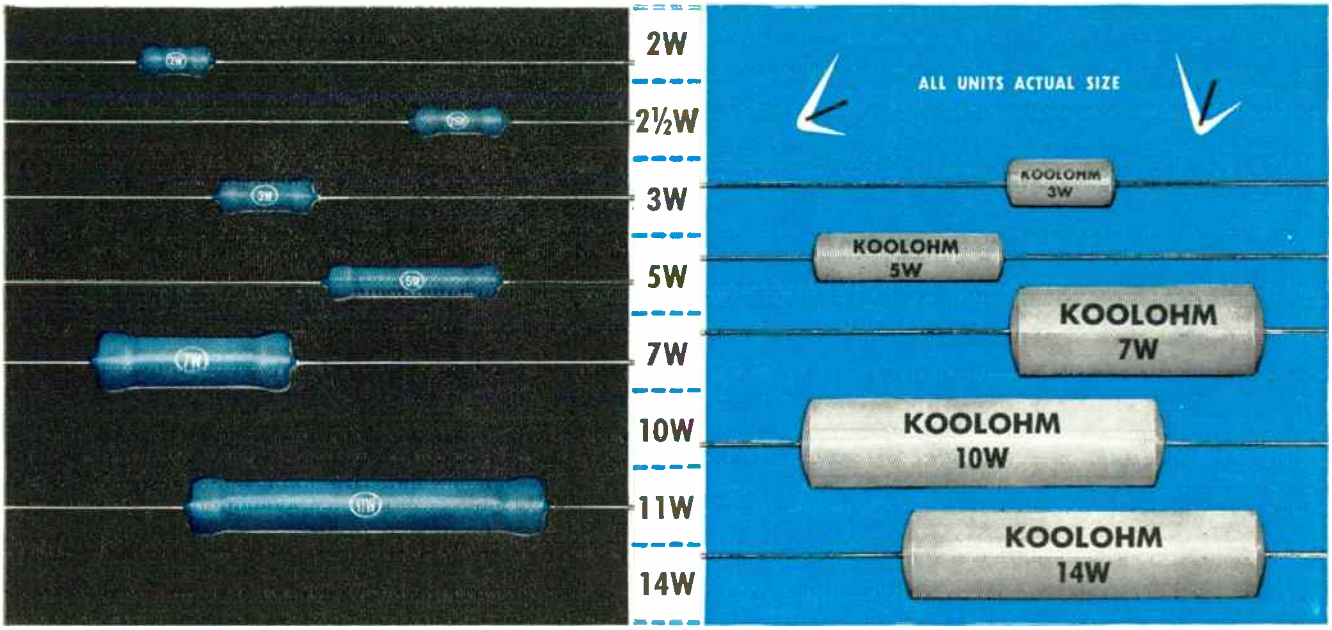
KOOLOHM®
INSULATED-SHELL POWER RESISTORS

New Koolohm construction features include welded leads and winding terminations—Ceron ceramic-insulated resistance wire, wound on special ceramic core—multi-layer non-inductive windings or high resistance value conventional windings—sealed, insulated, non-porous ceramic outer shells—*aged-on-load* to stabilize resistance value.

You can depend upon them to carry maximum rated load for any given physical size.

Send for Engineering Bulletin 7300 for complete technical data.



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Meetings with Exhibits

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

Δ

February 3-5, 1960
PGMIL Winter Meeting, Biltmore Hotel, Los Angeles, Calif.
Exhibits: Mr. Einer Ingebretson, Summers Gyroscope Co., Santa Monica, Calif.

March 21-24, 1960
IRE 1960 International Convention and Engineering Show, Waldorf-Astoria Hotel and New York Coliseum, New York, N.Y.
Exhibits: Mr. William C. Copp, Institute of Radio Engineers, 72 West 45th St., New York 36, N.Y.

April 3-8, 1960
Sixth Nuclear Congress, New York Coliseum, New York, N.Y.
Exhibits: Mr. F. M. Howell, c/o EJC, 29 W. 39th St., New York 18, N.Y.

April 20-22, 1960
SWIRECO, Southwestern IRE Regional Conference & Electronics Show, Shamrock-Hilton Hotel, Houston, Texas.
Exhibits: Mr. A. D. Seixas, SWIRECO, P.O. Box 22331, Houston, Texas.

May 2-4, 1960
National Aeronautical Electronics Conference, Dayton Biltmore Hotel, Dayton, Ohio.
Exhibits: Mr. Edward M. Lisowski, General Precision Lab., Inc., Suite 452, 333 West First St., Dayton 2, Ohio.

May 2-6, 1960
Western Joint Computer Conference, Fairmont Hotel, San Francisco, Calif.
Exhibits: Mr. H. K. Farrar, Pacific Tel. & Tel. Co., 140 New Montgomery St., San Francisco 5, Calif.

May 23-25, 1960
Seventh Regional Technical Conference & Trade Show, Olympic Hotel, Seattle, Wash.
Exhibits: Dr. Frank Holman, Boeing Airplane Co., 10708 39th Ave., S.W., Seattle 66, Wash.

May 24-26, 1960
Armed Forces Communications & Electronics Association Convention and Exhibit, Sheraton-Park Hotel, Washington, D.C.
Exhibits: Mr. William C. Copp, 72 West 45th St., New York 36, N.Y.

June 27-29, 1960
National Convention on Military Electronics, Sheraton-Park Hotel, Washington, D.C.
Exhibits: Mr. L. David Whitelock, BuShips, Electronics Div., Dept. of Navy, Washington, D.C.

(Continued on page 10A)

For High Accuracy—Broad Band Noise-Figure Measurement

KAY
Therma-Node

Basic Noise Source
CAT. NO. 770

Available Noise
Accurate to ± 0.1 db

3 Noise Heads
Cover 1 kc to 1000 mc

Portable
Can be Operated from
117 V, 60 cycles or 24 V dc

SPECIFICATIONS

STANDARD NOISE HEAD: (Head A furnished with Therma-Node), covers 2 to 1000 mc; output impedance 50 ohms, unbalanced. N type connectors.
Max. VSWR, variable tuned, 1.1 from 10 to 1000 mc.
Max. VSWR, fixed tuned, 1.1 from 10 to 100 mc; 1.2 from 6 to 300 mc; 1.4 from 4 to 400 mc; 2 from 2 to 500 mc.

INTERCHANGEABLE LOW FREQUENCY NOISE HEAD: (Head B), covers 1 kc to 400 mc; output impedance 50 ohms, unbalanced.
Max. VSWR 1.1 from 3 kc to 100 mc; 1.2 from 2 kc to 250 mc; 1.4 from 1 kc to 400 mc.

AMBIENT SOURCE PROBE for use with A and B Noise Heads:
Frequency range—0 to 1000 mc.
Output impedance—50 ohms, unbalanced.
Max. VSWR—1.1
Accuracy of indicated temperature—±1%.

SELECTABLE-IMPEDANCE NOISE HEAD: (Head C) covers 0.25 to 400 mc, balanced or unbalanced output. Selectable output impedances of 50, 100 and 200 ohms are provided.
Max. VSWR—1.1 from 1 to 75 mc; 1.2 from 0.5 to 100 mc; 1.4 from 0.25 to 400 mc.
Max. VSWR difference between ambient and hot source is 0.05 (hot and ambient sources contained in same probe).

Weight: 8 pounds in carrying case.
Dimensions: 11.5 x 8 x 4.75 inches.
Operates on 117 V, 60 cps, or 24 V dc
Price: $495.00, f.o.b. factory.
Low Frequency Noise Head (B): $125.00.
Selectable Impedance Head (C): $125.00.

Through refinement of a basic noise generation technique—thermal noise from a heated resistive element—the new Kay *Therma-Node* achieves high accuracy over an extremely wide range of frequencies. *Therma-Node's* resistive element, contained in the noise head, is heated to a normal 2200° K, generating adequate noise-power for accurate noise figure measurements to 10 db. Nominal fixed available noise temperature ranging between 2000 and 2400° K may be read directly on the panel meter to 2% accuracy. A single tuning element, contained in the noise head, provides a fixed range of 2 to 500 mc, and may be tuned to extend the range to 1000 mc. An optional, interchangeable head extends measurement down to 1 kc. Ambient termination is supplied. Both heads have output impedances of 50 ohms, unbalanced. Both heads can be used without connecting coaxial cables, thus eliminating cable errors.

A selectable-impedance noise head, covering the range .25 to 400 mc, and furnishing output impedances of 50, 100 and 200 ohms, balanced and unbalanced, is available as an accessory.

The inexpensive resistive element in the *Therma-Node* noise head has a life expectancy of 10,000 hours in either intermittent or continuous service. Because the few active components in the *Therma-Node* are solid state devices, its inherent stability results in long term accuracy and freedom from maintenance.

OTHER KAY NOISE GENERATORS

Instrument & Cat. No.	Frequency Range (mc)	Noise Figure Range (db)	Output Impedance (ohms)	Price f.o.b. factory
Mega-Node 240-B	5-220	0-16 at 50 ohms 0-23.8 at 300 ohms	unbal.—50, 75, 150, 300, ∞ bal.—100, 150, 300, 600, ∞	$365.00
Mega-Node 175-A	50-500	0-19	balanced—300	$365.00
Mega-Node 403-A	3-500	0-19	unbalanced—50	$365.00
Mega-Node 3000	10-3000	0-20	unbalanced—50	$790.00
Rada-Node 600-A	5-400	0-23.8 depending on impedance	unbalanced as specified	$1495.00
	10-3000	0-20	unbal. nom. 50	$1965.00
	1120-26,500	15.28 or 15.8	waveguide	†
Micro-Node 1080-A	3700-4200	0-15.8	waveguide	$795.00
*Microwave Mega-Nodes**	1120-26,500	15.28 or 15.8	waveguide	$175.00 to $595.00

† Price varies with Microwave Mega-Node discharge tube used as accessory.
* Ideally suited for noise figure measurement in radar communication.

WRITE FOR
KAY CATALOG 1959-A

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Meetings with Exhibits

(Continued from page 8A)

August 23-26, 1960

WESCON, Western Electronic Show and Convention, Ambassador Hotel & Memorial Sports Arena, Los Angeles, Calif.

Exhibits: Mr. Don Larson, WESCON, 1135 LaCienega Blvd., Los Angeles, Calif.

September 19-21, 1960

National Symposium on Space Electronics & Telemetry, Shoreham Hotel, Washington, D.C.

Exhibits: John Leslie Whitlock Associates, 6044 Ninth St., North, Arlington 5, Va.

October 3-5, 1960

Sixth National Communications Symposium, Hotel Utica & Utica Memorial Auditorium, Utica, N.Y.

Exhibits: Mr. W. R. Roberts, 102 Fort Stanwix Park N., Rome, N.Y.

October 10-12, 1960

National Electronics Conference, Hotel Sherman, Chicago, Ill.

Exhibits: Mr. Arthur H. Streich, National Electronics Conference, 184 E. Randolph St., Chicago, Ill.

October 24-26, 1960

East Coast Aeronautical & 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 14-16, 1960

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

Exhibits: Mr. John V. Parks, Bendix Aviation Corp., P.O. Box 1159, Kansas City 41, Mo.

November 15-17, 1960

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

Exhibits: Miss Shirley Whitcher, IRE Boston Office, 73 Tremont St., Boston, Mass.

December 1-2, 1960

PGVC Annual Meeting, Sheraton Hotel, Philadelphia, Pa.

Exhibits: Mr. E. B. Dunn, Atlantic Refining Co., 260 S. Broad St., Philadelphia 1, Pa.

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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RELIABLE
RAYTHEON
CIRCUIT-PAKS
CK709
RAYTHEON
* E PLURIBUS UNUM (one unit composed of many parts) aptly describes Raytheon Circuit-Pak.

Calendar of Coming Events and Authors' Deadlines*

1960

6th Natl. Symp. on Reliability and Quality Control, Statler-Hilton Hotel, Washington, D. C., Jan. 11-13.

PGMIL Winter Mtg., Biltmore Hotel, Los Angeles, Calif., Feb. 3-5.

1960 Solid State Circuits, Conf., Sheraton Hotel, Philadelphia, Pa., Feb. 10-12.

IRE National Conv., N. Y. Coliseum and Waldorf-Astoria Hotel, New York, N. Y., Mar. 21-24.

First Natl. Symp. on Human Factors in Electronics, New York, N. Y., Mar. 24-25.

Scintillation Counter Symp., Washington, D.C., Mar.

6th Nuclear Congress, N. Y. Coliseum. New York, N. Y., Apr. 4-8.

Conf. on Automatic Tech., Sheraton-Cleveland Hotel, Cleveland,, Ohio, Apr. 18-19.

Int'l Symp. on Active Networks and Feedback Systems, Engrg. Soc. Bldg. Auditorium, New York, N. Y., Apr. 19-21. (DL*: Jan. 15, H. J. Carlin, 55 Johnson St., Brooklyn, N. Y.)

Int'l Symp. on Active Networks and Feedback Systems, Polytechnic Inst. of Brooklyn, Brooklyn, N. Y., Apr. 19-21.

SWIRECO (Southwestern Regional Conference), Houston, Texas, Apr. 20-22.

Natl. Aeronautical Electronics Conf., Biltmore and Miami-Pick Hotels, Dayton, Ohio, May 2-4. (DL*: Jan. 15, NAECON Papers Committee, P.O. Box 621, Far Hills Branch, Dayton 19, Ohio)

Western Joint Computer Conf., San Francisco, Calif, May 2-6.

PGMTT Natl. Symp., San Diego, Calif., May 9-11. (DL*: Jan. 15, D.B. Medved, CONVAIR, Div. of Gen. Dynamics Corp., Mail Zone 6-172, San Diego, Calif.

Electronic Components Conf., Hotel Washington, Washington, D. C., May 10-12.

7th Reg. Tech. Conf. & Trade Show, Olympic Hotel, Seattle, Wash., May 23-25.

Conf. on Electronic Standards and Measurements, NBS Boulder Labs., Boulder, Colo., June 22-24.

Natl. Conv. on Mil. Elec., Sheraton Park Hotel, Washington, D. C., June 27-29.

* DL = Deadline for submitting abstracts.

(Continued on page 15A)

IRE Elects Officers for 1960

Ronald L. McFarlan (SM'51), consultant to the DATAmatic Corp. and the Raytheon Manufacturing Co., has been elected president for 1960 of the Institute of Radio Engineers. He succeeds Ernst Weber (M'41-SM'43-F'51), president of the Polytechnic Institute of Brooklyn, Brooklyn, N.Y.

In recognition of the rapid growth of the institute's activities, both here and abroad, the IRE will for the first time in its history have two vice presidents in 1960, one residing in North America and the other from abroad. The vice president representing overseas countries for 1960 will be J. A. Ratcliffe (M'29-A'32-F'53), head of radio research at the Cavendish Laboratory in Cambridge, England. The vice president representing North America will be J. N. Dyer (J'30-A'32-SM'45-F'49), vice president of the Research and Engineering Division of Airborne Instruments Lab., Melville, N. Y. Mr. Dyer and Mr. Ratcliffe will succeed Donald B. Sinclair (J'30-A'33-M'38-SM'43-F'43), vice president and chief engineer of General Radio Co., West Concord, Mass.

Elected as directors for the 1960-1962 term are W. G. Shepherd (A'42-SM'49-F'52), head of the department of Electrical Engineering, University of Minnesota, Minneapolis, Minn., and George Sinclair (A'37-SM'46-F'54), Professor of Electrical Engineering, University of Toronto, Toronto, Canada.

Regional Directors elected for 1960-1961 are as follows: Region 1—J. B. Russell, Jr. (A'36-SM'45-F'58) manager, Electronics Lab., General Electric Co., Syracuse, N. Y.; Region 3—B. J. Dasher (A'41-M'55-SM'57), director, School of Electrical Engineering, Georgia Institute of Technology, Atlanta, Ga.; Region 5—R. E. Moe (S'33-A'35-SM'46-F'55), manager of engineering, Receiving Tubes Dept., General Electric Co., Owensboro, Ky.; Region 7—C. W. Carnahan (A'34-SM'45-F'52), technical assistant to the president, Varian Associates, Palo Alto, Calif.

MIL-E-CON Issues Call for Papers

The 4th National Convention on Military Electronics—1960 (MIL-E-CON) will be held at the Sheraton-Park Hotel, Washington, D. C., on June 27-29, 1960, under the sponsorship of the Professional Group on Military Electronics. Robert H. Cranshaw, Manager, Advanced Space Projects, General Electric Company, Utica, N. Y., is Convention President for MIL-E-CON.

Papers presenting original work in military electronics are invited for this meeting. Suggested topics include, but are not limited to, the following: Current Problems of Space Technology, Space Electronics, Ranging and Tracking, Electronic Propulsion, Data Handling Systems, Guidance and Control, Inertial Systems, Reconnaissance Systems, Communication Systems, Operational Analysis, and Reliability.

The technical program will include both classified (limited to confidential) and unclassified sessions, with the Air Research and Development Command sponsoring the classified sessions. An unclassified and bound *Proceedings* of the convention will be available at the meeting.

Prospective authors are requested to furnish the following information, *not later than February 1, 1960:* Three copies of a 250-word unclassified abstract of the proposed paper, plus the name and position of the author and the name and address of his company or organization. Each author must obtain the appropriate military and company clearances for his abstract.

Send abstracts to: Dr. Craig M. Cranshaw, Dept. of Army, Office of the Chief Signal Officer, R & D Div., SIGRD-2, Washington 25, D. C.

J. W. McRae Receives Army's Top Civilian Award

Dr. James W. McRae, a Fellow of the IRE and former President of the Institute, was presented the Army's Distinguished Civilian Service Award in ceremonies at Fort Monroe, Va., October 5, 1959, in recognition of his contributions to the development of small nuclear weapons for use by ground forces. He is Vice President of the American Telephone and Telegraph Co., and coordinator of their defense activities.

Dr. McRae, who is also Chairman of the Army Scientific Advisory Panel, was cited for his aid in developing a series of small, tactical nuclear weapons while President of the Sandia Corp., a Western Electric Co. subsidiary, which operates the Sandia Laboratory for the Atomic Energy Commission at Sandia Base, near Albuquerque, N. Mex.

The presentation was made during the regular semi-annual meeting of the Army Scientific Advisory Panel by Richard S. Morse, civilian Director of Research and Development for the Army. The citation was read by Lt. Gen. Arthur G. Trudeau, Chief of Army Research and Development.

"His scientific knowledge and able leadership have contributed materially to the development for the Army of small, tactical nuclear weapons, thereby significantly increasing the Army's capability to carry out its combat mission," the citation accompanying the award stated. "His outstanding performance reflects great credit on him as a patriotic citizen and able scientist."

India Section Established by IRE

At its meeting on November 16, 1959, the IRE Executive Committee approved the establishment of a new IRE Section, to be known as the India Section. The new section will encompass the entire country of India.

At a recent convention held at the Stardust Hotel in Las Vegas, Nev., the IRE formed a new chapter. Shown above at the conclave are *(left to right)* Ben Ham, R. B. Patten and G. A. Fowler, Director of Region 7 all of Las Vegas, Ernest A. Hopkinson, and Ray Banks, ex-secretary of Los Angeles IRE chapter.

IRE Announces 1960 Awards

Haraden Pratt (A'14-M'17-F'29)(L), former telecommunications advisor to the President, and Harry Nyquist (A'39-SM'47-F'52), renowned former Bell Telephone Laboratories scientist, were among those named to receive IRE awards in 1960. Presentation of the awards will take place at the 1960 IRE International Convention banquet next March 23 at the Waldorf-Astoria Hotel in New York, N. Y.

Mr. Pratt, Secretary of the IRE and consulting engineer, has been named to receive the 1960 Founders Award "for outstanding contributions to the radio engineering profession and to The Institute of Radio Engineers through wise and courageous leadership in the planning and administration of technical developments which have greatly increased the impact of electronics on the public welfare." This is one of the two highest IRE awards and is bestowed only on special occasions.

Dr. Nyquist, also a consulting engineer, will receive the 1960 Medal of Honor, the highest annual technical award in the field of electronics. He will be given the award "for fundamental contributions to a quantitative understanding of thermal noise, data transmission and negative feedback."

Four additional awards will be given as follows:

1960 Morris Liebmann Memorial Prize Award (Awarded to a member of the IRE for a recent important contribution to the radio art.) To J. A. Rajchman (SM'46-F'53) RCA Laboratories, Princeton, N. J., "for contributions to the development of magnetic devices for information processing."

1960 Browder J. Thompson Memorial Prize Award (Awarded for an IRE paper combining the best technical contribution and presentation which has been written by an author under thirty.) To J. W. Gewartowski (S'53-M'57), Bell Telephone Laboratories, Murray Hill, N. J., for his paper entitled "Velocity and Current Distributions in the Spent Beam of the Backward-Wave Oscillator," which appeared in the October, 1958 issue of IRE Transactions on Electron Devices.

1960 Harry Diamond Memorial Award (Awarded to a person in government service for outstanding contributions in the field of radio or electronics as evidenced by publication in professional journals.) To K. A. Norton (A'29-M'38-SM'43-F'43), National Bureau of Standards, Boulder, Colo., "for contributions to the understanding of radio wave propagation."

1960 W. R. G. Baker Award (Awarded to the author of the best paper published in the IRE Transactions of the Professional Groups.) To E. J. Nalos (SM'54), General Electric Company, Palo Alto, Calif., for his paper entitled "A Hybrid Type Traveling-Wave Tube for High-Power Pulsed Amplification," which appeared in the July 1958 issue of the IRE Transactions on Electron Devices."

British IRE Gives Prize to Two American Engineers

The Council of the British Institute of Radio Engineers has announced the awards made for outstanding papers published in the Institution's Journal during 1958. One of these awards, the Marconi Premium, was given to Dr. Morton B. Prince (SM'55) and Martin Wolf (A'52-M'54) of Hoffman Electronics Inc., Evanston, Ill. Their paper, "New Developments in Silicon Photovoltaic Devices," was read at a meeting in London in June, 1958, and is considered to be the outstanding engineering paper published in the Journal during that year.

The Premium was presented by the President of the British IRE, Professor E. E. Zepler, at the Annual General Meeting of the Institution in London, December 2, 1959.

Army MARS Names Speakers

The First U. S. Army MARS SSB Technical Net, which can be heard Wednesday at 8 P.M. EST on 4030 kc upper sideband, has scheduled four speakers for January.

January 6—"The Atomichron," P. Heath, Manager, Field Service Engrg. Dept., National Co., Inc.

January 13—"Antenna Multi-Couplers," C. G. Southeimer Executive Vice President, CGS Labs.

January 20—"Television and the Amateur Operator," G. G. Lentzakis, Chief Instructor TV Course, Radio Div., USASCS, Fort Monmouth.

January 27—"Basics and Applications of Re-enforced Plastics in Communications Products," W. H. Greenberg, Director of Research, J. H. McCoy, Staff Engineer, Riverside Plastics Corp.

Calendar of Coming Events and Authors' Deadlines*

(Continued from page 14A)

Cong. Intl. Federation of Automatic Control, Moscow, USSR, June 25–July 9.

Int'l Conf. on Electrical Engrg. Education, Sagamore Conf. Center, Syracuse Univ., Syracuse, N. Y., Jul.

WESCON, Los Angeles Mem. Sports Arena, Los Angeles, Calif., Aug. 23–26.

Space Electronics and Telemetry Conv. and Symp., Shoreham Hotel, Washington, D.C., Sept. 19–22.

Industrial Elec. Symp., Sept. 21–22.

Sixth Natl. Communications Symp., Hotel Utica and Utica Memorial Aud., Utica, N. Y., Oct. 3–5.

Natl. Elec. Conf., Chicago, Ill., Oct. 10–12.

East Coast Conf. on Aero & Nav. Elec., Baltimore, Md., Oct. 24–26.

Electron Devices Mtg., Hotel Shoreham, Washington, D. C., Oct. 27–29.

Radio Fall Mtg., Hotel Syracuse, Syracuse, N. Y., Oct. 31, Nov. 1–2.

Mid-Amer. Elec. Conv., Kansas City, Mo., Nov. 14–16.

1960 NEREM (Northeast Electronics Res. & Engrg. Mtg.), Boston, Mass., Nov. 15–17.

13th Ann. Conf. on Elec. Tech. in Med. and Bio., Washington, D. C., Nov.

PGVC Ann. Mtg., Sheraton Hotel, Philadelphia, Pa., Dec. 1–2.

Eastern Joint Computer Conf., New Yorker Hotel, New York, N.Y., Dec.

1961

7th Natl. Symp. on Reliability and Quality Control, Bellevue-Strafford Hotel, Philadelphia, Pa., Jan. 9–11. (DL*: May 9, 1960, W. T. Summerlin, Philco Corp., 4700 Wissahickon Ave., Philadelphia 44, Pa.)

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

Electronic Computer Conf., West Coast, May 9–11.

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

Natl. Symp. on Space Elec. and Telemetry, Sept.

Industrial Elec. Symp., Sept. 20–21.

National Elec. Conf., Chicago, Ill., Oct. 9–11.

1961 NEREM (Northeast Electronics Res. & Engrg. Mtg.), Nov. 14–16.

1961 Electron Devices Mtg., Shoreham Hotel, Washington, D. C., Nov. 20–21.

Mid-America Elec. Conv. (MAECON), Kansas City, Mo., Nov. 14–16.

Eastern Joint Computer Conf., Sheraton Park Hotel, Washington, D. C., Dec.

* DL = Deadline for submitting abstracts.

IRE-URSI Spring Meeting Announces Paper Deadline

The IRE-URSI Spring Meeting will be held at the Sheraton-Park Hotel, Washington, D. C., on May 2–5, 1960. Authors are invited to submit, on or before March 1, 1960, titles and 100–200 word abstracts of papers offered for presentation. It is necessary that the papers be prepared *in duplicate*, and mailed to:

Commission Chairman (choose appropriate one for your paper)
c/o Mr. K. S. Kelleher
Chairman, Local Arrangements Committee
1200 Duke Street (Box 1082)
Alexandria, Va.

Defense Department Presents Award to DOFL Science Team

The Department of Defense has given its highest official award to J. R. Nall, T. A. Prugh (A'46–M'52–SM'56), Mrs. E. M. Olson, N. J. Doctor (M'58) and Dr. J. W. Lathrop. Secretary of the Army. Wilber M. Brucker presented the $25,000 award in a Pentagon ceremony on October 12, 1959.

Working as a science team at the Army's Diamond Ordnance Fuse Laboratories (DOFL), Washington, D. C., in 1957, the group's research culminated in the discovery of a new process to fabricate printed micro-circuits. By using photography and lithography in combination with other techniques, they constructed subassemblies so small and compact that an electronic system can be made with 1000 components per cubic inch compared to the 30 to 50 possible using conventional components.

For this breakthrough in miniaturization, the team has had recognition twice previous to the Defense Department Award. In January, 1958, Diamond Ordnance Fuse Laboratories awarded them a Certificate of Achievement. The $300 prize was "For advancing the art of micro-miniaturization through utilization of photolithographic and printed circuit techniques to produce integrated electronic sub-assembly." In turn, DOFL was presented with the first annual Miniaturization Award (sponsored by Miniature Precision Bearings, Inc.).

12th Annual ETMB Conference Technical Papers Available

Copies of the 12th Annual Conference on Electrical Techniques in Medicine and Biology Digest of Technical Papers are available from Henry G. Sparks, Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia 4, Pa., at $4.00 a copy. Remittance for these books should be made out to the order of: 12th Annual Conference, Electrical Techniques, Medicine and Biology.

The Digest is a 100-page (letterpress) report of all the papers presented, and more than 150 specially prepared illustrations are included.

Air Force MARS Announces January Program Schedule

The following is the schedule of programs for the Air Force MARS Eastern Technical Net for January and the first week of February. They can be heard on Sundays from 2 to 4 P.M. EST at 3295 kc, 7540 kc, and 15,715 kc.

January 3—Recess date.
January 10—Review of Technical Topics.
January 17—"Basics of Single Sideband," H. G. Adams and S. E. Piller, Eldico Electronics Div. of R.E.L.
January 24—"Fundamentals of Transistors," R. Gunderson, Braille Technical Press.
January 31—"Optics and the Visible Spectrum," E. C. Scott, Scientist, Rome Air Dev. Center.
February 7—"Infrared," R. D. Byrne, Scientist, Rome Air Dev. Center.

Baker Given Sarnoff Medal at Syracuse Pioneer Night

The Syracuse Section of the IRE played host to some of the nation's outstanding pioneers of early radio on November 5, 1959. Engineers in the area gathered at the Syracuse Museum of Fine Arts to hear and see the equipment and men who were the forerunners of today's dynamic communication and electronic industry. Another highlight of the meeting was the presentation of the David Sarnoff Gold Medal of the Society of Motion Picture and Television Engineers to Dr. W. R. G. Baker, President of Syracuse University Research Corporation, Syracuse, N. Y.

Dr. Baker was awarded the medal "for his outstanding work as Chairman of the National Television System Committee which functioned on two separate occasions when standards were urgently needed to promote the growth of television in the United States." The medal was formally presented at an earlier meeting of the Society of Motion Picture and Television Engineers, but Dr. Baker was unable to be present to receive the award. Dr. Alfred N. Goldsmith, cofounder and editor emeritus of the IRE, received the award for him on this occasion, and presented it to him personally at the Syracuse meeting.

The four radio pioneers who participated in the meeting were Raymond A. Heising, William C. White, Haraden Pratt and Dr. Goldsmith. Mr. Heising designed and constructed the first phone company transmitters to successfully send speech from Washington to the West Coast, Panama and Hawaii. Later he worked in development of trans-Atlantic shortwave telephone service. He holds 130 U. S. patents and is best known for the Heising Modulation System.

Mr. White, now retired from General Electric, is well known for his participation in the first long distance radio communication made in the United States, in which the first two words were "Hello, White." He has held important positions in the formulation of standards for electronics.

Mr. Pratt was one of the first amateur wireless telegraphers, having started in 1906 on the West Coast. He worked for Marconi Company and Mackay Radio, retiring from the International Telephone Corp. in 1951.

Dr. Goldsmith has spent most of his life teaching in New York, N. Y. He holds over 200 U. S. patents.

In addition to the forum conducted by these men, an exhibit by the Rochester Antique Wireless Club actually transmitted and received messages with equipment dating back to 1887.

Circuit Theory Transactions To Publish Special Issue

The Professional Group on Circuit Theory is planning a special issue on "Applications of Electronic Computers to Network Design," which is tentatively scheduled for September, 1960.

The first purpose of the special issue is to accumulate digital computer programs of circuit theory interest for a PGCT program library. The custodian of the library will issue copies of the programs to members, and to the public, for a nominal charge appropriate to the medium employed, *i.e.*, punched cards, magnetic tape, etc. The special issue will publish the texts of instructions for use of the various programs.

In addition, there is a need for original and creative papers on any phase of this general subject. Papers, for example, on the potential analog would be of interest, and many other topics are possible. The tentative deadline for papers is June, 1960.

Please address contributions to Philip R. Geffe, Guest Editor, PGCT Transactions, North Hills Electric Co., Inc., 402 Sagamore Avenue, Mineola, L. I., N. Y.

A recent meeting of the 1960 International Solid-State Circuits Conference national committee in the Sheraton Hotel, Philadelphia, Pa., where the details for the February 10-12, 1960 program, were finalized. *Left to right:* William Howard, sponsors advisory committee; Lewis Winner, public relations chairman; Robert F. Cotellessa, Technical Digest editor; Tudor R. Finch, program chairman; Arthur P. Stern, conference chairman, S. K. Ghandhi, conference secretary; Solomon Charp, conference treasurer; Murlan S. Corrington, chairman of the sponsors advisory committee; Robert Mayer and Henry Sparks, local arrangements.

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OBITUARIES

Cyril Nathaniel Hoyler (A'35–SM'45), Manager of Technical Relations of RCA Laboratories in Princeton, N. J., died recently after a station wagon in which he was riding was struck by a train at a grade-crossing near Moosomin, Saskatchewan, Canada. He was fifty-four.

A widely known lecturer on electronic science, he was in Canada on one of a series of demonstration lecture tours that had carried him through virtually every state in this country and every province in Canada during the past ten years. He had appeared before college, professional and civic groups in scores of cities.

He joined RCA in 1941 as a member of the technical staff at Camden, N. J. In 1942, he transferred to the newly formed RCA Laboratories in Princeton, where he became Administrative Assistant in 1945. He was appointed Manager of Technical Relations in 1952.

Born in Edmonton, Alberta, in 1905, he taught for two years in the rural schools of Canada before coming to the United States to attend Moravian College, Bethelhem, Pa. He was graduated from Moravian in 1928 with the B.S. degree. In 1935, he received the M.S. degree in Physics from Lehigh University, and was awarded the professional degree of Electrical Engineer by Lehigh in 1957. In 1958, Moravian College conferred upon him its Comenius Award as a distinguished alumnus.

Prior to his association with RCA, he was an instructor of mathematics and German at the Junior High School of Irvington, N. J., before joining the Faculty of Moravian in 1929, where he was successively instructor, professor and head of the Physics Department.

Mr. Hoyler was the author and co-author of a number of technical articles and a book on radio frequency heating. He was a member of Sigma Xi and the American Institute of Electrical Engineers.

Balthasar van der Pol (M'25–F'29) (L), a former Vice-President and Director of the IRE, died recently at the age of 70.

Born in 1889, he studied physics at the University of Utrecht, Netherlands, from 1911 to 1916, and from 1916 to 1919 he studied in England under J. A. Fleming in London and J. J. Thomson in Cambridge. In 1919 he received the D.Sc. degree from the University of Utrecht. During the next year, while studying under H. A. Lorentz at Teyler's Foundation, Haarlem, Netherlands, he was co-founder of the "Nederlands Radiogenootschap."

From 1922 to 1949 he was on the staff of the Philips Research Laboratories, Eindhoven, Netherlands, in the later years as "Director of Fundamental Radio Research." He was also Extraordinary Professor, Technical University, Delft, Netherlands, from 1938 to 1949. He became Director of the International Radio Consultative Committee (CCIR) in 1949, the position which he held until 1956, when he retired.

He was Vice-President of URSI from 1934 to 1950, and in 1952 he was named Honorary President of that organization. He was made an Honorary Member of the Netherlands Radio Society in 1920. Other awards he received included the Valdemar Poulsen Gold Medal of the Danish Aacdemy of Technical Sciences; Knight, Order of Orange Nassau; Knight, Order of the Netherlands Lion; Doctor, honoris causa, Technical University, Warsaw, Poland; and an Honorary Doctor's degree, University of Geneva, Switzerland. He was a member of the Royal Netherlands Academy of Sciences.

Dr. van der Pol's scientific work was characterized by his great interest in the mathematical side of radio science. He investigated many problems in this field and made substantial contributions to their solution. The first of these was the propagation of radio waves. His further investigations in this field culminated in 1937 in the calculation of the diffraction of radio waves around a finitely conducting sphere. The study of triode oscillations resulted in a number of papers dealing with the influence of the nonlinearities of the characteristics. This led him eventually to the study of relaxation oscillations, described by the now-called "Van der Pol equation." His work also contributed to advancement in work on the early theory of antennas, and to the theory of circuits and of frequency modulation. In recent years he was engaged in many applications of his knowledge of Laplace transforms to pure mathematics: special functions, number theory.

In addition to being a member of several committees of the IRE, Dr. van der Pol was a Vice-President of the Institute in 1934 and a Director in 1934. He received the IRE Medal of Honor in 1935.

1960 Winter Convention on Military Electronics

BILTMORE HOTEL, LOS ANGELES, CALIF., FEBRUARY 3–5, 1960

Wednesday Morning, February 3
Keynote Session

"The Satellite as a Transmission Medium for Global Communications," Moderator: *J. M. Bridges, Dir., Office of Electronics, Office of the Dir. of Defense Res. and Engrg., Dept. of Defense, Washington, D. C.* Panel Members: *Paul Price, Chief, Commun. and Tracking Branch, ARPA; Dr. J. R. Pierce, Dir., Res. Electrical Commun. Principles, Bell Telephone Labs., Murray Hill, N. J.; Dr. H. K. Ziegler, Chief Scientist, U. S. Army Signal Res. and Dev. Labs., Fort Monmouth, N. J.; Brig. General J. B. Bestic, U.S.A.F., Deputy Dir., Communication-Electronics, Dept. of the Air Force, Washington, D. C.*

Wednesday Afternoon
Session 1—Surveillance (Confidential)—Sponsor: Air Res. and Dev. Command

"Single Aperture CW Radar Techniques," *W. S. O'Hare, Raytheon Mfg. Co., Maynard, Mass.*

"A Single-Look Dual-Threshold Acquisition Automatic Ranging System," *G. Strauss, W. L. Maxson Co., New York, N. Y.*

"High-Power Sonar Equipment Design," *P. S. Goodwin, Ling-Altec Electronics, Anaheim, Calif.*

"The AN/USA-3 (XE-1) Infrared Surveillance System," *M. H. Mehr, Perkin-Elmer, Norwalk, Conn.*

"Anti-Submarine Warfare," *Dusseau, Bendix-Pacific, North Hollywood, Calif.*

Session 2—Satellite Systems

"A Satellite Navigation System Employing a Synchronized Pattern," *E. Cole, The Martin Co., Baltimore, Md.*

"Satellite Auxiliary Power Systems," *N. B. Palmquist, Lockheed Aircraft Corp., Missile and Space Div., Sunnyvale, Calif.*

"Explorer VI Digital Telemetry-Telebit," *R. Gottfried, Space Technology Labs., Los Angeles, Calif.*

"Communications System For Project MERCURY Space Capsule," *W. Brenner, McDonnell Aircraft, St. Louis, Mo.*

Session 3—Instrumentation

"The Mathematics and Hardware of Takeoff Monitors For Turbojet Aircraft," *A. L. deGraffenried, Avien, Inc., Woodside, L. I., N. Y.*

"Analysis of Airborne Ablation Instrumentation Measurements For Second Generation ICBM Re-entry Vehicles," *L. Foster, General Electric, Missile and Space Div., Philadelphia, Pa.*

"PERCOS—Performance Coding System," *Dr. E. Keller, Motorola, Inc., Chicago Military Electronics Center, Chicago, Ill.*

"An RF Instrument For Wide-Band Low-Level Voltage Measurements," *D. Fryberger, R. Schultz, L. Greenstein, Armour Res., Chicago, Ill.*

Session 4—Communications I

"System Design For The Aircom Modernization Program," *T. J. Heckelman, Philco Corp., Philadelphia, Pa.*

"An Operations Research Approach For Dealing With The Functional Behavior of Communications Systems," *F. Wertheimer, Vitro Labs., West Orange, N. J.*

"A New Method of Forecasting Maximum Transmision Range At UHF," *L. B. Gardner, Litton Ind., Beverly Hills, Calif.*

"Evaluating Alphanumeric Codes When Considering Terminal Equipment," *H. Landerer, ITT, Communications System, Inc., Paramus, N. J.*

Thursday Morning, February 4
Session 5—Reconnaissance (Confidential)—Sponsor: Air Res. and Dev. Command

"Specialized Reconnaissance Techniques," *J. DeBroekert, Stanford Electronic Lab., Stanford Univ., Stanford, Calif.*

"A Fully Transistorized Electronic Reconnaissance System AN/DLD-1 (XH-1)," *R. Lowan, J. Kerney, A. Boecker, Airborne Instruments Lab., Huntington, L. I., N. Y.*

"Time Difference Intercept System," *R. F. Borofka, ITT, San Fernando, Calif.*

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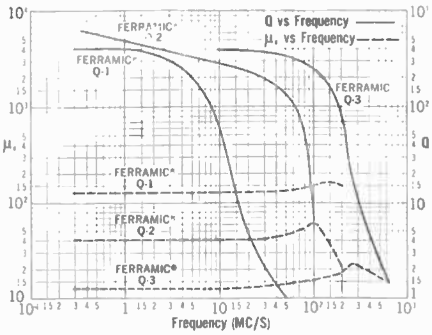
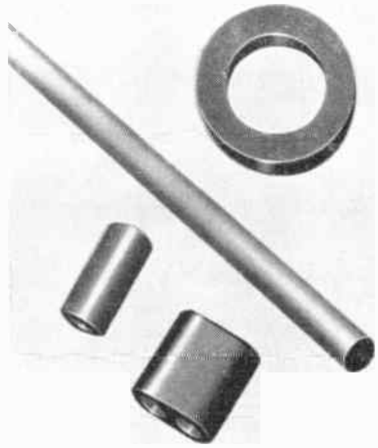


This curve demonstrates the exceptional qualities of G.C's "Q" family of materials —high permeability and low loss.



Q-1 Material

Ideally suited for IF transformers at 455 kc and antenna cores from 500 kc to 1700 kc.

Magnetic properties include: Initial permeability at 1 mc/sec. – 125; Maximum permeability – 400; Saturation flux density – 3300 gauss; residual mag. – 1800 gauss; coercive force – 2.1 oersted; temperature coefficient of initial permeability – .10%/°C max; Curie point – 350°C; vol. resistivity – high; loss factor at 1 mcs/sec. – .000020; loss factor at 5 mcs/sec. – .000050.

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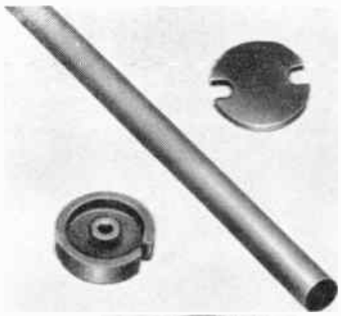

This material produces excellent results in TV receiver IF applications at 45 mc; and FM radio applications at 10.7 mc.

Magnetic properties include: Initial permeability at 1 mc/sec. – 40; Maximum permeability – 115; Saturation flux density – 2400 gauss; residual mag. – 750 gauss; coercive force – 4.7 oersted; temperature coefficient of initial permeability – 0.10%/°C max; Curie point – 450°C; vol. resistivity – high loss factor at 10.0 mc/sec. – 0.000085; loss factor at 50.0 mc/sec. – 0.000017.

Q-3 Material

The exceptionally high Q, excellent temperature stability and high permeability of Q-3 ideally suits it for applications in the 88 mc–108mc FM band and television RF frequencies from 50 mc–216 mc.

Preliminary specifications include: Initial permeability μ_o – 14; Maximum permeability μ_{max} – 42; Saturation flux density B_s – 2600 gauss; Residual magnetism B_r – 1470 gauss; Coercive Force – 21 oersted; Q at 100 mc – 250; Q at 150 mc – 150; Q at 200 mc – 70; Resistivity high; Curie point 330°C; Temperature coefficient of μ_o – .09%/°C max.

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"The AN/AAS-5 Infrared Reconnaissance System," *W. Birtley and D. Chaffee, Jr., HRB-Singer, State College, Pa.*

"Prediction and Simulation of Infrared Reconnaissance Maps," *R. Replinger, Bendix-Aviation, Systems Div., Ann Arbor, Mich.*

Session 6—Satellite Subsystems

"A Stationary Sun Position Sensor," *G. Anthony, IBM, Owego, N. Y.*

"A Memory System For Satellites," *G. Boyer, DataLab, Consolidated Electrodynamics Corp., Pasadena, Calif.*

"Directional Magnetometer For Low Intensity Magnetic Field," *H. DeBolt, Fairchild Camera, Syosset, L. I., N. Y.*

"Development of Portable Atomic Frequency Standard for Missile Borne and Satellite Borne Applications," *M. L. Stitch and H. Lyon, Hughes Aircraft Co., Culver City, Calif.*

Session 7—Reliability I

"How Can We Attain High Reliability of Complex Military Electronic Equipment?" *M. Halio, Hq. Air Defense Command Detachment 2, McGuire AFB, N. J.*

"Product Assurance," *E. Winlund, General Electric, Computer Dept., Phoenix, Ariz.*

"Controlling Weapon System Reliability Through An Effective Reliability Specification," *C. Bird, IBM, Owego, N. Y.*

"A Navigational Computer Reliability Program History," *F. M. Schriever, Ford Instrument, Long Island City, L. I., N. Y.*

Session 8—Communications II

"Prediction and Siting Techniques For Tactical Tropospheric Scatter Systems," *K. B. Roe, Collins, Tucson, Ariz.*

"AutoData-RCA's Automatic Message Switching System," *T. Patterson, RCA, Camden, N. J.*

"The Use of Phase of a Sinusoidal Signal as the Controlling Variable," *A. Kline, Motorola Inc., Western Military Electronics Center, Phoenix, Ariz.*

"Synthesis of Multipole Narrow Band Filters," *J. Hupert, De Paul Univ., Chicago, Ill., and A. Sances, American Machine and Foundry, Chicago, Ill.*

Thursday Afternoon

Session 9—Guidance and Control (Confidential)—Sponsor: Air Res. and Dev. Command

"Cyrogenic Gyros," *W. H. Culver and M. H. Davis, The RAND Corp., Santa Monica, Calif.*

"The AN/UPW-1 System, a Tracking and Command Guidance Radar," *S. McAloney, Ford Instr., L. I., N. Y.*

"Error Analysis of a Radar Guided Ballistic Missile On Non-Optimum Trajectories," *H. Perry, ARGMA, Redstone Arsenal, Ala.*

"Basic Principles of Fuzing High Explosive Warheads for Use Against Air Targets," *J. Brown and A. Sylvester, U. S. Naval Ordnance Lab., Corona, Calif.*

Session 10—Data Handling

"Range Safety Instrumentation At Vandenberg AFB," *J. Moeller, Space Technology Labs., Los Angeles, Calif.*

"Mobile HF Data Transmission," *W. Nupp, U. S. Naval Air Dev. Center, Johnsville, Pa.*

"An Integrated Checkout Equipment for Evaluating the Performance of Satellite Data Handling Systems," *W. Dickens, et al., Philco, Palo Alto, Calif.*

"High Performance Airborne Analog-To-Digital Converter Using Germanium Transistors," *S. Ruhman, Packard-Bell Computer Corp., Santa Monica, Calif.*

Session 11—Reliability II

"Study of Electronic Equipment Life Prediction," *A. Kneale, Motorola Inc., Western Mil. Elect. Center, Phoenix, Ariz.*

"Reliability Through Component Part Standardization," *H. Allers and P. Magnani, IBM, Owego, N. Y.*

"Redundancy In Non-Repairable Equipment," *S. Steel and C. Hillesdale, Philco, Palo Alto, Calif.*

"A System Reliability Analysis," *R. Lowe, Aeronutronics, Newport Beach, Calif.*

Session 12—Ranging and Tracking

"Polystation Doppler Radar Tracking System," *C. Hammack, Philco, Palo Alto, Calif.*

"Derivations, Discussions, and Error Analysis Pertaining To Ideal Coplanar Elliptic Orbits and Ballistic Trajectories," *C. McClure, IBM, Owego, N. Y.*

"LOCTRACS, Air Traffic Surveillance Systems," *K. Bailey, Lockheed Aircraft Corp., Lockheed Electronics and Avionics Div. (LEAD), Los Angeles, Calif.*

"The Use of Geometrical Invariants in Polystation Doppler Tracking Systems," *C. Dawson, Philco, Palo Alto, Calif.*

Friday Morning, February 5

Session 13—Communications (Confidential)—Sponsor: Air Res. and Dev. Command

"Ballistic Missile Space Communications System Concept," *C. Strom et al., Rome Air Dev. Center, Griffiss AFB, N. Y.*

"Sub-surface Radio Communication Links," *H. A. Norby, Space Electronics, Glendale, Calif.*

"Erectable Antennas For Space Vehicles," *Lockheed Aircraft Corp., Missile and Space Division, Sunnyvale, Calif.*

"Considerations For Inflatable Antennas For The Space Age," *S. H. Saulson and C. N. Gosnell, Westinghouse Electric Corp.*

Session 14—Re-entry

"Re-entry Radiation From An IRBM," *W. Arnquist and D. Woodbridge, Systems Dev. Corp., Santa Monica, Calif.*

"Re-entry Guidance and Flight Path Control," *J. Vaeth, The Martin Co., Baltimore, Md.*

"Re-entry Stability and Control of a Lifting Body," *J. Hinson, The Martin Co., Baltimore, Md.*

Session 15—Telemetry

"High Accuracy Telemetry For SERGEANT Weapon System," *W. Arens, Jet Propulsion Lab., Pasadena, Calif.*

"Recent Advances in Transistorized Telemetry," *J. Smith, Texas Instruments, Dallas, Tex.*

"Maximum Utilization of Narrow Band Data Link For Interplanetary Communications," *W. F. Sampson, Hallamore, Anaheim, Calif.*

"Philco Three-Axis 60 Foot Telemetry and Data Acquisition Antenna," *R. Hundley, Philco, Palo Alto, Calif.*

Session 16—Components

"Resistor Micro-Elements," *Dr. R. C. Laneford, Weston Instruments, Newark, N. J.*

"Thin Ferromagnetic Film Balanced Modulators," *R. Samuels and A. Reed, Iowa State College, Ames.*

"Solid State Modulators For Military Electronics Equipment," *R. Biesele, Jr., Shockley Transistor, Mountain View, Calif.*

"Life Time of Transistors In The Van Allen Belts," *D. Pomeroy, Hughes Aircraft Co., Culver City, Calif.*

Friday Afternoon

Session 17—Electronic Warfare (Confidential)—Sponsor: Air Res. and Dev. Command

"A New Approach To Communication Jamming—Uniformly Distributed Expendable Jammers," *J. Fritchel, General Electric, Utica, N. Y.*

"Jamming of Communication Systems Using FM, AM, and SSB Modulation," *H. Magnuski, Motorola Inc., Chicago Military Electronic Center, Chicago, Ill.*

"Training Personnel In Electronics Countermeasure Techniques," *J. Leskinen, U. S. Navy Training Device Center, Port Washington, N. Y.*

"A Variable Pattern, Variable Polarization Antenna For ECM Applications," *S. R. Jones, Melpar, Falls, Church, Va.*

Session 18—Reconnaissance

"The Application of Doppler Navigation Equipment To Aerial Mapping and All Weather Reconnaissance Missions," *A. Roberts, General Precision Lab., Pleasantville, N. Y.*

"The Evaluation of a Real Time Reconnaissance System," *H. Weitz, Bulova Res. and Dev. Labs., Woodside, N. Y.*

"The Antenna System of the Reconnaissance Satellite," *R. B. MacAskill, Cook Res. Labs., Morton Grove, Ill.*

"Photographic Rectification By Image Scanning," *H. Sandberg, Hycon Mfg. Co., Pasadena, Calif.*

Session 19—Guidance and Control

"Radio Control Receiver Developments in 400 MC Band," *G. Meredith and O. D. Embree, White Sands Missile Range, N. M.*

"A Novel, High Performance, Solid-state Autopilot," *R. Moses and J. Porter, RCA, West Coast Missile and Surface Radar Div., Van Nuys, Calif.*

"A Simple Hyperbolic Coordinate Converter," *E. J. Groth, Motorola Inc., Western Military Electronics Center, Phoenix, Ariz.*

"The Design of a 'Universal' Type of Missile Programmer," and "A Guidance Programmer Utilizing Ferrite-Core Techniques," *D. L. Collins, I. L. Wieselman and J. P. Byrd, Telemeter Magnetics, Inc., Los Angeles, Calif. and Jet Propulsion Lab., Pasadena, Calif.*

Session 20—Ranging and Tracking Subsystems

"Ballistic Camera Synchronization System," *J. Shannon, Electronic Engineering Co., Santa Ana, Calif.*

"The AN/DPN-63 Subminiaturized C-Band Transponder," *L. Diven, Motorola Inc., Western Mil. Elect. Center, Phoenix, Ariz.*

"Precision Phase Measuring Equipment For ULF Frequency Band," *C. Hellman, Philco, Western Dev. Labs., Palo Alto, Calif.*

"Multi-Element Antennas," *C. Palermo, Univ. of Michigan, Willow Run Labs., Ann Arbor.*

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1960 Solid-State Circuits Conference

UNIVERSITY OF PENNSYLVANIA AND HOTEL SHERATON, PHILADELPHIA, PA., FEBRUARY 10–12, 1960

The 1960 Solid-State Circuits Conference will feature eight sessions devoted to broad advances in the field of solid-state device applications and circuits. Forty-four papers will be offered covering new magnetic and semiconductor devices and circuits for digital storage and logic. A number of survey reports on tunnel diodes, thin magnetic films and microwave properties of semiconductors will also be presented.

Twelve informal discussion sessions, conducted by leaders in the solid-state field, will be held on Wednesday and Thursday evenings to give registrants an opportunity to discuss tunnel diodes and their applications, storage techniques, reliability, noise theory, logic circuits, microelectronics, parametric applications and energy conversion.

A 100-page technical digest, containing substantial abstracts and a complement of illustrations of every technical paper, will be furnished without charge to every registrant.

Wednesday Morning, February 10

Session 1—Applications of Tunnel Diodes

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

"Esaki Diodes: A Survey," *G. C. Dacey, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Tunnel Diodes and Their Use as Multi-Junction Circuit Elements," *J. J. Tiemann, General Electric Co., Schenectady, N. Y.*

"The Tunnel Diode as a Logic Element," *M. H. Lewin, A. G. Samusenko and A. W. Lo, RCA, Princeton, N. J.*

"Circuit Principles for Application of Esaki Diodes at Microwave Frequencies," *M. E. Hines, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Tunnel Diode Radio Frequency Amplifier," *E. Miller, H. B. Yin and J. Schultz, RCA, Princeton, N. J.*

"Esaki Diode Logic Circuits," *G. W. Neff, S. A. Butler and D. L. Critchlow, IBM, Yorktown, N. Y.*

Wednesday Afternoon

Formal Opening of Conference

Introductory Comments, *A. P. Stern, General Electric Co., Syracuse, N. Y.*

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

"The Impact of Solid-State Research on U. S. Industry and Technology," *C. G. Suits, General Electric Co., Schenectady, N. Y.*

"Solid-State Circuits Research in Europe," *M. J. O. Strutt, Swiss Federal Inst. of Technology, Zurich, Switzerland.*

Session 2—Thin Magnetic Films for Logic and Memory

Chairman: *E. W. Fletcher, M.I.T., Cambridge, Mass.*

"Thin Magnetic Films: A Survey," *A. V. Pohm, Iowa State College, Ames, Iowa.*

"High-Speed Magnetic-Film Logic," *W. E. Proebster, IBM Res. Lab., Zurich, Switzerland.*

"A Thin Magnetic-Film Shift Register," *K. D. Broadbent and F. J. McClung, Jr., Hughes Res. Labs., Los Angeles, Calif.*

"An Evaporated-Film Cryotron Circuit," *C. R. Smallman, M. L. Cohen, A. E. Slade and J. L. Miles, Arthur D. Little, Inc., Cambridge, Mass.*

Wednesday Evening

Informal Discussion Sessions

"Tunnel Diode Characterization," Moderator: *G. C. Dacey, Bell Telephone Labs., Inc., Murray Hill, N. J.* Panel Members: *D. E. Thomas, Bell Telephone Labs., Inc., Murray Hill N. J.; J. J. Tiemann, General Electric Co., Schenectady, N. Y.; S. L. Miller, IBM, Poughkeepise, N. Y.; H. Somers, RCA, Princeton, N. J.*

"Thin Films for Memory," Moderator: *A. V. Pohm, Iowa State College, Ames, Iowa.* Panel Members: *J. I. Raffel, M.I.T., Lincoln Labs., Lexington, Mass.; S. M. Rubens, Remington Rand, St. Paul, Minn.; D. A. Meier, National Cash Register, Dayton, Ohio; R. A. Tracy, Burroughs Corp., Paoli, Pa.; D. H. Looney, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Reliability Considerations," Moderator: *W. D. Rowe, Westinghouse Elec. Corp., Buffalo, N. Y.* Panel Members: *J. J. Scanlon, Bell Telephone Labs., Inc., Whippany, N. J.; G. T. Ross, RCA, Camden, N. J.; K. Davidson, Texas Instruments, Dallas, Tex.; C. J. Thorton, Lansdale Tube Co., Lansdale, Pa.; G. E. Cunnihan, Stromberg Carlson, Rochester, N. Y.*

"Noise Theory," Moderator: *J. G. Linvill, Stanford University, Stanford, Calif.* Panel Members: *R. P. Rafuse, M.I.T., Cambridge, Mass.; W. E. Dahlke, Telefunken, Ulm, Germany; R. B. Adler, M.I.T., Cambridge, Mass.; D. A. Linden, Stanford University, Stanford, Calif.; P. Penfield, Jr., M.I.T., Cambridge, Mass.; C. S. Kim, General Electric Co., Syracuse, N. Y.*

"Energy Conversion," Moderator: *S. J. Angello, Westinghouse Elec. Corp., Pittsburgh Pa.* Panel Members: *H. Reiner, Standard Elektrik, Stuttgart, Germany; S. R. Hoh, ITT Labs., Nutley, N. J.; P. S. Castro, Lockheed Aircraft Corp., Sunnyvale, Calif.; D. A. Paynter, General Electric Co., Syracuse, N. Y.; R. P. Putkovich, Westinghouse Elec. Corp., Cheswick, Pa.*

"Magnetic Logic," Moderator: *D. C. Engelbart, Stanford Res. Inst., Menlo Park, Calif.* Panel Members: *E. E. Newhall, Bell Telephone Labs., Inc., Murray Hill, N. J.; E. P. Stabler, General Electric Co., Syracuse, N. Y.; A. J. Meyerhoff, Burroughs Corp., Great Valley, Pa.; H. P. Wolff, IBM, Poughkeepsie, N. Y.; W. E. Proebster, IBM, Zurich, Switzerland; T. H. Bonn, Remington Rand Univac, Philadelphia, Pa.; D. Muir, National Cash Register, Hawthorne, Calif.; H. D. Crane, Stanford Res. Inst., Palo Alto, Calif.*

Thursday Morning, February 11

Session 3—Digital Logic

Chairman: *R. A. Henoe, IBM, Poughkeepsie, N. Y.*

"Improvements in Current Switching," *F. K. Buelow, IBM, Poughkeepsie, N. Y.*

"Tunnel-Diode Digital Circuitry," *W. F. Chow, General Electric Co., Syracuse, N. Y.*

"A Modulation-Demodulation Scheme for Ultra High-Speed Computing and Wide-Band Amplification," *W. Eckhardt and F. Sterzer, RCA, Princeton, N. J.*

"Adaptive Logic," *R. J. Domenico and R. A. Henle, IBM, Poughkeepsie, N. Y.*

"Design of a Solid-State Neuron Circuit for Use in Self-Organizing Systems," *C. L. Coates and E. A. Fisch, General Electric Co., Syracuse, N. Y.*

"Fast Logic Using Transistor Current-Routing Techniques," *D. B. Jarvis, L. P. Morgan and J. A. Weaver, Mullard Res. Labs., Salsford, Eng.*

Session 4—Applications of New Devices

Chairman: *H. W. Katz, General Electric Co., Syracuse, N. Y.*

"A Broad-Band, Fast-Acting, Ferrite Switch," *D. E. Allen and N. G. Sakiotis, Motorola, Inc., Phoenix, Ariz.*

"A Solid-State Modulator for Millimeter Waves," *E. T. Harkless and R. Vincent, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Low-Noise Tunnel-Diode Down-Converter Having Conversion Gain," *K. K. N. Chang, W. Y. Pan, H. B. Yin, G. H. Heilmeier, D. J. Carlson and H. J. Prager, RCA, Princeton, N. J.*

"Photorectifier Based on a Combination of a Photoconductor and an Electret," *J. G. van Santen and G. Diemer, N. V. Philips Gloeilampenfabrieken, Eindhoven, The Netherlands.*

"A New High-Speed Effect in Solid-State Diodes," *A. F. Boff, R. Shen, Hewlett-Packard Co., Palo Alto, Calif., and J. L. Moll, Stanford University, Stanford, Calif.*

Thursday Afternoon

Session 5—Information Storage Techniques

Chairman: *A. W. Lo, RCA, Princeton, N. J.*

"The Tunnel Diode as a Memory Element," *J. C. Miller, K. Li and A. W. Lo, RCA, Princeton, N. J.*

"Sub-Microsecond Core Memories Using Multiple Coincidence," *H. P. Schlaeppi and I. P. V. Carter, IBM, Zurich, Switzerland.*

"Fluxlok—A Non-Destructive High-Speed Memory Technique Using Standard Ferrite Cores," *R. M. Tillman, Burroughs Corp., Paoli, Pa.*

"Low Coercive-Force Ferrite Ring Cores for a Fast Non-Destructively Read Store," *G. H. Perry and S. J. Widdows, Royal Radar Establishment, Great Malvern, Eng.*

"Memory for a 40-Megacycle Deltic Correlator," *J. V. Peilland and J. Hesler, General Electric Co., Syracuse, N. Y.*

"A 512-Word, 5-Microsecond, Variable Twistor Store," *D. C. Weller, Bell Telephone Labs., Inc., Murray Hill, N. J., and J. L. Rogers, Space Technology Labs., Los Angeles, Calif.*

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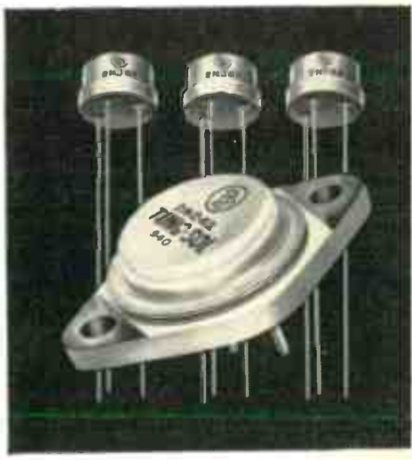

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Read-Out Storage Systems," *W. L. Shevel, Jr. and O. A. Gutwin, IBM, Poughkeepsie, N. Y.*

Session 6—Linear Amplification and Generation

Chairman: *F. H. Blecher, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Circuit Applications of a Coaxially Encapsulated Microwave Transistor," *V. R. Saari, C. A. Bittmann and R. E. Davis, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Tunnel Diode Amplifiers with Unlimited Gain-Bandwidth Products," *E. W. Sard, Airborne Instruments Lab., Melville, N. Y.*

"Superconductive Devices and Electrical Measurements at Low Temperatures," *I. M. Templeton, National Res. Council, Ottawa, Can.*

"Solid-State Sampled-Data Bandpass Filters," *L. E. Franks and F. J. Witt, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Analysis of Transistor Class C Oscillators," *J. A. Narud, Harvard University, Cambridge, Mass.*

Thursday Evening Informal Discussion Sessions

"Tunnel Diode Applications," Moderator: *A. W. Lo, RCA, Princeton, N. J.* Panel Members: *G. W. Neff, IBM, Ossining, N. Y.; M. E. Hines, Bell Telephone Labs., Inc., Murray Hill, N. J.; W. F. Chow, General Electric Co., Syracuse, N. Y.; E. W. Sard, Airborne Instruments Lab., Melville, N. Y.; K. K. N. Chang, RCA, Princeton, N. J.*

"Storage Techniques," Moderator: *J. A. Rajchman, RCA, Princeton, N. J.* Panel Members: *D. C. Weller, Bell Telephone Labs., Inc., Murray Hill, N. J.; W. L. Shevel, IBM, Poughkeepsie, N. Y.; R. M. Tilman, Burroughs Corp., Great Valley, Pa.; W. Peil, General Electric Co., Syracuse, N. Y.; G. H. Perry, Royal Radar Establishment, Great Malvern, Eng.; H. P. Schlaeppi, IBM, Zurich, Switzerland; J. C. Miller, RCA, Princeton, N. J.; J. E. Mack, Bell Telephone Labs., Inc., Whippany, N. J.*

"Microelectronics," Moderator: *G. Sziklai, Westinghouse Electric Corp., Pittsburgh, Pa.* Panel Members: *I. N. Ross, Bell Telephone Labs., Inc., Murray Hill, N. J.; D. C. Engelbart, Stanford Res. Inst., Menlo Park, Calif.; P. B. Myers, Motorola, Phoenix, Ariz.; W. Adcock, Texas Instruments, Dallas, Tex.; H. W. Henkels, Westinghouse Electric Corp., Youngwood, Pa.; R. Norman, Fairchild Semiconductor Corp., Palo Alto, Calif.; A. McWhorter, M.I.T., Cambridge, Mass.*

"Parametric Applications," Moderator: *B. Salsberg, Airborne Instruments Lab., Melville, N. Y.* Panel Members: *R. S. Engelbrecht, Bell Telephone Labs., Inc., Whippany, N. J.; H. Hsu, General Electric Co., Syracuse, N. Y.; L. Neergard, RCA, Princeton, N. J. A. Berk, Hughes Aircraft Co., Culver City, Calif.; K. E. Shreiner, IBM, Yorktown, N. Y.*

"High-Frequency Amplification and Generation," Moderator: *J. B. Angell, Philco Corp., Philadelphia, Pa.* Panel Members: *V. H. Grinich, Fairchild Semiconductor Corp., Palo Alto, Calif.; J. M. Early, Bell Telephone Labs., Inc., Murray Hill, N. J.; F. H. Blecher, Bell Telephone Labs., Inc., Murray Hill, N. J.; C. H. Knowles, Motorola, Phoenix, Ariz.; L. D. Wechsler, General Electric Co., Syracuse, N. Y.; R. L. Pritchard, Texas Instruments, Inc., Dallas, Tex.*

"Semiconductor Logic," Moderator: *W. B. Cagle, Bell Telephone Labs., Inc., Whippany, N. J.* Panel Members: *L. P. Morgan, Mullard Res. Labs., Salsford, Eng.; J. D. Schmidt, General Electric Co., Syracuse, N. Y.; F. K. Buelow, IBM, Poughkeepsie, N. Y.; M. H. Lewin, RCA, Princeton, N. J.; R. F. Schauer, Iowa State College, Ames; J. T. Lynch, Burroughs Corp., Paoli, Pa.*

Friday Morning, February 12
Session 7—Microelectronic Considerations

Chairman: *J. R. Nall, Fairchild Semiconductor Corp., Mountain View, Calif.*

"Microelectronics and the Art of Similitude," *D. C. Engelbart, Stanford Res. Inst., Menlo Park, Calif.*

"Limitations in Multielement High-Speed Logic Systems," *J. M. Early, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Concept of Molecular Engineering," *H. W. Henkels, Westinghouse Electric Corp., Youngwood, Pa.*

"Solid-State Micrologic Elements," *I. Haas, J. T. Last and R. H. Norman, Fairchild Semiconductor Corp., Palo Alto, Calif.*

"Semiconductor Inductance Diode," *J. Nishizawa and Y. Watanabe, Tohoku University, Sendai, Japan.*

Friday Afternoon
Session 8—Parametric Circuit Techniques

Chairman: *H. Heffner, Stanford University, Stanford, Calif.*

"Microwave Properties of Semiconductors: A Survey," *H. Kroemer, Varian Assoc., Palo Alto, Calif.*

"Low-Frequency Reactance Amplifier," *J. R. Biard, Texas Instruments, Inc., Dallas, Tex.*

"A Passive Parametric Limiter," *A. A. Wolf and J. E. Pippin, Sperry Microwave Electronics Co., Clearwater, Fla.*

"Backward Traveling Wave Parametric Amplifier," *Hsiung Hsu, General Electric Electronics Lab., Syracuse, N. Y.*

"An Analysis of Sub-Harmonic Pumping of Parametric Amplifiers," *K. E. Mortenson, General Electric Co., Schenectady, N. Y.*

"Measurement of Absolute Noise Performance of Parametric Amplifiers," *R. P. Rafuse, M.I.T., Cambridge, Mass.*

Professional Groups*

Aeronautical & Navigational Electronics (G-11)—L. M. Sherer, RTCA, Washington, D. C.; H. R. Mimno, Harvard Univ., Cambridge, Mass.

Antennas & Propagation (G-3)—A. Dorne, Dorne and Margolin, Westbury, L. I., N. Y. S. A. Bowhill, Pennsylvania State Univ., University Park, Pa.

Audio (G-1)—Prof. A. B. Bereskin, E.E. Dept., Univ. of Cincinnati, Cinn. 21, Ohio; M. Camras, Armour Res. Found. Tech. Ctr., Chicago 16, Ill.

Automatic Control (G-23)—J. E. Ward, Servomechanisms Lab., M.I.T., Cambridge 39, Mass.; G. S. Axelby, Westinghouse Air Arm Div., Friendship Airport, Baltimore 3, Md.

Broadcast & Television Receivers (G-8)—G. C. Larson, Raytheon Mfg. Co., Industrial Apparatus Div., River Road Plant, Waltham, Mass.; C. W. Sall, RCA, Bldg. 13-4, Camden, N. J.

Broadcasting (G-2)—G. E. Hagerty, Westinghouse Elec. Corp., 122 E. 42 St., Suite 2100, N. Y. 17, N. Y.; W. L. Hughes, Elec. Engrg. Dept., Iowa State College, Ames, Iowa.

Circuit Theory (G-4)—S. Darlington, Bell Telephone Labs., Murray Hill, N. J.; W. Bennett, Bell Tel. Labs., Murray Hill, N. J.

Communications Systems (G-19)—J. E. Schlaikjer, IT&T Co., 67 Broad St., N. Y., N. Y.; M. R. Donaldson, Electronic Comm. Inc., St. Petersburg, Fla.

Component Parts (G-21)—J. J. Drvostep, Sperry Gyroscope Co., Mail Station 1A36, Great Neck, L. I., N. Y.; G. Shapiro, Engr. Electronics Sec., Div. 1.6, NBS, Conn. Ave. & Van Ness St., Washington, D. C.

Education (G-25)—J. G. Truxal, Head, Dept. of E. E., Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; W. R. Le Page, Dept. of E. E., Syracuse Univ., Syracuse, N. Y.

Electron Devices (G-15)—W. M. Webster, RD 2, Cherry Valley Rd., Princeton, N. J.; E. L. Steele, Hughes Prods., Inc., Internatl. Airport Sta., Los Angeles 45, Calif.

Electronic Computers (G-16)—R. O. Endres, Rese Engrg. Co., Philadelphia, Pa.; H. E. Tompkins, Moore School of Elec. Engrg., Univ. of Penn., Philadelphia, Pa.

Engineering Management (G-14)—H. M. O'Bryan, Sylvania Elec. Prods., 730 3rd Ave., N. Y. 17, N. Y.; A. H. Rubenstein, Northwestern Univ., Evanston, Ill.

Engineering Writing and Speech (G-16)—T. T. Patterson, Jr., RCA, Bldg. 13-2, Camden, N. J.; J. Kinn, Electronics, 330 W. 42 St., N. Y., N. Y.

Human Factors in Electronics (G-28)—C. M. Jansky, Royal McBee Corp., Port-chester, N. Y.; J. I. Elkind, Bolt, Beranek and Newman, Cambridge, Mass.

Industrial Electronics (G-13)—J. E. Eiselein, RCA Victor Div., Camden, N. J.; R. W. Bull, Armour Res. Found., Chicago, Ill.

Information Theory (G-12)—Peter Elias, M.I.T., Rm. 26-347, Cambridge 39, Mass.; G. A. Deschamps, Elec. Engrg. Dept., Univ. of Ill., Urbana, Ill.

* Name listed are Group Chairmen and Transactions Editors.

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Instrumentation (G-9)—L. C. Smith, Image Instr. Inc., Newton Lower Falls, Mass.; G. B. Hoadley, Dept. of Elec. Engrg., N. Carolina State College, Raleigh, N. C.

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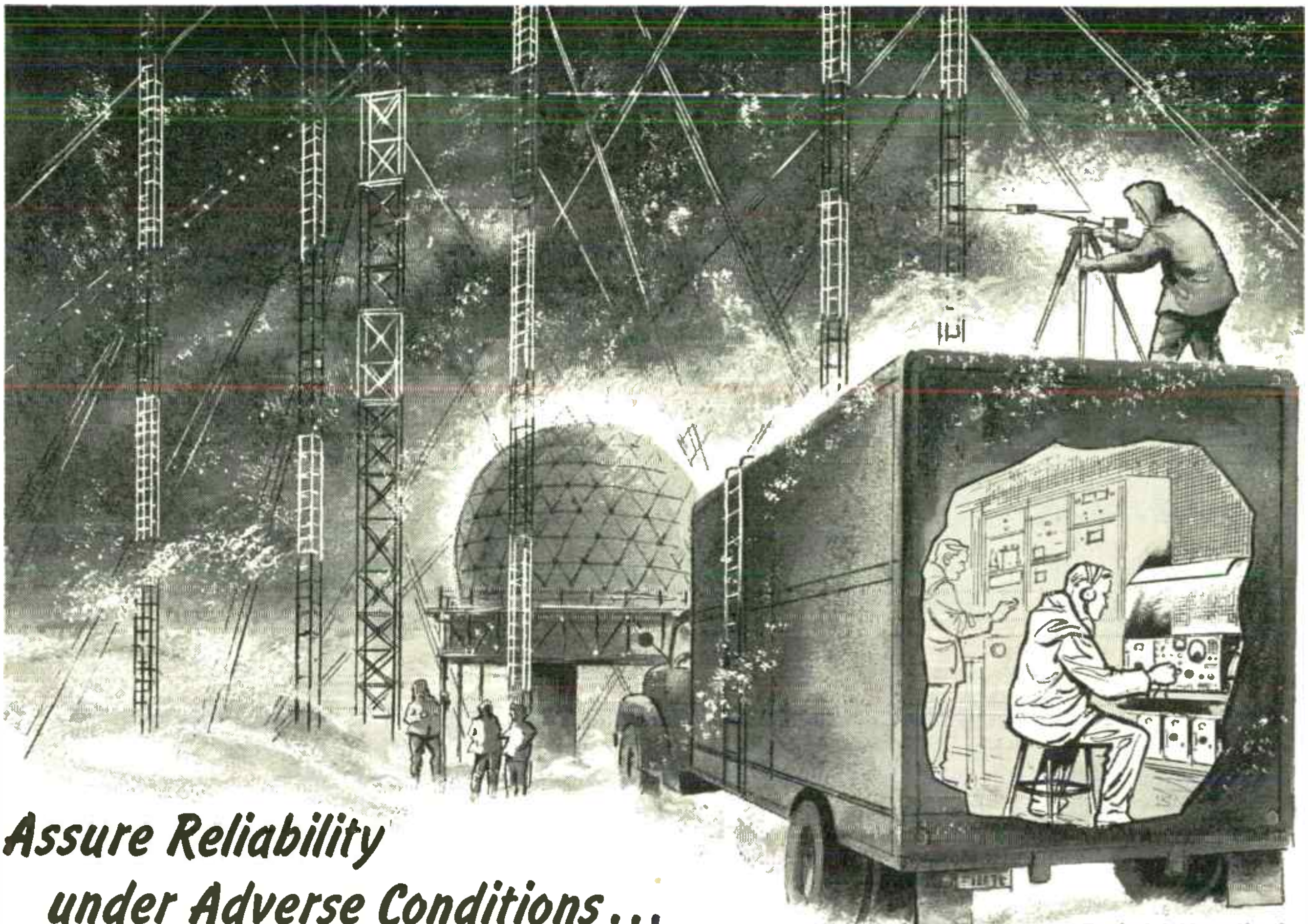

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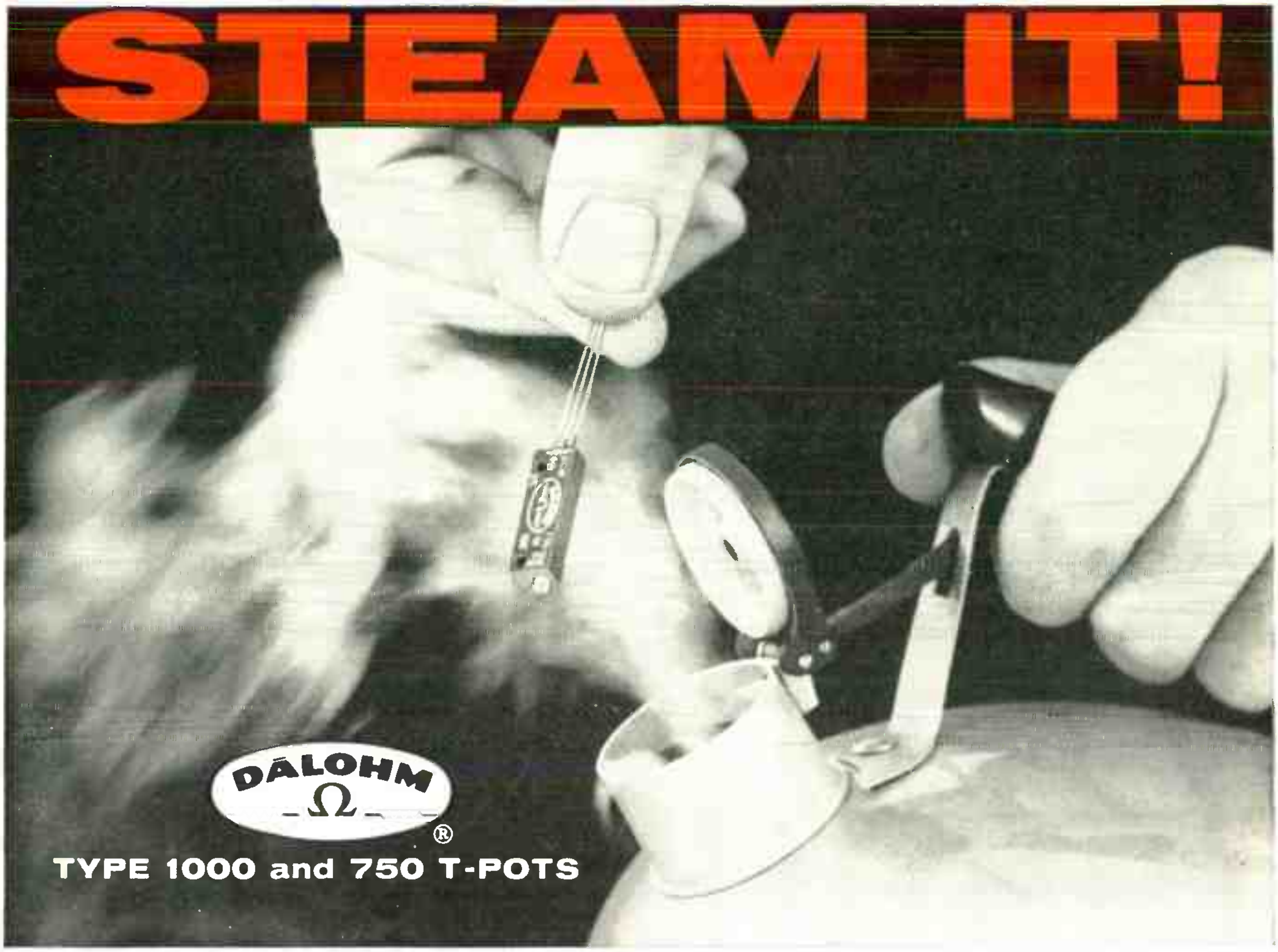

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San Fernando Valley (7)—R. A. Lamm, 15573 Briarwood Dr., Sherman Oaks, Calif.; J. D. Wills, 6606 Lindley Ave., Reseda, Calif.

Santa Ana (7)—T. W. Jarmie, 12345 Cinnabar Rd., Santa Ana, Calif.; R. F. Geiger, Aeronutronic, A Div. of Ford Motor Co., Ford Rd., Newport Beach, Calif.

Santa Barbara (7)—C. P. Hedges, 316 Coleman Ave., Santa Barbara, Calif.; J. A. Moseley, 4532 Via Huerto, Santa Barbara, Calif.

South Western Ontario (8)—W. A. Ruse, Bell Tel. Co., 1149 Goyeau St., Windsor, Ont., Canada; G. L. Virtue, 959 Rankin Blvd., Windsor, Ont., Canada.

Westchester County (2)—M. J. Lichtenstein, 52 Sprain Valley Rd., Scarsdale, N. Y.; M. Ziserman, 121 Westmoreland Ave., White Plains, N. Y.

Western North Carolina (3)—L. L. Caudle, Jr., Box 2536, 1925 N. Tryon St., Charlotte, N. C.; J. I. Barron, Southern Bell T&T Co., Box 240, Charlotte, N. C.

PROVED DESIGN APPLICATION PERFORMANCE

TYPE C Temperature Compensating RMC DISCAPS

TC	1/4 Dia.	5/16 Dia.	1/2 Dia.	5/8 Dia.	3/4 Dia.	7/8 Dia.
P-100	1- 5 MMF	6- 10 MMF	11- 20 MMF	—	—	—
NPO	2-15	16- 33	34- 69	70- 85 MMF	86-115 MMF	116-150 MMF
N- 33	2-15	16- 33	34- 69	70- 85	86-115	116-150
N- 75	2-15	16- 33	34- 56	57- 68	69-125	126-150
N- 150	2-15	16- 33	34- 67	68- 75	76-140	141-200
N- 220	3-15	16- 33	34- 75	76-100	101-140	141-200
N- 330	3-15	16- 47	48- 75	76-100	101-150	151-200
N- 470	3-20	21- 51	52- 80	81-120	121-200	201-250
N- 750	5-30	31- 75	76-150	151-220	221-300	301-470
N-1500	10-56	57-120	121-220	220-300	300-470	471-560
N-2200	20-75	76-150	151-200	201-300	301-680	—

Temperature coefficients up to N-5200 available on special order.

SPECIFICATIONS

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.

RMC Type C DISCAPS meet or exceed all specifications of the ETA standard RS-198. Rated at 1000 working volts, Type C DISCAPS provide a higher safety factor than other paper or mica capacitors.

Constant production checks assure that all specifications and temperature characteristics are met. Another phase of complete quality control consists of 100% testing of capacities.

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

DISCAP CERAMIC CAPACITORS RMC

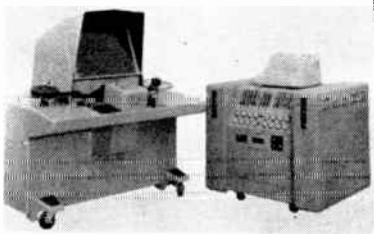
RADIO MATERIALS COMPANY
A DIVISION OF P. R. MALLORY & CO., INC.
GENERAL OFFICE: 3325 N. California Ave., Chicago 18, Ill.
Two RMC Plants Devoted Exclusively to Ceramic Capacitors
FACTORIES AT CHICAGO, ILL. AND ATTICA, IND.

NEWS
New Products

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

Film Reader

A new, semi-automatic film reader with an automatic, high speed electronic digitizing unit which accurately measures distances along two axes on 16mm to 70mm sprocketted film has been announced by **Data Instruments**, a Division of Telecomputing Corp., 12838 Saticoy St., N. Hollywood, Calif.



The Dilog 510 displays a magnified image of the film being measured. The digitizer is an indicating and recording accumulator which counts and stores measurement pulses generated by the reader. These counts are displayed visually, or may be readout on electric typewriter, punched cards, perforated type or x-y plotter.

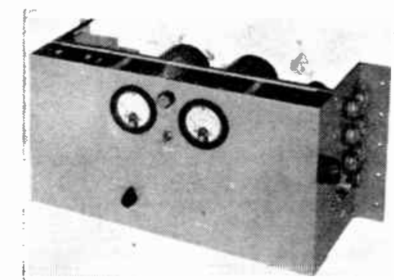
The Dilog 510 has a counting rate of 20,000 counts per second with a maximum storage of 100,000 counts along each axis. Card punching speed is 50 cards per minute, and it will also type up to 600 characters per minute, or will provide 20 columns per second of punched tape with optional output equipment. The optical system provides the operator with a resolution of up to 60 lines per millimeter.

The film reader has a power requirement of 8 amperes at 110 volts ac. The reading unit is 52 inches high by 53.5 inches wide and 36 inches in depth. The separate digitizing unit is 32 inches high by 38 inches wide and 33 inches in depth.

Low Voltage Power Supply

A new, low-voltage, high-current, transistor regulated power supply, designed for ultra-stable output under varying line, load and temperature conditions has been developed by **Power Sources, Inc.**, Burlington, Mass. Type PS4022 operates on an input of 105–125 volts, 57–63 or 380–420 cps ac to produce 4.5 to 9.0 volts dc output with load currents from 0–10 amperes.

This heavy-duty supply provides total regulation of better than ±0.1% change in set output voltage for any combination of input voltage or load current conditions. Total ripple and noise is less than 2.0 mv



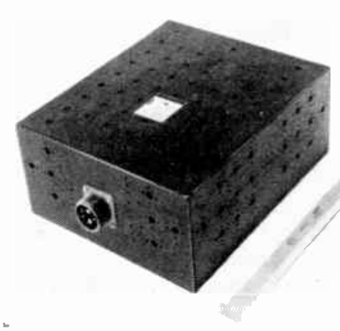
RMS. Each unit has specially designed circuits to prevent damage to components, even when a dead short appears across the output. In addition, temperature stabilization insures minimum drift in the output for operating temperatures from −20° to +65°C.

Output impedance is less than 0.02 ohms from dc to 1000 cps. Voltage across the output is set by a three-step range switch and a vernier potentiometer.

Additional information and complete prices and specs may be obtained from the firm.

2 KVA Solid State Inverter

A new 2 kva static inverter has been completed by **Kinetics Corp**, 410 S. Cedros Ave. Solana Beach, Calif. The three-phase Kinetics unit is one of the largest static inverters ever built using silicon semiconductors. Efficiency of the inverter is greater than 82% at full load and exceeds 77% at only half load. This is said to be the nearest approach yet to a flat efficiency curve for a static inverter. Regulation is ±2% from 30% of load to full load.

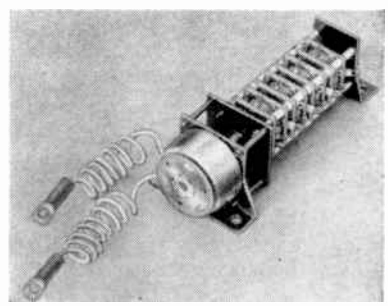


There are no moving parts, no vacuum tubes and no transistors in high current circuits in the inverter. The design features rugged silicon semiconductor elements.

Output is very nearly a perfect sine wave. A high degree of reliability is achieved by the use of rugged components and the lack of complex circuitry. Input voltage of the 2 kva static inverter is 26–29 vdc. Output voltage is 115 vac, ±2%. Power factor is unity to 0.85. The inverter will supply 150% of rated current for 15 minutes. It will supply 200% of rated current for five minutes. The 2 kva Kinetics static inverter is designed to operate under environments imposed by missiles and aircraft, including all the MIL-E-2572 requirements.

Solenoid-Operated Rotary Switch

The type MA-12-S solenoid-operated switches by **Electro Switch Corp., 167** King Ave., Weymouth 88, Mass., provide for remote control of multiple circuits with a compact rotary switch mechanism. Up to five poles, in a tap switch arrangement having twelve taps per pole, can be supplied.



The "Ledex" solenoid coil is available for dc voltages ranging from 6 volts to 230 volts. Self-interrupting contacts as well as suitable rectifiers for ac operation are also available. The solenoid requirements in terms of available power supply, pulsing method (mechanical and/or electrical), frequency of operation, number of operations, and ratio of energized time to deenergized time are arranged to customer specifications.

All current carrying parts are silver-alloy material. Insulation is molded alkyd type MAI-60, per MIL-M-14E. The basic switch is designed to meet MIL-S-3786.

For details write to the firm.

(Continued on page 34A)

Creative Microwave Technology

Published by MICROWAVE AND POWER TUBE DIVISION, RAYTHEON COMPANY, WALTHAM 54, MASS., Vol. 1, No. 8

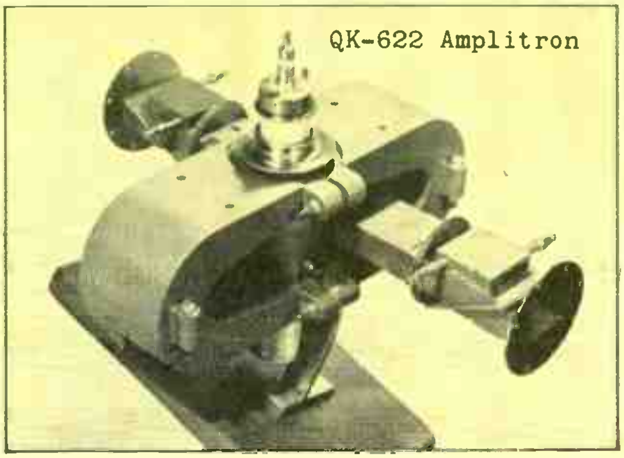
NEW RAYTHEON HEATERLESS AMPLITRONS EXCEED 1,000 HOURS AT RATED POWER OUTPUT

Two new 3-megawatt, S-band Amplitrons have demonstrated an operating life of more than 1,000 hours at rated power output. The QK-622 covers the 2,900 to 3,100 Mc band; the QK-783, the 2,700 to 2,900 Mc band. Both tubes supply full power with low phase pushing characteristics over their entire operating bands at efficiencies greater than 70%--making them unquestionably the most highly efficient microwave tubes thus far developed.

Tubes may be operated at reduced peak power levels to serve as driver stages. High efficiencies are retained at peak power of 600 Kw and gain of 10 db.

Exceptionally long tube life is made possible by the fact that no cathode warmup is required. Starting takes place whenever RF input is present prior to application of modulating pulse. Heater supplies may be omitted entirely from the equipment.

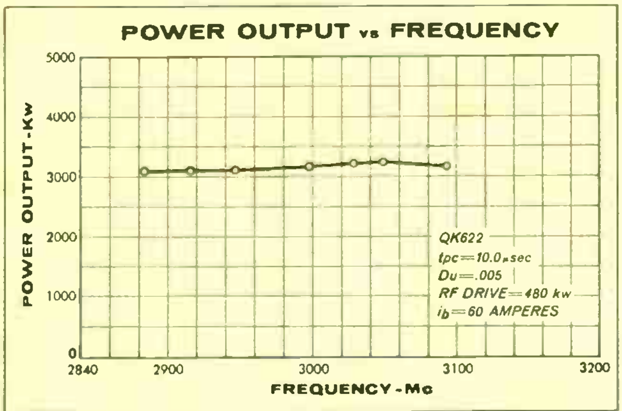
Applications include power-amplifier stages for long-range radars. The tube has been used successfully as an RF power source for linear accelerators.

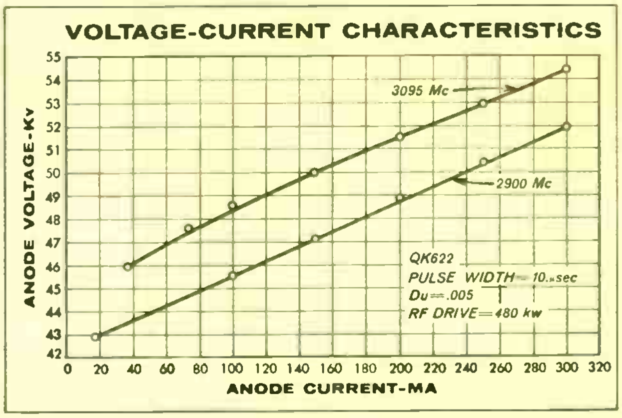


QK-622 Amplitron

Typical Operating Characteristics
(QK622 and QK783 Amplitrons)

Peak Power Output (min.)	3 Mw
Average Power Output	15 Kw
Pulse Duration	10 μ sec
Band Width	200 Mc
Duty Cycle	.005
Pulse Voltage	50-55 Kv
Peak Anode Current	65 amps
Efficiency	70%
RF Input	475 Kw
Weight (with permanent magnet)	125 lbs.





Excellence in Electronics

RAYTHEON

You can obtain detailed application information and special development services by contacting: Microwave and Power Tube Division, Raytheon Company, Waltham 54, Massachusetts

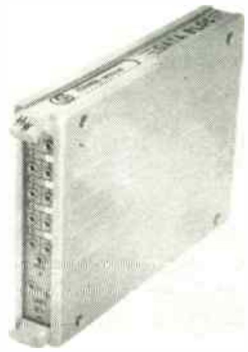
A LEADER IN CREATIVE MICROWAVE TECHNOLOGY

NEWS New Products

(Continued from page 32A)

Four-Stage Shift Register

A new low cost, high-speed, four-stage shift register has been added to the line of digital computer and data handling equipment produced by **Harvey-Wells Electronics, Inc.**, East Natick Industrial Park, Natick, Mass.



This digital component, featuring three inputs and eight outputs, may be cascaded to form a multistage shift register. It has the same electrical characteristics as other units in the Data Bloc and Data Pac lines.

The new unit replaces four standard Flip Flop B units for shift register applications. The advantages of the Model 1801 Shift Register are small size and low cost.

Complete information on this and other units in the high-speed modular building block line is available from the Research and Development Division, Harvey-Wells.

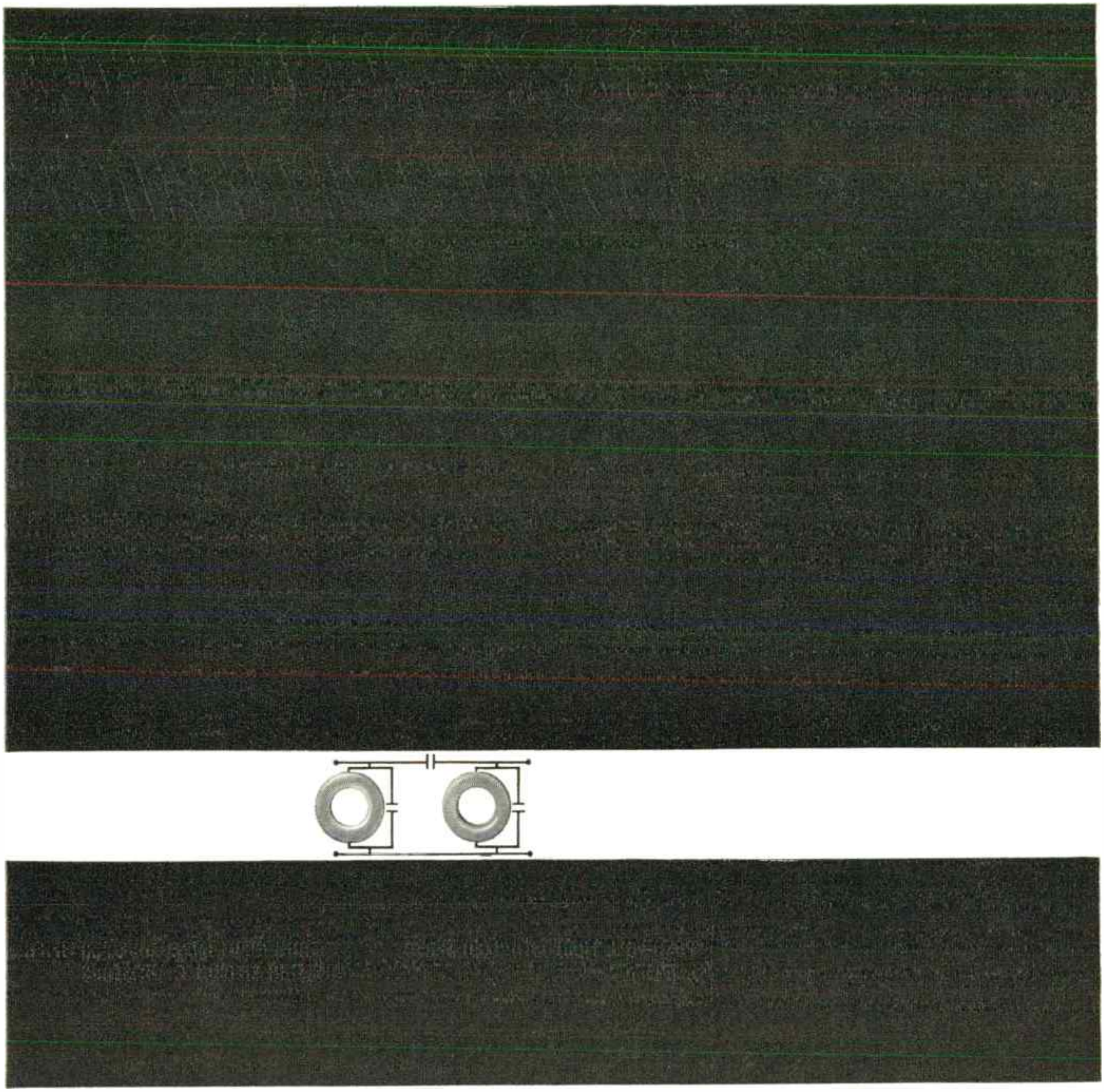
Sperry Semiconductor Appoints Montgomery

Sperry Semiconductor Div. of Sperry Rand Corp., South Norwalk, Conn., has announced the appointment of William M. Montgomery, Jr. as Field Sales Manager. Mr. Montgomery will supervise regional sales offices now being established in major areas throughout the United States.

Montgomery comes to Sperry Semiconductor from his most recent position as Western Regional Sales Manager, Motorola Semiconductor Div. and was previously associated with Texas Instruments as a Sales Engineer. He holds the B.S.E.E. degree from Texas A. & I.

Sperry has already announced the establishment of two regional sales offices, in Chicago and Los Angeles, to promote sales of its silicon diodes and transistors. Other offices are being planned.

(Continued on page 36A)



Smaller filters ease the squeeze!

Filter designers! First 160-mu moly-permalloy powder cores pack high performance into smaller space

Filter and inductor designers specify our 160-mu moly-permalloy powder cores for low frequency applications. Where space is precious, such as in carrier equipment and telemetering filters, the high permeability of these 160-mu cores eases the squeeze.

In many cases, 160-mu cores offer designers the choice of a smaller core. In others, because inductance is 28 percent higher than that of 125-mu cores, at least 10 percent fewer turns are needed to yield a given inductance.

If Q is the major factor, 160-mu cores permit the use of heavier wire with a resultant decrease in d-c resistance.

Like all of our moly-permalloy powder cores, the 160's come with a *guaranteed* inductance. We can ship eight sizes from stock, with a choice of three finishes—standard enamel, guaranteed 1,000-volt breakdown finish, or high temperature finish. Further information awaits your inquiry. *Magnetics Inc., Dept. P-78, Butler, Pa.*

MAGNETICS inc.

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 34A)

Ciochetti SM At Dallons

Tom Ciochetti has been appointed to the position of Sales Manager of **Dallons Semiconductors,** a division of Dallons Laboratories, Inc.

The semiconductor division is located at 5066 Santa Monica Blvd., Los Angeles 29, Calif. The announcement came from the office of Dr. Oscar Dallons, President.

In his new position as Sales Manager of the Semiconductor Division, Ciochetti will be responsible for marketing activities in connection with Dallons' line of solid state devices. The product line includes Premium Quality Silicon Rectifiers in the 50 to 600 PIV range and other related advance design solid state devices.

Ciochetti's background includes a BSEE from the University of Arizona and post graduate work in special instruments and rectifier cathodic protection. Previous experience includes 8 years of industrial and technical sales management in the electronic industry.

Stanton Board Chairman Of American Avionics

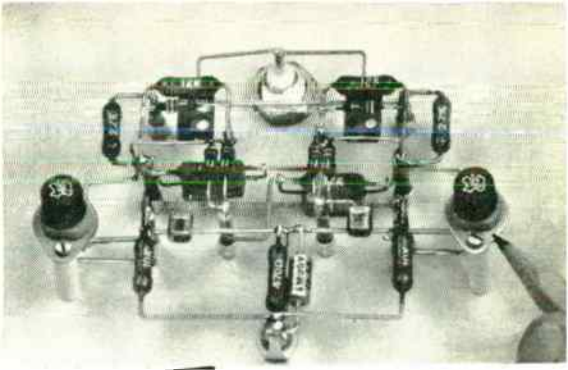
Election of Peter H. Stanton as Chairman of the Board and Chief Executive Officer of **American Avionics, Inc**, Los Angeles electronics manufacturer, was announced by Harold Moss, President.

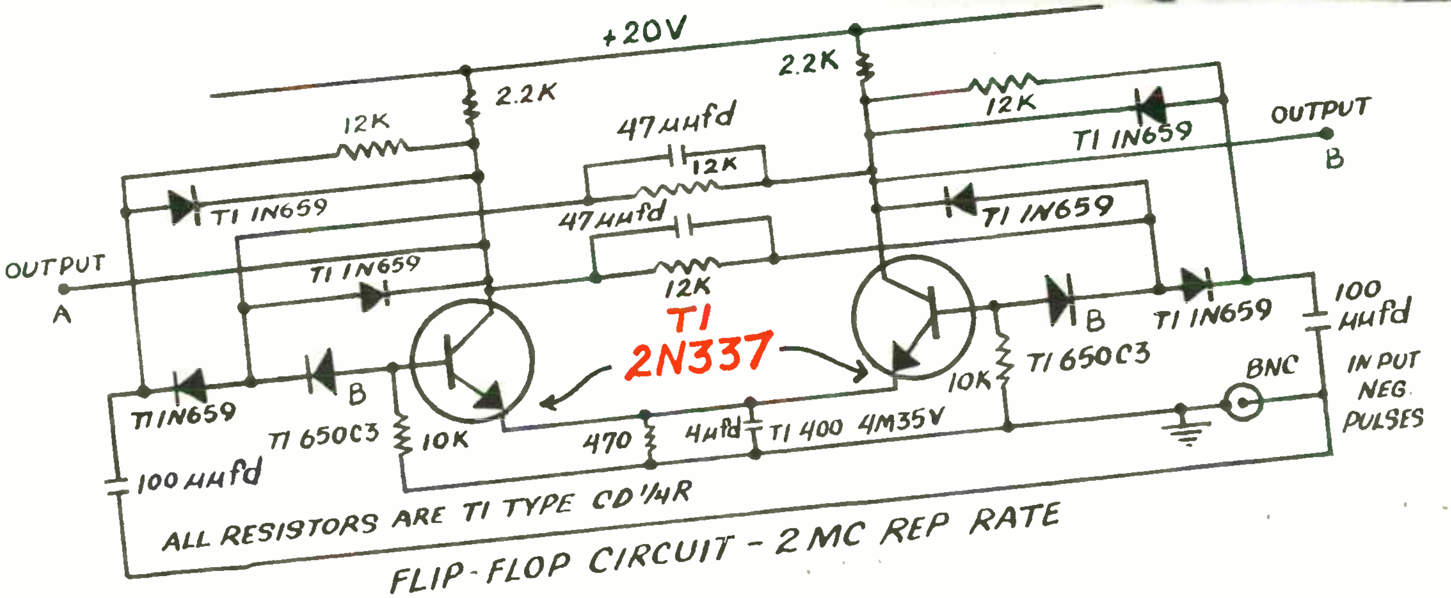
"The addition of an executive of Mr. Stanton's caliber is a signal achievement for our company," Moss stated. He added that Stanton was formerly President of $8,000,000-a-year U. S. Science Corporation, main operating subsidiary of Topp Industries, Inc. Previously, he was Vice President Finance and Administration of Topp Industries and also held key management positions with Lear, Inc. and Giannini Controls Corporation. Stanton has a Master's degree from the Harvard Business School and is a graduate electronics engineer.

Stanton said that he and George Otis, a close associate for many years, have acquired majority control of American Avionics, but that Otis would not be active in its management.

(Continued on page 158A)

HIGH SPEED SWITCHING





. . . with reliable T/I silicon transistors

New improved TI 2N337 and 2N338 specifications provide greater design flexibility for your switching circuits . . . nuclear counters . . . pre-amplifiers . . . RF amplifiers . . . 455 KC IF amplifiers . . . and many other high frequency applications.

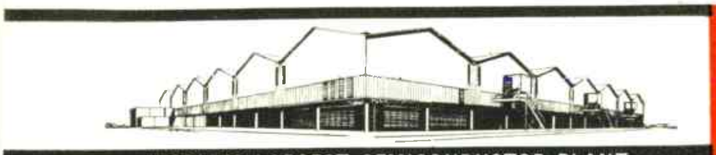
You get high gain at low current levels with TI diffused silicon transistors. High alpha cutoff . . . 10 mc min for 2N337, 20 mc min for 2N338 . . . and extremely low collector capacitance assure optimum performance in your switching and high frequency amplifier applications.

Over four years of mass production and successful use in the most advanced military and industrial applications have proved the value of the TI 2N337 series. Consider TI's guaranteed specs when you select devices for your next transistor circuit. These units are immediately available in production quantities or from large stocks at all authorized TI distributors.

design characteristics at 25° C ambient (except where advanced temperatures are indicated)

		test conditions		2N337 min	2N337 design center	2N337 max	2N338 min	2N338 design center	2N338 max	unit
I_{CBO}	Collector Cutoff Current	$V_{CB} = 20V$	$I_E = 0$	—	—	1	—	—	1	μA
	at 150°C	$V_{CB} = 20V$	$I_E = 0$	—	—	100	—	—	100	μA
BV_{CBO}	Breakdown Voltage	$I_{CB} = 50\mu A$	$I_E = 0$	45	—	—	45	—	—	V
BV_{EBO}	Breakdown Voltage	$I_{EB} = 50\mu A$	$I_C = 0$	1	—	—	1	—	—	V
h_{ib}	Input Impedance	$V_{CB} = 20V$	$I_E = -1mA$	30	50	80	30	50	80	Ohm
h_{ob}	Output Admittance	$V_{CB} = 20V$	$I_E = -1mA$	—	0.2	1	—	0.2	1	μmho
h_{rb}	Feedback Voltage Ratio	$V_{CB} = 20V$	$I_E = -1mA$	—	200	2000	—	300	2000	$X10^{-6}$
h_{fb}	Current Transfer Ratio	$V_{CB} = 20V$	$I_E = -1mA$	0.95	0.985	—	0.975	0.99	—	—
h_{FE}	DC Beta	$V_{CE} = 5V$	$I_C = 10mA$	20	35	55	45	80	150	—
$f_{\alpha b}$	Frequency Cutoff	$V_{CB} = 20V$	$I_E = -1mA$	10	20	—	20	30	—	mc
C_{ob}	Collector Capacitance*	$V_{CB} = 20V$	$I_E = -1mA$	—	1.2	3	—	1.2	3	μμf
R_{cs}	Saturation Resistance†	I_B ‡	$I_C = 10mA$	—	75	150	—	75	150	Ohm
h_{fe}	Current Transfer Ratio	$V_{CB} = 20V$	$I_E = -1mA, f = 2.5mc$	14	22	—	20	24	—	db
t_r	Rise time§			—	0.05	—	—	0.06	—	μsec
t_s	Storage Time			—	0.02	—	—	0.02	—	μsec
t_f	Fall time			—	0.08	—	—	0.14	—	μsec

* Measured at 1 mc † Common Emitter ‡ I_B = 1mA for 2N337, 0.5mA for 2N338 § Includes delay time (t_d)



WORLD'S LARGEST SEMICONDUCTOR PLANT

TEXAS INSTRUMENTS
INCORPORATED
SEMICONDUCTOR-COMPONENTS DIVISION
POST OFFICE BOX 312 · 13500 N. CENTRAL EXPRESSWAY
DALLAS, TEXAS

Industrial Engineering Notes*

Association Activities

A research program sponsored jointly by the American Management Association and approximately thirty industrial companies, including a number of EIA members, has come to the attention of EIA. Called the Group 10 Research Project, the program, recognizing the upward trend in the "non-productive" or overhead and administrative labor categories, is developing new criteria for analyzing costs and results in these areas, William F. E. Long, Manager of the EIA Marketing Data Department, said. It involves both exchanges of manpower utilization data and periodic seminars to discuss findings. "Significant correlations within given industries are being developed, but a larger sample from within the electronics industry is urgently needed. The value to the electronics industry as a whole, of participation of EIA members in this program—particularly in the light of present political and military criticisms that the industry's cost patterns are too high—can hardly be over-emphasized. The additional value to individual companies in such areas as planning staffing levels, exchanging latest control techniques, and comparing company trends to industry averages, is self-evident." Companies that wish to participate (membership in AMA is not required) will be given both the data, the techniques, and the findings developed to date. Inquiries for further information should be made either to the American Management Association, 1515 Broadway, New York 36, N. Y., or to one of the following EIA members who are on the Project's Steering Committee: Lindley K. Nichols, Manager, Management Engineering Studies, Radio Corp. of America, 30 Rockefeller Plaza, New York, N. Y.; or Lawrence I. Marks, Treasurer, Adler Electronics, Inc., 1 LeFevre Lane, New Rochelle, N. Y. **The 1959–60 edition of the EIA Membership List and Trade Directory has been published and mailed to the first and second representative of each member company, according to Executive Vice President James D. Secrest.** As heretofore, the directory contains listings of all member companies, together with their principal executives and electronic products. It also lists all EIA officers, committees, divisions, sections, and departments. Trade names of member companies also are included. Copies of the directory will be distributed widely this week among appropriate Federal agencies, trade press, foreign embassies, and the like. Nonmembers of the Association may purchase copies of the directory at $2.50 each.

* The data on which these Notes are based were selected by permission from *Weekly Report*, issues of October 26 and November 2, 9 and 16, 1959, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

Government and Legislative

Based on expressed needs of the Federal Aviation Agency and the OCDM for additional freequencies to permit greater air safety and earth-space and space communications, the FCC has issued a proposed amendment to Parts 2 and 9 of its rules to reallocate frequencies in the band 118–136 mc. The proposed changes would become effective July 1, 1960. The Commission said the proposal would provide five megacycles of additional spectrum space for air traffic control with the frequency range of existing equipment. Three megacycles, 132–135, would come from the present government band 132–144 mc and would be reallocated for use by government and non-government aeronautical and aircraft station. The band 135–136 mc would be reallocated for the (1) earth-space and (2) space services as a joint government and non-government band. Also, use of this band for government fixed and mobile services and its continued use by government radio positioning service would be permitted on a noninterference basis, the FCC said. **The recently published report to the Navy which calls basic research "the life blood of the entire system of technological innovation" and advises the Navy to do more of it has just been made available to the public through the Office of Technical Services, Commerce Department.** The two volume study, basic research in the Navy, is based on findings of a two year investigation of basic research effort by U. S. industry, government, and universities. It was prepared for the Naval Research Advisory Committee, the Navy's foremost public advisory group on research, by Arthur D. Little, Inc. The Office of Naval Research sponsored the work. The report, numbered PB 151925, is priced at $7 for the two volumes. The NRA committee has said the story "breaks new ground in applying new methods for measuring the amount of basic research that is done, illustrating the mission orientation of basic research in the Navy, and depicting the anatomy of basic research. "The story we feel deserves to be called to the attention of many segments of the public in order that they might benefit from this 'research upon research' which was done with public funds." Copies are available from the OTS, Commerce Department, Washington 25, D. C., at the reported price per copy. Order by number.

Industry Marketing Data

The number of radios produced in September totaled 1,981,208, including 717,501 automobile receivers, compared with 1,009,423 radios made in August in-

(*Continued on page 40A*)

For Capacitors with **GREATER RELIABILITY . . .**

Choose

El-Menco

The Capacitors You Find Wherever There's Electronics!

EL-MENCO DUR-MICA CAPACITORS

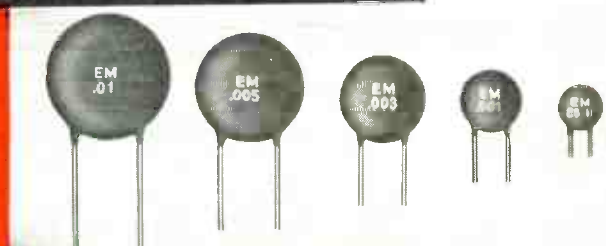
Only 1 Failure Per 43,000,000 Unit-Hours!

- It has been computed that "debugged" DM30, 10,000 MMF units, when subjected to 257,000 hours of life at 85°C with 100% of the rated DC voltage applied, will yield only 1 FAILURE PER 43,000,000 UNIT-HOURS!
- DM15, DM16, DM19, DM20 . . . perfect for miniaturization and for new designs using printed wiring circuits. Also available in DM30, DM42 and DM43.
- New "hairpin" parallel leads insure easy application.
- Exceed all electrical requirements of military specification MIL-C-5A.

EL-MENCO CERAMIC DISC CAPACITORS

Toughest Ever!

- Available in 500 working volts DC and 1,000 working volts DC ratings.
- Low-loss phenolic coating that is wax impregnated.
- Flat design assures reduced self-inductance . . . particularly adaptable to very high frequency applications.
- Insulation resistance far exceeds the 10,000 megohms minimum requirement.
- Exceed all electrical requirements of E.I.A. specifications RS-198.



EL-MENCO *MYLAR-PAPER DIPPED CAPACITORS

Only 1 Failure in 7,168,000 Unit-Hours

- Life tests at 100°C with rated voltage applied have yielding only 1 FAILURE PER 716,800 UNIT-HOURS for 1 MFD. Since the number of unit-hours of these capacitors is inversely proportional to the capacitance, 0.1 MFD Mylar-Paper Dipped capacitors will yield only 1 FAILURE PER 7,168,000 UNIT-HOURS!
- Working volts DC: 200, 400, 600, 1000 and 1600.
- Durez phenolic resin impregnated.
- Tolerances: ± 10% and ±20% (closer tolerances available).
- Dielectric strength: 2 or 2½ times rated voltage, depending upon working voltage.
- Exceed all electrical requirements of E.I.A. specification RS-164 and military specifications MIL-C-91A and MIL-C-25A.

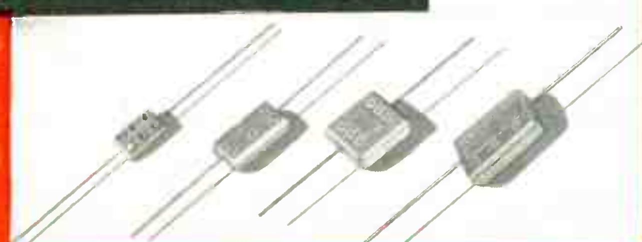
*Registered Trademark of DuPont Co.

EL-MENCO MOLDED MICA CAPACITORS

Superior Performance!

- Unmatched for excellent stability, dielectric strength, high insulation resistance, extremely high "Q" and correspondingly low power factor.
- Units can be subjected to a short "debugging" life test at elevated voltage and temperature for removal of early life failures and for improved reliability.

Write for Free Samples and Booklets on Any of The Above Capacitors



El-Menco *Capacitors*

EL-MENCO OFFERS A COMPLETE LINE OF CAPACITORS ... STANDS READY TO SERVE ALL YOUR CAPACITOR NEEDS.

THE ELECTRO MOTIVE MFG. CO., INC.

Manufacturers of El-Menco Capacitors

WILLIMANTIC CONNECTICUT

• molded mica • dipped mica • mica trimmer • dipped paper
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Section Meetings

Akron

"Traveling Wave Amplifiers in Missile-borne Receivers," C. E. Denton, Bell Telephone Labs, 10/20/59.

Alamogordo-Holloman

"How It All Began," Capt. J. L. Reinartz, Eitel McCollough, 10/19/59.

Albuquerque-Los Alamos

"Activities of the New Mexico Science Center and Museum, Inc.," Dr. Alma Wittlin, New Mexico Science Center and Museum, 10/27/59.

Atlanta

"The RC-130A Aerial Mapping System," Robert Duggan, Lockheed Aircraft Corp. 10/23/59.

Baltimore

"Baltimore's Tallest TV Story," J. T. Wilmer, WBAL-TV; Carl Nopper, WMAR TV; Ben Wolfe, WJZ-TV, 10/12/59.

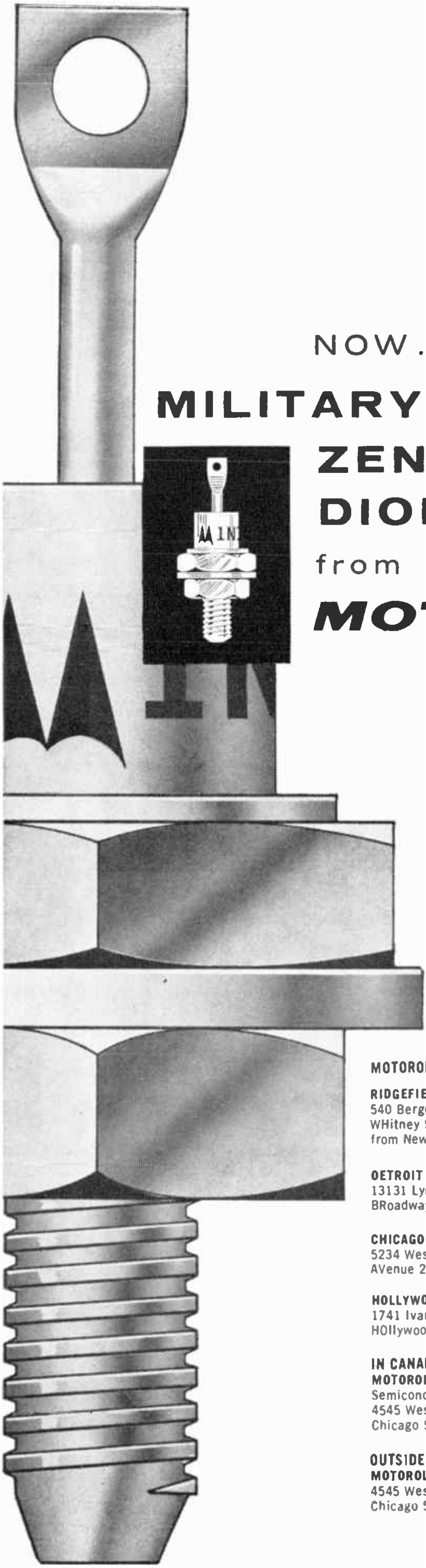
"Solion, Electromechanical Devices," N. M. Potter, National Carbon Co. 11/9/59.

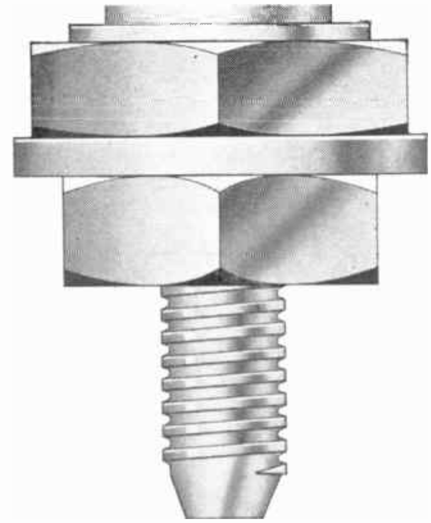
Bay of Quinte

"Whistler Propagation," R. E. Pugh, Queens University, Ontario, 10/21/59.

"Magnetic Amplifier," A. G. Carter, Canadian Westinghouse Co., Ltd. 11/18/59.

(Continued on page 104A)





NOW . . .

MILITARY TYPE ZENER DIODES from MOTOROLA

Three new silicon diffused-junction zener diodes, produced to military specifications, are now available from Motorola. These precision components are designed for highest reliability under the toughest of environmental conditions.

1N1353 MIL-E-1/1236 (Sig C)	10 watts @ 125°C stud temperature	at nominal 12 volts
1N1358 MIL-E-1/1236 (Sig C)	10 watts @ 125°C stud temperature	at nominal 20 volts
1N1361 MIL-E-1/1236 (Sig C)	10 watts @ 125°C stud temperature	at nominal 27 volts

Since these units meet or exceed MIL-E-1/1236 specs, they make possible the utilization of versatile zener diodes in military equipment. Typical applications include:

- regulation of vacuum tube heaters
- surge protection
- arc reduction
- improved circuit fusing
- voltage regulation
- improved coupling and filtering
- more reliable relay operation

Useful in both AC and DC circuits, their small size and light weight make them particularly adaptable to miniature equipment.

TEMPERATURE COMPENSATED ZENERS ALSO AVAILABLE FROM MOTOROLA

For applications demanding a high degree of stability under temperature extremes, Motorola now supplies zener diodes (types 1N2620 through 1N2624) with warranted temperature coefficients as low as .0005% per °C.

FOR COMPLETE SPECIFICATIONS

on the more than 950 Motorola zener diode types contact your nearest Motorola Semiconductor office. Engineering quantities are available from the 22 authorized Motorola Semiconductor Distributors.

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540 Bergen Boulevard
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5234 West Diversey Avenue
AVenue 2-4300

HOLLYWOOD 28, CALIFORNIA
1741 Ivar Avenue
HOllywood 2-0821

IN CANADA WRITE:
MOTOROLA, Inc.
Semiconductor Products Division
4545 West Augusta Boulevard
Chicago 51, Illinois

OUTSIDE USA & CANADA WRITE:
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MOTOROLA SEMICONDUCTORS

Motorola, Inc., Semiconductor Division
5005 East McDowell, Phoenix, Arizona

- *Greatest versatility!*
- *Highest accuracy!*
- *Widest range!*

NARDA transistorized
POWER METER

MODEL 440 . . . **$250**

What's most important to you in a power meter? Accuracy? Portability? Independence from line voltage deviations? Wide range? Stability? Rapid warm-up?

Not that you have to make a choice . . . or a compromise . . . any longer. The Narda Model 440 Power Meter gives you all these features! Completely transistorized and powered by a nickel-cadmium battery, rechargeable during operation or overnight, it offers two low-power scales in addition to the five standard scales (see below), a built-in charger with state-of-charge indicator and protection against overcharging, and freedom from internal heating caused by vacuum tubes.

Moreover, the 440 provides up to 18 ma bias current, enabling you to use the widest selection of bolometers and thermistors. In short, the 440 is the most versatile unit available to provide accurate direct-reading measurements of cw or pulsed-power automatically, over any frequency range for which there are bolometer or thermistor mounts. For complete data, contact your nearest Narda representative, or write us directly. Address: Dept. PIRE-11.

SPECIFICATIONS

POWER RANGES: 7 SCALES

***0.01 mw full scale**	−30 to −20dbm
***0.03 mw full scale**	−25 to −15dbm
0.1 mw full scale	−20 to −10dbm
0.3 mw full scale	−15 to −5dbm
1.0 mw full scale	−10 to 0dbm
3.0 mw full scale	−5 to +5dbm
10 mw full scale	0 to +10dbm

*4.5 ma bolometers give best results on these scales.

Range Switch: 0.01 to 10 mw (full scale)

Accuracy: 3% of full scale reading

Bolometers & Thermistors: All 100 and 200 ohm, requiring up to 18 ma bias.

Battery Charger: Built-in; continuous or overnight. (Battery operable 16 hrs. before recharge required.)

the narda microwave corporation

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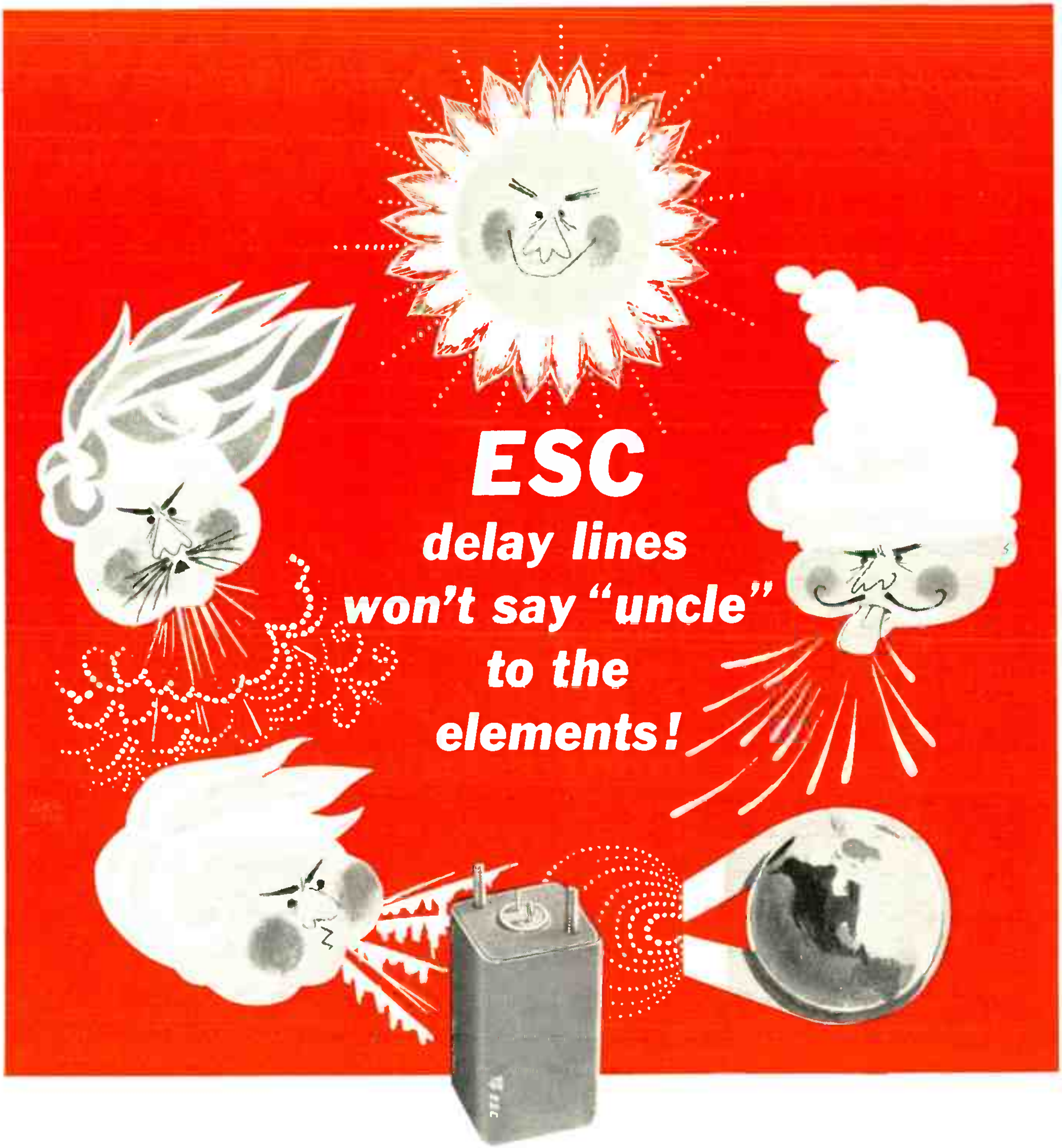

In ESC's environmental testing laboratories, the most grueling elements in the world are unleashed against finished delay lines—temperatures from −55°C to +125°C, sand and dust storms, 100 g shock, vibration, 100% humidity at elevated temperatures. And through them all, the rugged, precise ESC delay lines continue to function perfectly . . . never say "uncle".

Merciless, meticulous testing is just one more reason why ESC is the world's leading producer of custom-built and stock delay lines. Write today for a complete technical data file!

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exceptional employment opportunities for engineers experienced in computer components... excellent profit-sharing plan.

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Distributed constant delay lines • Lumped-constant delay lines • Variable delay networks • Continuously variable delay lines • Pushbutton decade delay lines • Shift registers • Pulse transformers • Medium and low-power transformers • Filters of all types • Pulse-forming networks • Miniature plug-in encapsulated circuit assemblies

Section Meetings

(Continued from page 42A)

CEDAR RAPIDS

Annual IRE picnic. 6/20/59.
Symposium on Antennas and Propagation 9/18–19/59.

"Communication and Navigation in Global Flight," E. S. Pedersen, Scandinavian Airlines System. 10/14/59.

CINCINNATI

"Problems and Solutions of Jet Engine Noise," R. Lee, General Electric Co.; "Demonstration of Stereo and Hi-Fi Techniques and Equipment," A. Rice, Hi-Fi Audio, Inc. 10/20/59.

CLEVELAND

"Project Mercury," W. H Hunter, NASA; Guided tour through NASA Facilities. 11/12/59.

COLUMBUS

"The Space Age and the IRE," Dr. Ernst Weber, IRE President. 10/15/59.

CONNECTICUT

"What we Learned with Satellites during IGY," Dr. J. P. Hagen, National Aeronautics & Space Administration. 10/15/59.

DALLAS

"Infrared Detectors," Dr. H. Levinstein, Syracuse University. 10/20/59.

"Acoustical Research in the South African National Physical Labs," Dr. J. P. A. Lochner, South African Council of Scientific Research. 10/30/59.

"Thermoelectric Energy Conversion," Dr. Paul Egli, Naval Research Lab. 11/17/59.

(Continued on page 50A)

Section Meetings

(Continued from page 46A)

DENVER

Guided tour of Thompson-Ramo-Wooldridge Inc. 10/15/59.

DETROIT

"Application of Electronic Techniques in Medicine," Dr. V. K. Zworykin, RCA Labs. 10/20/59.

"Basic Operation of the Plasma Thermocouple," J. R. Reitz, Case Institute of Technology, "Thermoelectric Power Conversion," Dr. R. P. Ruth, Bendix Aviation Corp. 11/20/59.

EL PASO

"Why" Showing of Movie of "Fire at Our Lady of Angels School," Chicago, Speaker Roy Shanks, El Paso Fire Dept. 10/22/59.

EMPORIUM

"Future Aspects of Space Technology," Dr. P. A. Castruccio, Washington Air Arm Div. 9/15/59.

"Medical Electronics," Dr. W. A. Shafer, St. Vincent Hospital, Erie, Pa. 10/20/59.

EVANSVILLE-OWENSBORO

"Radio Astronomy," W. R. Iliff, Collins Radio 10/14/59.

"Superconducting Devices and Associated Circuits," Dr. D. R. Young, IBM. 11/11/59.

FLORIDA WEST COAST

"Semiconductor Survey," Arthur Barko, General Electric Co. 10/21/59.

FORT HUACHUCA

"Electromagnetic Interference," F. J. Nichols, Genistron Co. 10/26/59.

FORT WAYNE

"Preliminary Design of Communication Satellite Systems," A. T. Mayle, I. T. T. Laboratories. 10/8/59.

FORT WORTH

"Analog to Digital Converters," Robert Knapp, Rensselaer Institute. 8/25/59.

"New Materials and Methods in Telephone Switching," R. A. LeMond, Southwestern Bell Telephone Co. 10/15/59.

"Component Quality Assurance in Reliability Programs," Dr. A. W. Wortham, Texas Instruments 11/10/59.

GAINESVILLE

"University of Florida Linear Accelerator," W. F. Fagen, University of Fla. 10/14/59.

"Electronic Switching," Dr. W. Chen, University of Florida. 11/11/59.

HAMILTON

"Education for the Electronic Age," Dr. Ernst Weber, IRE President. 6/26/59.

"Video Tape Recording," W. E. Jeynes, CHCH-TV, Hamilton, Ontario, 9/14/59.

HOUSTON

"Radio Astronomy," Dr. W. W. Salisbury, Varo Mfg. Co. 10/20/59.

HUNTSVILLE

"Parametric Amplifiers," R. C. Haraway, ARGMA. 11/18/59.

INDIANAPOLIS

"Pocket Size Police Radar System," G. C. Hopkins, Electrofab Co. 10/19/59.

(Continued on page 54A)

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**

For all power supply needs through 1.5 amperes:

LAMBDA
COM-PAK®
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 magnetic components and capacitors for long trouble-free service.

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

OVERLOAD PROTECTION AC and DC fuses; built-in blown-fuse indicators.

SEND FOR COMPLETE 32-PAGE CATALOG

Contains full information and specifications on Lambda's full line of transistor-regulated and tube-regulated power supplies.

ALL LAMBDA POWER SUPPLIES ARE GUARANTEED FOR FIVE YEARS.

LAMBDA ELECTRONICS CORP.

11-11 131 Street, College Point 56, N. Y.

Announcing . . .

SILICON RECTIFIERS

from

DELCO RADIO

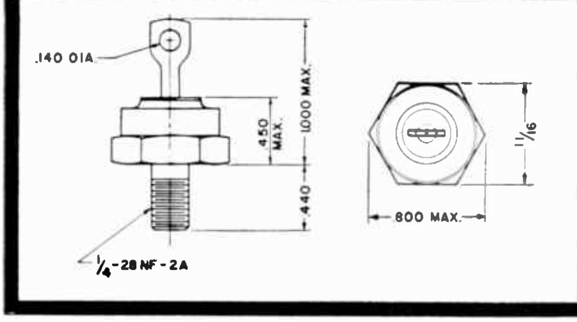
High Quality

High Performance

Extreme Reliability

From the leading manufacturer of power transistors, new Silicon Power Rectifiers to meet your most exacting requirements. Even under conditions of extreme temperatures, humidity and mechanical shock, these diffused junction rectifiers continue to function at maximum capacity! Thoroughly dependable, completely reliable—new Delco Rectifiers are an important addition to Delco Radio's high quality semiconductor line.

Conservatively rated at 40 and 22 amperes for continuous duty up to case temperatures of 150°C.



TYPE	AVG. DC CURRENT	PIV	NORMAL MAX. TEMP.	MAX. FORWARD DROP	MAX. REVERSE CURRENT
1N1191A	22A	50V	150°C	1.2V at 60 amps.	5.0 MA
1N1192A	22A	100V	150°C	1.2V at 60 amps.	5.0 MA
1N1193A	22A	150V	150°C	1.2V at 60 amps.	5.0 MA
1N1194A	22A	200V	150°C	1.2V at 60 amps.	5.0 MA
1N1183A	40A	50V	150°C	1.1V at 100 amps.	5.0 MA
1N1184A	40A	100V	150°C	1.1V at 100 amps.	5.0 MA
1N1185A	40A	150V	150°C	1.1V at 100 amps.	5.0 MA
1N1186A	40A	200V	150°C	1.1V at 100 amps.	5.0 MA at 150°C case temperature and rated PIV

For full information and applications assistance, contact your Delco Radio representative.

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1180 Raymond Boulevard
Tel: Mitchell 2-6165

Chicago, Illinois
5750 West 51st Street
Tel: Portsmouth 7-3500

Santa Monica, California
726 Santa Monica Boulevard
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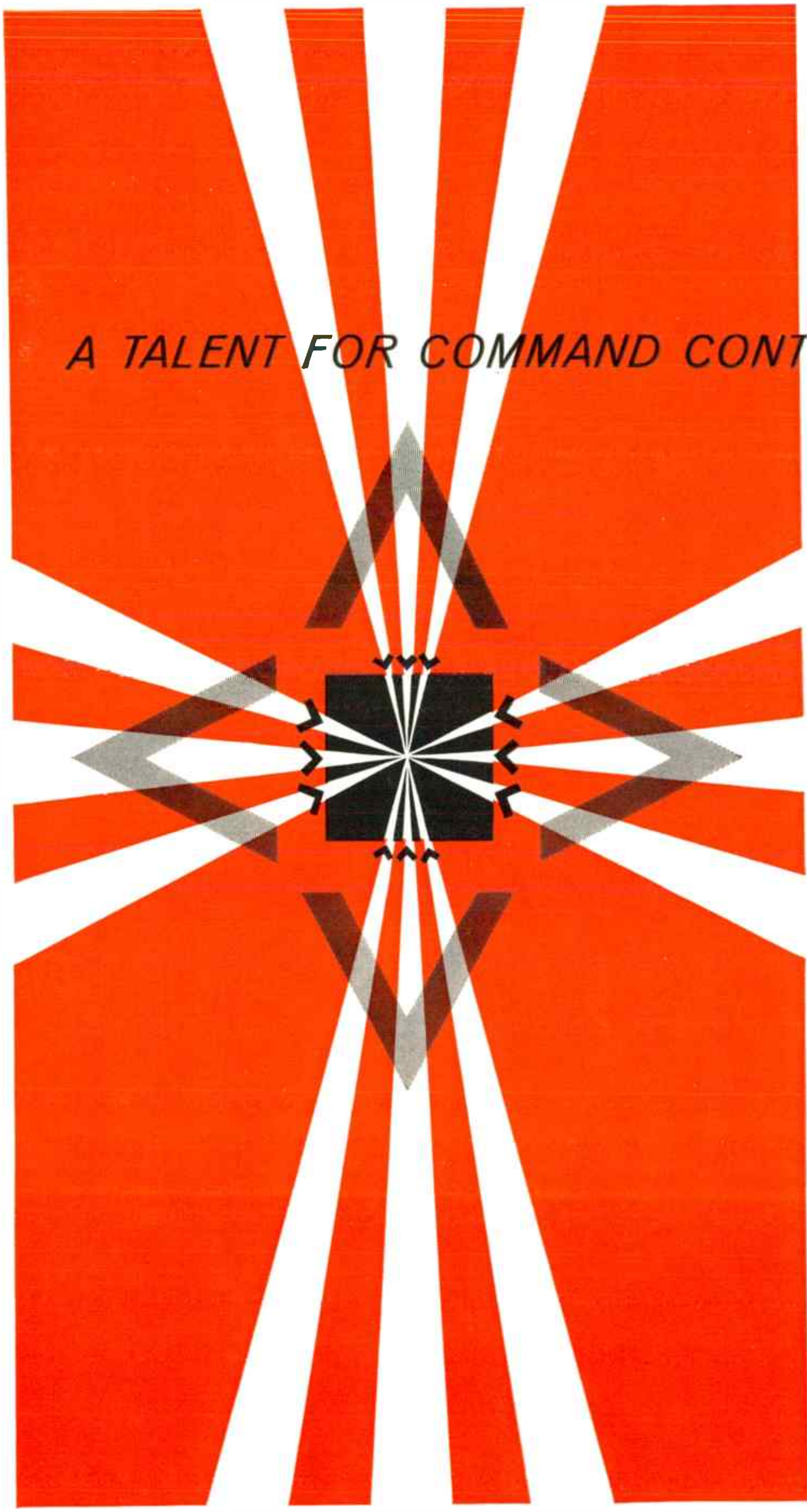
Division of General Motors • Kokomo, Indiana

DELCO RADIO

A TALENT FOR COMMAND CONTROL SYSTEMS



TIME . . . vital element in our nation's defense.

Response to attack must be immediate. The higher speed and greater range of more lethal weapons demand that Commanders of future military systems have a new degree of control that only electronics can provide.

Before the command decision is possible, a mass of information must be gathered from far-flung defense elements, correlated, and presented to the Commander and his staff in readily understandable form . . . this is the function of the Command Control System.

The speed and efficiency of this system must be such that it shall never have to be used of necessity.

Stromberg-Carlson advanced developments in electronic transmission, processing and display of intelligence . . . the nucleus of a proven talent for Command Control Systems.

Brochure on request.

STROMBERG-CARLSON
A DIVISION OF **GENERAL DYNAMICS**
1400 NORTH GOODMAN STREET • ROCHESTER 3, N. Y.

Section Meetings

(Continued from page 50A)

LITTLE ROCK

"Design and Development of Analog Computers," J. H. Hoseman, Electro-Precision, Inc. 11/16/59.

LONDON (ONTARIO)

"High Frequency PNP Diffused Base Transistors," Dr. R. E. Davis, Bell Telephone Labs. 11/9/59.

LONG ISLAND

"Electronic Computers in Control Systems," Dr. J. Truxal, Polytechnic Institute of Brooklyn. 10/20/59.

LOS ANGELES

"Frontiers at Infinity," Prof. A. C. B. Lovell, University of Manchester, England. 10/1/59.

LUBBOCK

"An Engineering Problem Solution on a Digital Computer," J. P. Craig, Texas Technological College. 10/27/59.

MIAMI

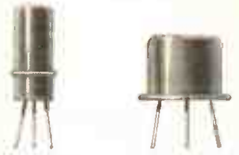
"Fish Noises," Roger Dana, University of Miami Marine Lab.; "New Semiconductor Developments," J. W. Keller, Cordis Corp. 11/4/59.

MONTREAL

Plant tour of the Northern Electric Company Semi-Conductor Lab. Comments by J. A. McDonald, Northern Electric Co., Ltd. After tour films on "Crystals" Bell Telephone Labs, and "Semi-conductor Physics," Dr. Walter Brittain. 10/21/59.

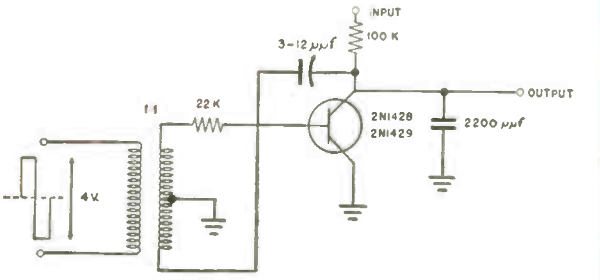
(Continued on page 58A)

PHILCO ANNOUNCES...
HIGHER FREQUENCY...HIGHER BETA
with Extremely Low I_{CO} in 2 New SILICON TRANSISTORS

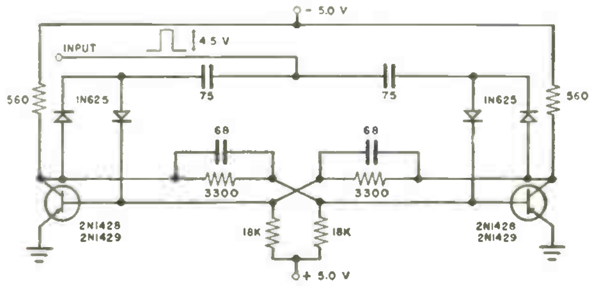


SAT* 2N1428 . . . 2N1429

Stable at Temperatures up to 140° C



LOW LEVEL (<1MV) CHOPPER



1 MC BINARY STAGE (−55°C TO +125°C)

These two new Philco PNP Silicon Surface Alloy Transistors are designed for general purpose high frequency amplifying and switching applications. They offer extremely low saturation resistance (approaching that of germanium transistors) . . . high f_T . . . low leakage current . . . good inverse Beta . . . low offset voltage . . . and excellent frequency response. The combination of high Beta and low I_{CO} makes these transistors excellent for use in DC amplifiers, low level choppers and other critical control circuits. They are suitable for:

. . . high speed switches, operating at speeds up to 5 mc.
. . . general purpose high frequency amplifiers.
. . . DC amplifiers.
. . . high input impedance low frequency amplifiers and choppers.

These two transistors are electrical equivalents, but offer a choice of packages . . . the popular small TO-1 package and the standard TO-5 package. Designers of industrial control and test equipment will find that they deliver excellent performance at high ambient temperatures.

Write Dept. IR-160 for complete information and application data.

*SAT . . . trademark PHILCO Corporation for Surface Alloy Transistor

Absolute Maximum Ratings

Storage Temperature	−65 to +140° C
Junction Temperature	+140° C
Collector to Base Voltage, V_{CB}	−6 volts
Collector to Emitter Voltage, V_{CEO}	−6 volts
Collector Current, I_C	−50 ma
Total Device Dissipation at 25° C (Note 2)	100 mw

Immediately Available in Production Quantities . . . and from 1 to 99 from your local Philco Industrial Semiconductor Distributor.

PHILCO®

LANSDALE DIVISION • LANSDALE, PENNSYLVANIA

TRANSISTOR CENTER U.S.A.®

DESIGN TO-DAY FOR RELIABLE PERFORMANCE THROUGH THE YEARS TO COME

Today's circuits must operate reliably and precisely over long periods of time and under a great variety of exacting conditions. The resistors you choose must match the high performance you project for the entire system.

SAGE PRECISION WIREWOUND POWER RESISTORS give you:

Close Tolerance .... to .05%
T.C. .... ± 20 ppm/°C.
Insulation .... 1,000 V-rms Minimum
Stability .... .5% Maximum Lifetime Drift

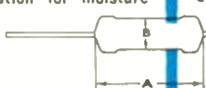
Beyond these characteristics, however, SAGE skills and in-process testing methods developed through more than a decade of specialization assure you of maximum protection against catastrophic failure and give Sage Resistors the *Priceless Plus of Reliability* . . . SAGE offers a variety of power ratings, from 1 Watt to 50, in three basic types, as detailed below, with resistances to 200,000 ohms. We shall be happy to send detailed specifications.

Check your requirements against these SAGE "Silicohm" Resistors

In 13 sizes from 1 to 10 Watt Ratings

TYPE "S," SILICONE COATED POWER RESISTORS

Compact, light weight, single layer "Silicohm" type "S" units are made to MIL-R-26C Characteristic G. Exclusive high temperature insulation for moisture environment . . . minimum dielectric strength of 1000 V-rms . . . precision .05% . . . T.C. of 20 ppm/°C.



MODEL	ratings, WATTS	dimensions, A*	B*	resistances Max. Ohms
SA1W	1	.500	.125	12,000
SB1W	1	.687	.125	20,000
SA2W	2	.500	.187	15,000
SB2W	2	.812	.187	25,000
S2W	2	.625	.250	20,000
S3W	3	.750	.250	30,000
SS5W	5	.875	.312	50,000
SR5W	5	1.000	.312	60,000
SL5W	5	1.125	.312	70,000
SS7W	7	1.250	.312	80,000
SR7W	8	1.375	.375	125,000
SS10W	10	1.812	.375	175,000
S10W	10	1.937	.375	200,000

*nominal inches

In 3, 5, 7, 8 and 10 WATT ratings

TYPE "CS" CLIPPER METAL SHEATHED RESISTORS

The new standard in clip mounting, assembly protected units, Type "CS" Resistors feature metal sleeves shrunk by a Sage developed process to conform exactly to fully insulated type "S" resistors. Smallest in size and coolest in operation of comparable units. **Wattage Ratings double when CS is clip mounted on full heat sink.** Tolerance from .05 to 5% available.

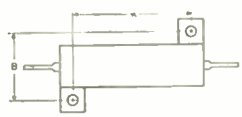
MODEL	ratings, WATTS	dimensions, A*	B*	resistances Max. Ohms
CS3W	3	.750	.250	30,000
CSR5W	5	1.000	.312	60,000
CSS7W	7	1.250	.312	80,000
CSR7W	8	1.375	.375	125,000
CS10W	10	1.937	.375	200,000

*nominal inches

Write now for free samples and complete specifications.

In M10W, M25W, and M50W Sizes

TYPE "M" METAL CLAD CHASSIS MOUNTED RESISTORS



Aluminum housed resistors for direct plate mounting, Type "M" units feature considerably less heat rise than similar units. Lower **Hot Spot** means long service life, near perfect stability (Average resistance shift is only 0.4% after 1000 cycled hours at recommended loads), and exceptional reliability under extreme conditions. MIL-R18546 Characteristic G models in 15 and 30 Watt Ratings. Dielectric strengths to 2500 V-rms.

MODEL	ratings, WATTS	resistances MAX. OHMS
M10W	10	25,000
M25W	20	50,000
M50W	40	175,000

Nominal Mounting Dimensions, inches		
MODEL	A	B
M10W	.562	.625
M25W	.719	.781
M50W	1.562	.844

SAGE ELECTRONICS CORPORATION
COUNTRY CLUB ROAD • EAST ROCHESTER, N. Y.

Section Meetings

(Continued from page 54A)

NEW YORK

"Biological Effects of Microwave Energy." Panel discussion by Col. G. M. Knauf, Dr. J. Vogelman, Dr. T. Ely, USAF Dynamics Electronics, Atomic Energy Commission. 10/7/59.

NORTH CAROLINA

IRE Space Age Electronics Symposium. 11/6–7/59.

NORTHERN NEW JERSEY

"Scope and Objectives of the Signal Corps Micro-Miniature Module Program," R. Gerhold, USASRDL Research & Development Labs. "Present Progress Status and Future Plans for Micro-Miniature Module Program," D. Mackey, RCA Semi-Conductor and Materials Division. 10/14/59.

NORTHWEST FLORIDA

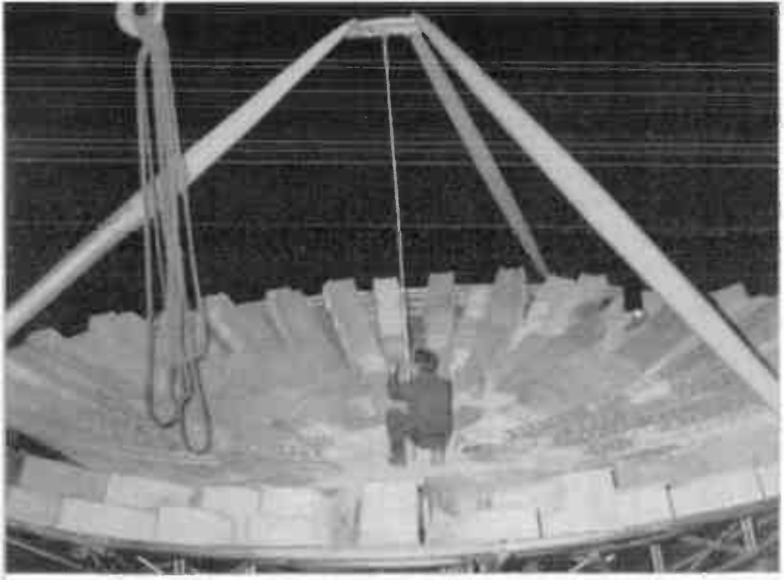
"Environmental Testing as Applied to Military Electronics," Dr. F. R. Britton, Electronic Communications, Inc. 10/20/59.

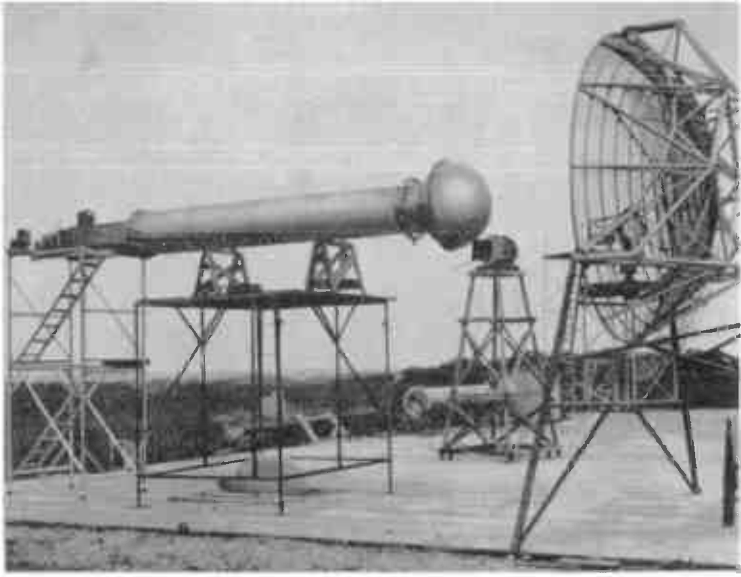
"Range Instrumentation Using Radioactive Sensing Technique," Dr. M. J. Cohen, Frankin Systems, Inc. 11/2/59.

OKLAHOMA CITY

"Activities of the Radio Propagation Division of the NBS Labs," Dr. F. W. Brown, National Bureau of Standards. 10/13/59.

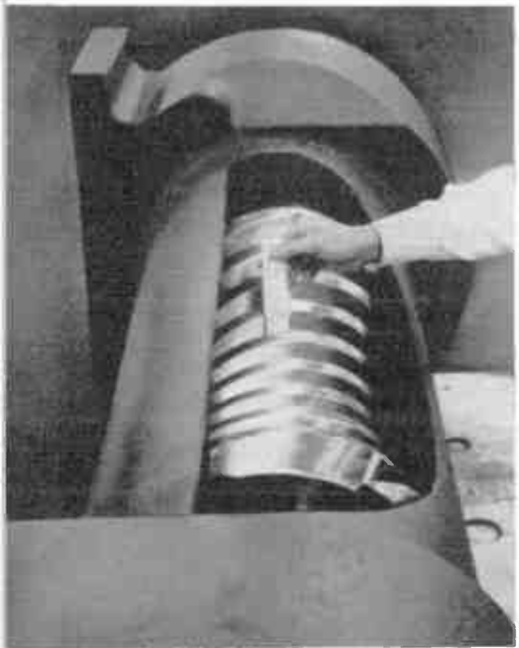
(Continued on page 60A)

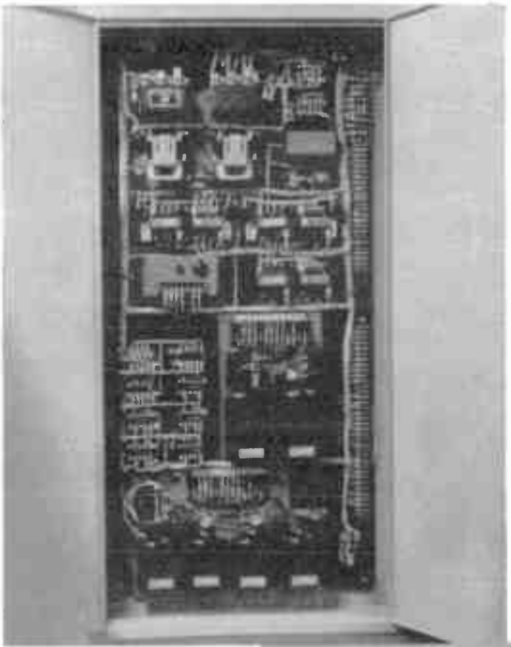




In antenna systems

KENNEDY capability is total capability





TOTAL *capability in the field of antenna systems?*

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Tracking Antennas-Radio Telescopes-Radar Antennas

"Trans-Horizon" Antennas-Tropospheric Scatter

Ionospheric Scatter

Section Meetings

(Continued from page 58A)

OMAHA-LINCOLN

"Electronics, Education, and the IRE," Dr. Ernst Weber, IRE President, 9/23/59.

"FM Broadcast Multiplex Techniques and Possibilities," B. L. Dunbar, KQAL-FM 10/30/59.

ORLANDO

"Broad-Band Crossed-Slot Telemetry Antenna," R. C. Payne, Dynatronics, Inc. 10/21/59.

"Inertial Guidance—Principles, Applications & Problems," W. E. Shepard, Minneapolis-Honeywell Regulator Co. 11/18/59.

OTTAWA

"Fuel Cells," E. L. R. Webb, National Research Council. 10/15/59.

"Ionospheric Research With Rockets and Satellites During IGY," J. E. Jackson, Goddard Space Flight Center, NASA. 11/19/59.

PHILADELPHIA

"The Cold War in Computers," Dr. M. Rubinoff, University of Pennsylvania; and Dr. S. Alexander, National Bureau of Standards. 10/14/59.

PHOENIX

"Computers in Industry," R. E. Clark, General Electric Co. 10/20/59.

PORTLAND

"Interesting Aspects of High Speed Oscilloscope Projects," J. Kobbe and C. Moulton, Tektronix, Inc. 10/13/59.

Field trip to USAF Radar Site & Telco Microwave Repeater Installations. 10/17/59.

"A Unique Forced Complement Audio Crossover System," C. Moulton, Tektronix, Inc. 10/22/59.

PRINCETON

"Infrared and Optical Masers," A. L. Schawlow, Bell Telephone Labs. 10/8/59.

QUEBEC

Visit to CFCM-TV Studios, Transmitter and Antenna. 10/27/59.

REGINA

"Electronics in Solid, Space, and Sound," Dr. C. N. Hoyler, RCA. 10/20/59.

RIO DE JANEIRO

"National Manufacturing of Electronic Equipment," J. L. Palhares, Standard Electrica S.A. 10/14/59.

"Some Modern Aspects of the Electronics," Dr. O. A. Cardim, Corcao Cardim S.A. 11/11/59.

ROME-UTICA

"The Role of Electronics in the Space Age," Dr. S. Ramo, Ramo-Wooldridge Corp. Communications Symposium. 10/5-6-7/59.

SACRAMENTO

"Single Sideband Communications Systems," G. R. Fuller, McClellan AFB. 9/24/59.

"SAGE System on the West Coast," George Story, McClellan AFB. 10/10/59.

"Filtering and Equalization in Optical Systems," Dr. B. M. Oliver, Hewlett Packard Co. 11/10/59.

ST. LOUIS

"The Ion Rocket For Space Propulsion," A. E. Lennert, Martin Co. 11/12/59.

(Continued on page 64A)

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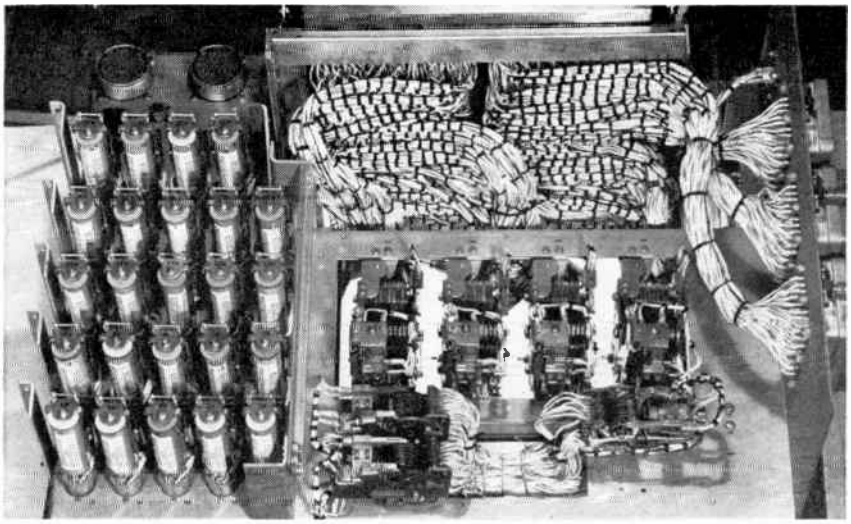

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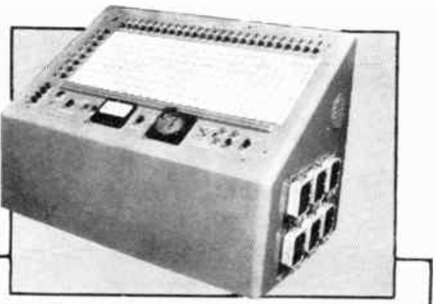

SPECIFICATIONS:

1. Continuity Test
   - A. Test Voltage .... 28 V.D.C.
   - B. Continuity Current .... 1 ampere
   - C. Continuity Resistance .... adjustable from 0.3 ohms to 10 ohms
2. Continuity-Discontinuity Test
   - A. Test Voltage .... 28 V.D.C.
   - B. Continuity Current .... 1 ampere
   - C. Continuity Resistance 0.3 ohms to 10 ohms
   - D. Discontinuity Resistance .... 2.5 megohms reject, 3 megohms accept
3. Short Test
   - A. Test Voltage .... 28 V.D.C.
   - B. Test Current .... 0.03 ma (max)
   - C. Short Resistance Range .... 2.5 megohms reject, 3 megohms accept
4. Ohmmeter
   - A. Range .... 0 to 200 megohms
   - B. Accuracy .... ±3%
5. Timer (Standard)
   - A. 60 minute range, 0.2 second scale division
   - B. Accuracy 0.1 sec. per operation at 60 cycles

   Timer (Optional)
   - A. 60 minute range, 0.01 second scale division
   - B. Accuracy 0.002% per operation ±1 division
6. Power Requirements
   - A. 100 to 125 V.A.C. 55 to 65 cycles (standard timer)
   - B. 100 to 125 V.A.C. 50 to 400 cycles (optional timer)
7. External Energization
   - A. 28 V.D.C. and 110 V.A.C., 60 cycles are provided for external energization of relays or other resistive devices, isolated from test voltage.
   - B. Other voltages may be supplied by external power supplies and switched as external energization or other test purposes.

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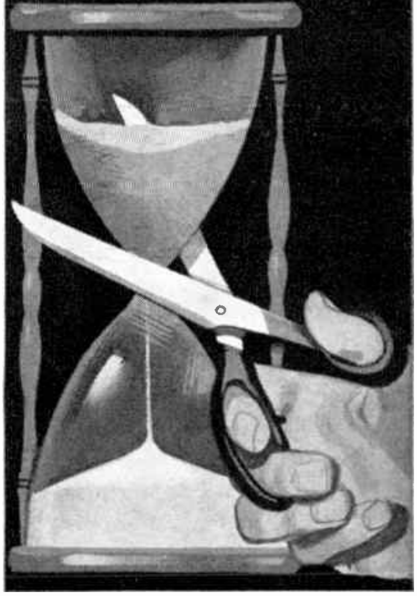

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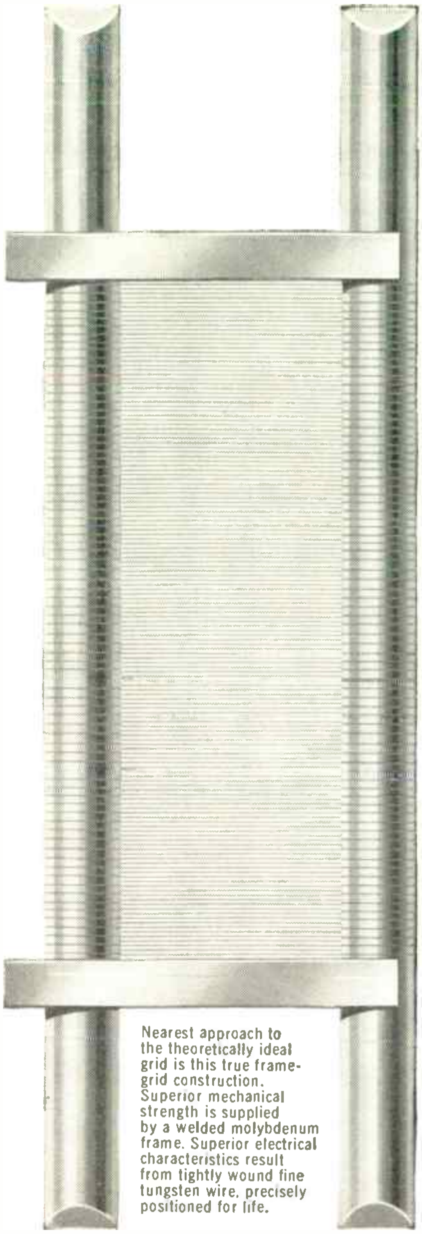

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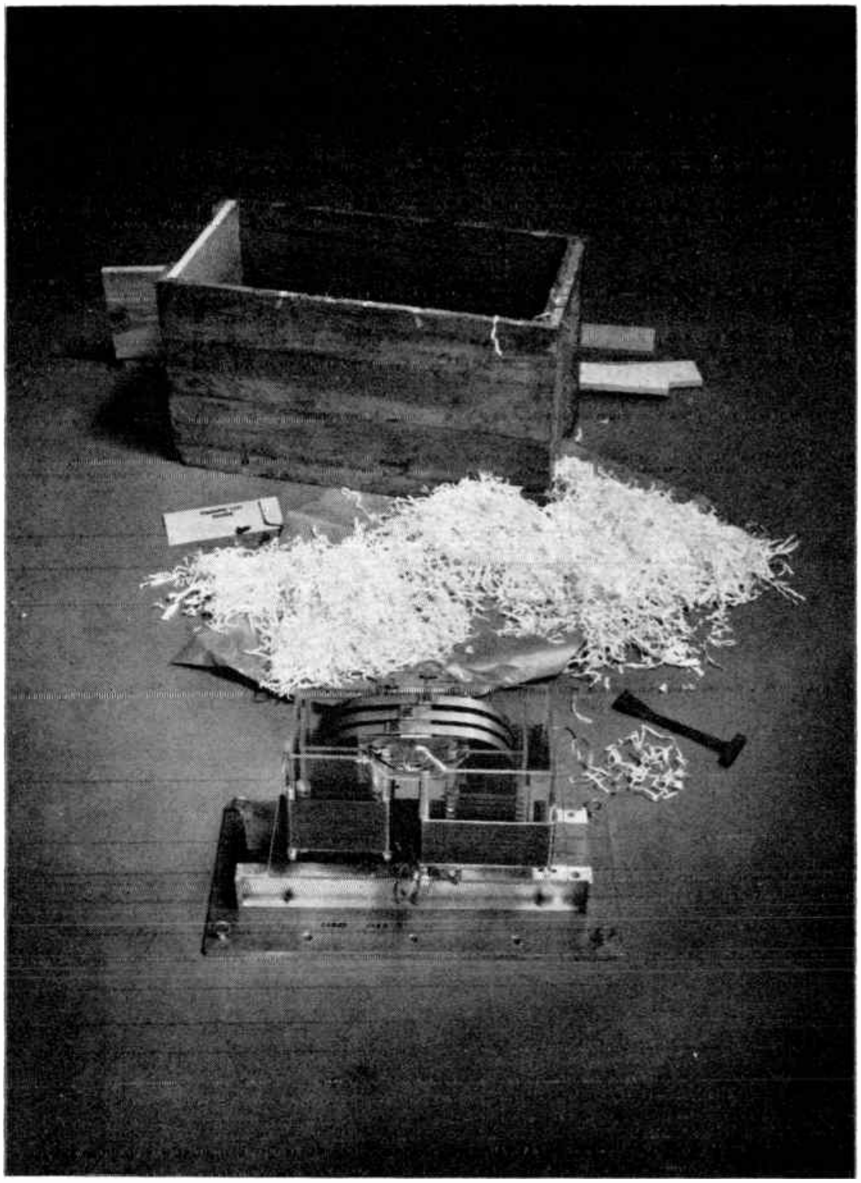

Section Meetings

(Continued from page 60A)

SAN ANTONIO

"Parametric Amplifiers," Dr. K. Katzebue, Texas Instruments. 10/14/59.

SAN DIEGO

"The Use of Electronics in Neurophysiology," Dr. Paul Shea, San Diego Mercy Hospital. 10/7/59.

SCHENECTADY

"A System for Electronic Control of Highway Vehicles," L. E. Flory, RCA. 10/27/59.

"Digital Computer Simulation in Speech Research," M. V. Mathews, Bell Telephone Labs. 11/10/59.

SOUTH CAROLINA

"The DEW Line," Admiral Sowell, Federal Electric. 10/23/59.

SYRACUSE

Pioneer Night. Panel discussion with panel of the following: Dr. A. N. Goldsmith, R. A. Heising, H. Pratt and W. C. White. Presentation of David Sarnoff 1959, Award to Dr. W. R. G. Baker.

TOKYO

"Linear Accelerator and Klystron Tube," Prof. M. Chodorow, Stanford Microwave Lab. 9/9/59.

"Comments on the Institute of Radio Engineers," Prof. A. A. Oliner, Polytechnic Institute of Brooklyn. 10/21/59.

"Stereo Broadcasting Systems Discussing Recent Proposals," Dr. D. H. Ewing, RCA. 10/26/59.

TORONTO

"Wideband Scatter Equipment," M. O. Felix, Canadian Westinghouse. 11/2/59.

IRE Canadian Convention. 10/7-8-9/59.

TUCSON

"University of Arizona Nuclear Reactor," R. L. Chapman, University of Arizona. 10/23/59.

TULSA

Tour of American Airlines Facilities. Commentator—W. Welder. 10/22/59.

"Characteristics and Applications of Solion Electrochemical Devices," C. D. Anderson, Texas Research Associates. 11/19/59.

WASHINGTON

"Space Surveillance," R. L. Easton, U. S. Naval Research Lab. 10/5/59.

"Men, Money and Management," Rear Admiral R. Bennett, U. S. Dept. of Navy. 11/2/59.

WESTERN MICHIGAN

Annual Business Meeting. 6/10/59.

"Miniature Airborne Military Electronic Devices," C. Van Namen, Lear, Inc. 10/8/59.

WINNIPEG

Annual Stag Meeting. 10/2/59.

Field trip to Grand Forks, N. D. to Visit SAGE Headquarters. Film on NORAD shown. 10/17/59.

(Continued on page 66A)

Section Meetings

(Continued from page 64A)

SUBSECTIONS

Buenaventura

"Some Military Aspects of the Space Age," J. Kschurgi, Douglas Aircraft, 10/21/59.

"The Potential Uses and Abuses of Gigantic Machine Memories," Mr. Benson, Benson-Lehner Corp. 11/13/59.

Burlington

"Deposited Carbon Resistors," N. B. Healey, International Resistance Co.; "Metal Film Resistors," R. W. Hamilton, International Resistance Co.; "Resistors of the Future," H. A. Clark, International Resistance Co. Tour of International Resistance Company's facilities. 10/30/59.

Eastern North Carolina

Tour of Microwave Carrier Equipment Manufacturing Plant. Commentator—Henry Horne, III, Kellogg Co. 10/9/59.

Memphis

Field Trip Inspection of the IBM 705 Computer. 7/30/59.

"A Preview of Plans for an Expedition to Study the Forthcoming Solar Eclipse," Prof. J. Freimuth, Southwestern at Memphis. 8/28/59.

Merrimack Valley

"Use of Educational Aids in Electronics," Dr. H. E. Stockman, Merrimack College. 10/19/59.

New Hampshire

"Ceramic Tube Developments," J. LeVan, Sanders Associates, Inc.; "A Two-Port Cross-Field Micro-Wave Device," J. Kline, Sanders Associates, Inc. 10/22/59.

Reading

"Recent Sophistications in Communications," E. M. Noll, Writer. 9/16/59.

"The Brains in Telephone Switching Systems," L. A. Moretz, Bell Telephone Company. 10/21/59.

Santa Ana

"Traffic Control Through Electronic Devices," Dr. D. Gerlough, Thompson-Ramo-Wooldridge Corp. 10/27/59.

Santa Barbara

"Explorer VI and Other Space Satellites," Dr. G. Mueller, Space Technology Labs. 10/20/59.

Southwest Ontario

"History and Use of Stereo-Sound," J. Allen, K.L.A. Industries. 10/20/59.

Westchester

"Feedback, Life and Reliability of Feedback Amplifiers," Dr. A. Uiga, Ballantine; "Sampling Oscilloscope," K. Magleby, DuMont. 10/27/59.

Western North Carolina

Tour of WSOC-TV Studio Facilities. Comments by L. L. Caudle. 10/20/59.

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The standard SMART machine enables you to measure up to 16 different d-c parameters of a transistor or other semiconductor device and record these data within 12 seconds. A minimum time of .5-second is required to test each parameter and an additional .2-second records the intelligence on an IBM 526 Summary Punch or other digital recording device. Using all 16 parameters, of course, 300 transistors may be tested per hour; however, fewer parameters would be desired on most testing runs and upwards of 500 semiconductors/hour could be handled easily.

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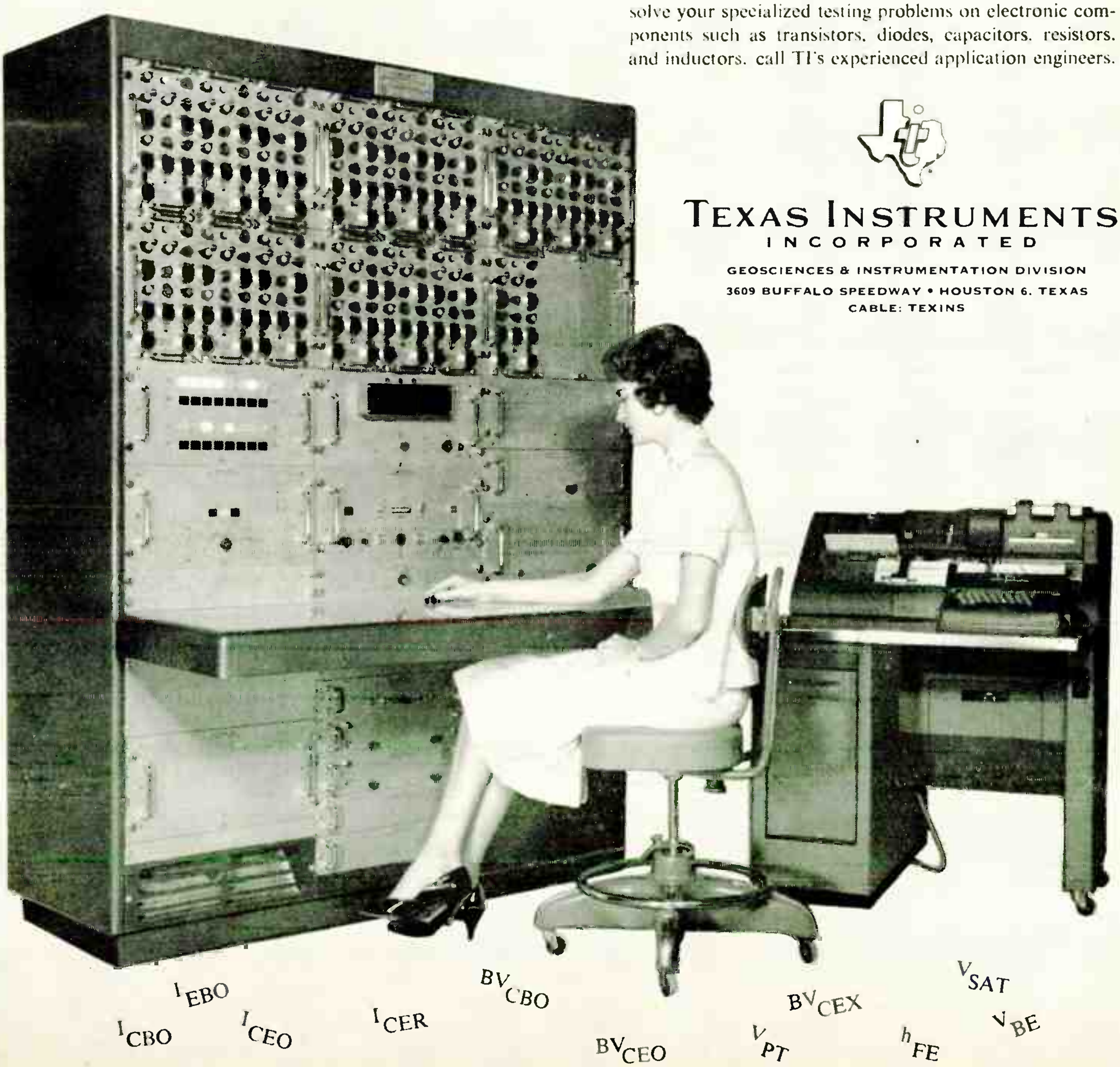

IRE People

Wellesley Dodds (SM'45), manager of Varian Associates' traveling-wave tube and backward-wave oscillator development department, has been made vice president of engineering and a member of the board of directors for Bomac Laboratories, Inc., a Varian subsidiary located at Beverly, Mass. As Bomac's vice president of engineering, he will head all research and development on present and future product lines, including radar switching tubes, magnetrons, silicone diodes and duplexers.

W. J. Dodds

He is also chairman of the working group on microwave tubes of the advisory group on electron tubes for the U. S. Department of Defense.

Before joining Varian in 1956, he was with Radio Corporation of America for 13 years, serving as research engineer and manager of RCA's tube division's microwave tube advanced development program. After receiving the M.S. degree in physics at the University of Kansas, and before joining RCA, he was in university teaching and research.

Mr. Dodds holds several patents in the field of microwave tubes and circuits and is a member of Sigma Xi, Sigma Pi Sigma, and Pi Mu Epsilon.

❖

Dr. Paul M. Erlandson (SM'50), has been named Director of Research of the Schlumberger Well Surveying Corporation and will head the company's theoretical and experimental programs in physics, mathematics, and related engineering sciences at its Research Center in Ridgefield, Conn. He was formerly Director of Physics Research for Continental Can Co. in Chicago, Ill. Prior to that he was Assistant Vice President and Chairman of the Physics Department of the Southwest Research Institute, San Antonio, Texas.

P. M. Erlandson

A graduate of the Massachusetts Institute of Technology and of the University of Texas, where he received the Ph.D. degree in Physics, Dr. Erlandson is a Senior Member of the Instrument Society of America, and a member of the Optical Society of America, the Acoustical Society of America, and the American Physical Society.

❖

The appointment of **Dr. William G. Hoover** (A'44–M'55), professor of electrical engineering at Stanford University, as technical director of Granger Associates has been announced by Dr. John V. N. Granger, president.

Dr. Hoover has been director of Stanford's Ryan High Voltage Laboratory since 1957. He will retain an association with the University's academic and research activities.

With Granger Associates, he will supervise the company's expanding technical programs with particular emphasis on pulse systems, microwave tube applications and advanced circuit techniques, Dr. Granger said.

A native of Pueblo, Colo., he spent his youth in San Mateo and Burlingame on the San Francisco Peninsula. He received the B.A., Electrical Engineering and Ph.D. in E.E. degrees from Stanford, the latter awarded in 1936.

He was employed by Westinghouse Electric and Manufacturing Co. from 1929 to 1931, returning to Stanford in 1932 for advanced studies and to teach.

The new Granger technical director has had extensive experience in lightning protection and corona breakdown, radar modulator and pulse transformer design and in pulse circuitry.

In recent years he has been a consultant to Granger, Pacific Switchgear Division of Federal Pacific Electric Co., Stanford Research Institute, Ampex Corp., General Electric Co., Allis-Chalmers Manufacturing Co., Grumman Aircraft Engineering Corp., Dalmo-Victor Co. and Dale Products, Inc., among others.

A registered professional engineer in the State of California, he is a member of the American Institute of Electrical Engineers, Sigma Xi, Tau Beta Pi, the American Society for Engineering Education and Phi Beta Kappa.

❖

Promotion of **Dr. Joseph F. Hull** (M'50), to Director of the Research Laboratory of the Litton Industries Electron Tube Division has been announced by Dr. Norman Moore, Division General Manager.

Dr. Hull has been Chief Scientist of the San Carlos, Calif., laboratory since April, 1955, working primarily in the fields of backward-wave oscillators and crossed-field devices. He received the B.S.E.E. degree from the University of Wisconsin, the M.S.E.E. degree from Rutgers University and the D.S.E.E. degree from the Polytechnic Institute, Brooklyn. Prior to joining Litton Industries, he was employed with the General Electric Research Laboratory and the Signal Corps Engineering Labs.

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(*Continued on page 70A*)

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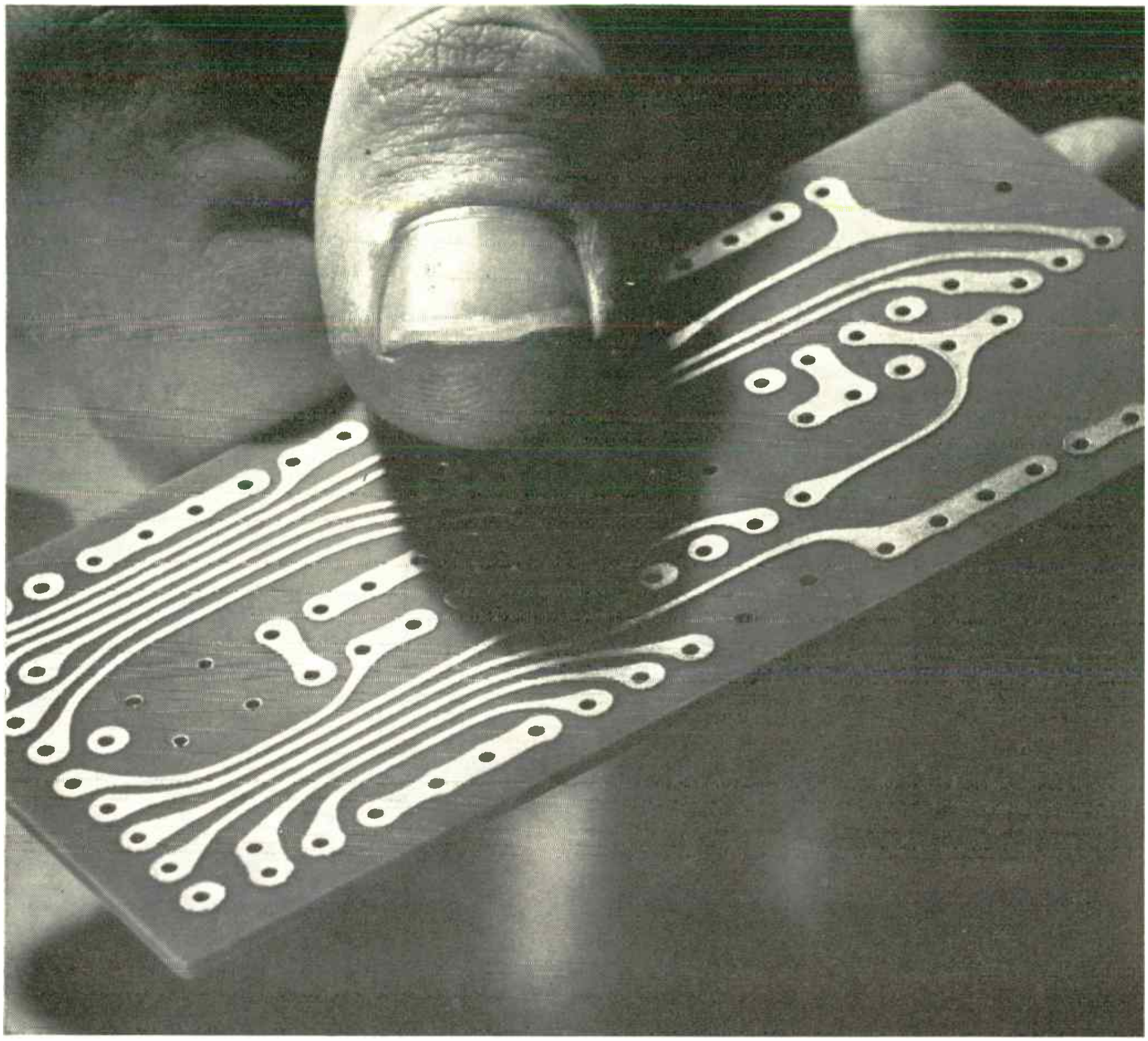
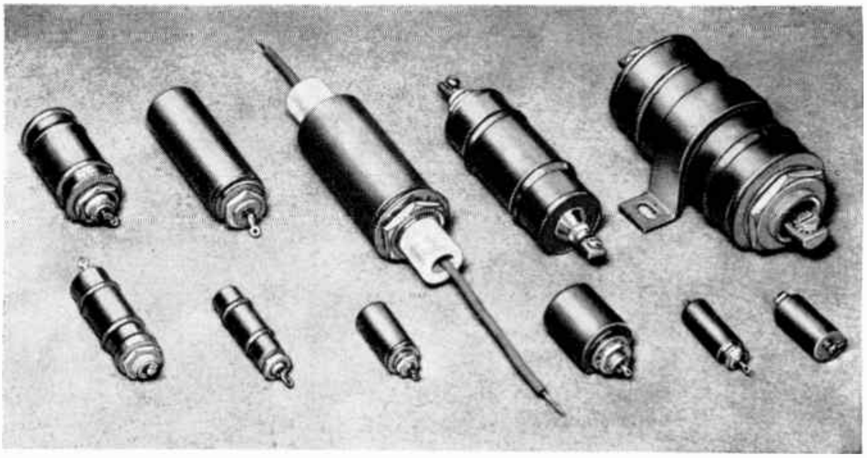


FOTOCERAM circuit board blanks are made using the micro-accuracy of photography. All holes and shapes are produced by simple exposure to light, heat, and an etching operation.



New Series of Sprague Cylindrical-Style Radio Interference Filters: top row, l. to r.—4JX14, 5JX94, 1JX115, 20JX15, 50JX20 bottom row—5JX27, 1JX54, 1JX113, 1JX117, 2JX49, 1JX118.

New Series of Small, Light Radio Interference Filters

The new cylindrical-style radio interference filters recently announced by Sprague Electric Company are the smallest and lightest filters of their type available for military and industrial electronic and electrical equipment. Their basic design was pioneered by Sprague in order to achieve maximum miniaturization.

This new series of standard filters, believed to be the most complete in the industry, ranges in current rating from 5 milliamperes to 50 amperes covering the majority of applications.

The natural shape of the rolled capacitor section and of the toroidal inductors dictates the cylindrical form. All filters have threaded-neck mountings for use on panels or bulkheads. This assures both the proper isolation between input and output terminals as well as a firm peripheral mounting with minimum impedance to ground.

Listed in Sprague Engineering Bulletin 8100 (available upon request to the Technical Literature Department) are 68 of the more popular low-pass filter designs intended for use as three-terminal networks connected in series with the circuits to be filtered. The excellent interference attenuation characteristics reflect the use of Thrupass® capacitor sections.

Since maximum effectiveness of filtering involves elimination of mutual coupling between input or noise source and output terminals, filters should be mounted where the leads being filtered pass through a shielded chassis or bulkhead. The threaded neck mounting is designed to give a firm metallic contact with the mounting surface over a closed path encircling the filtered line and to eliminate unwanted contact resistance so that the theoretical effectiveness of these units is realized in practice.

Typical insertion loss is determined by measurements made in conformance with Military Standard MIL-STD-220. Minimum curves for specific filters are available upon request.

For assistance in solving unusual interference, rating, or space problems, contact Interference Control Field Service Manager, Sprague Electric Co., at 12870 Panama Street, Los Angeles 66, California; 224 Leo Street, Dayton 4, Ohio; or 235 Marshall Street, North Adams, Massachusetts.

IRE People

(Continued from page 70A)

J. T. Lindsay Brown (A'37–VA'39–M'55) has recently joined Gordos Corp., where he will head the exploratory development of mercury-wetted relays and dry-reed switches.

J. T. L. Brown

He recently retired after 44 years of research and development with Bell Telephone Laboratories and, previously, with Western Electric Co. He holds 20 patents, and was responsible for the origination and development of Bell System mercury-wetted relays. In the large number of communication and defense applications where these are now used, the new design principles which they employ, are recognized as an outstanding contribution to switching.

A graduate of College of the City of New York, Mr. Brown is a member of the A.I.E.E. and the Acoustical Society of America.

❖

(Continued on page 78A)

Problem: HYDROSPACE

Hydrospace — concept of space in ocean, motion through an opaque medium — offers a complex of challenge unique in engineering. To solve this tri-dimensional riddle, Raytheon's Submarine Signal offers an equally unique concept — industry's first ASW CENTER.

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For information on select staff openings in several technical specialties, write: Mr. Donald H. Sweet, Executive & Engineering Placement, Raytheon Company, 624 G Worcester Road, Framingham, Massachusetts (suburban Boston).

SUBMARINE SIGNAL

GOVERNMENT EQUIPMENT DIVISION

RAYTHEON

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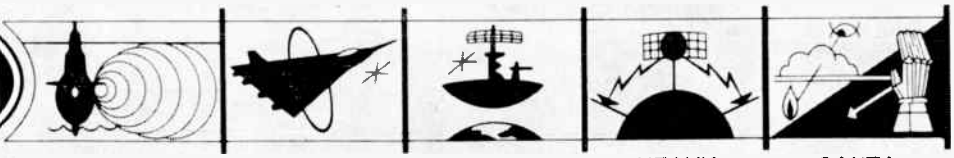
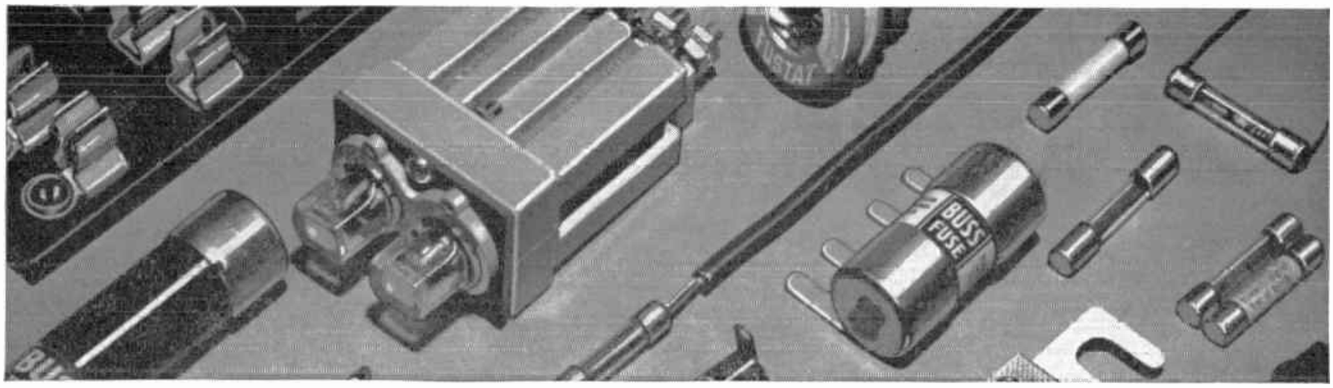


SUBMARINE SIGNAL | AIRBORNE ELECTRONIC | SYSTEMS MANAGEMENT | HEAVY ELECTRONIC | SANTA BARBARA



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Because there's a BUSS or FUSETRON fuse to fit Your Every Electrical Protection Need.

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The BUSS and FUSETRON fuse lines includes:

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Indicating fuses where signal must be given when fuses open, or to activate an alarm.

BUSS and FUSETRON fuses range in size from 1/500 amperes up — and there's a companion line of fuse clips, blocks and holders. Whatever your fuse requirements, the chances are more than good there's a BUSS or FUSETRON fuse to satisfy them.

If you have a special protection problem . . . extensive BUSS laboratory facilities and a large engineering staff are at your disposal to help you save money and engineering time.

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BUSS fuses are made to protect - not to blow, needlessly.

BUSS makes a complete line of fuses for home, farm, commercial, electronic, electrical, automotive and industrial use.

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McGRAW EDISON

IRE People

(Continued from page 74A)

Bernard Hecht (M'45–SM'54), Quality Control Consultant, has been appointed by the U. S. International Cooperation Administration for an important temporary assignment in Israel. He will work with the Israeli Institute of Productivity in the establishment of modern scientific quality control organizations and methods in the young and fast growing industrial firms of that country. He will work in conjunction with the U. S. Overseas Mission in Tel Aviv, Israel.

B. HECHT

Mr. Hecht was recently elected to the grade of Fellow of the American Society for Quality Control. He expects to return to the United States after the first of the year and will resume his regular activities in the firm of Bernard Hecht and Associates, Los Angeles, Calif.

❖

William J. Hildebrandt (S'49–A'50–M'56), has been promoted to the position of Development Engineer with managerial responsibility for Applied Research, Component Development, and the Electro-mechanical Laboratory of Underwood Corp.

W. J. HILDEBRANDT

He received the B.S.M.E degree from Stevens Institute of Technology and the M.S.E.E. degree from University of Connecticut. Over the past eight years he has specialized in electro-mechanical engineering and digital computer systems. He is a Lecturer in Electrical Engineering at the University of Connecticut Graduate School and holds membership in Tau Beta Pi and Phi Kappa Phi.

❖

The appointment of **Dr. R. J. Filipowsky** (A'50), as the Head of the newly-formed Advance Development Department has been announced by Collins Radio Company, Western Diviison in Burbank, Calif.

In his new position he will be responsible for the direction of research in the field of digital data communication.

He received the M.S. degree and the D.Sc. degree from the Technical University of Austria. Prior to joining Collins, he was Professor and Head of the Department of Electronics at Madras Institute of Technology in India, and later became Ad-

(Continued on page 82A)

IRE People

(Continued from page 78A)

visory Engineer for Communications at the Electronics Division, Westinghouse Electronic Corp., Baltimore, Md.

He is a member of British and Australian IRE and is affiliated with many other technical societies.

❖

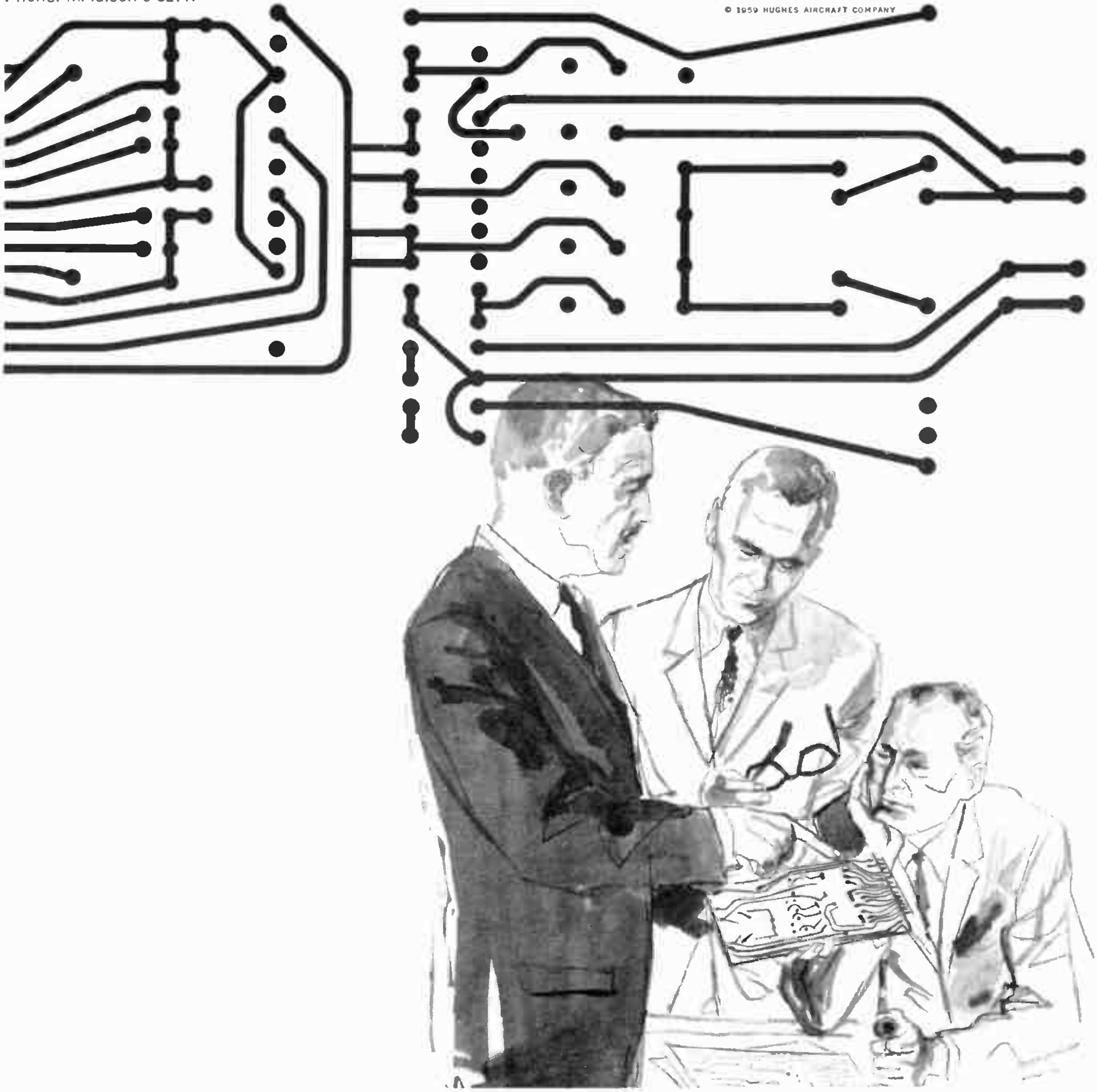
Stephen J. Jatras (M'52) has been appointed director of engineering of the Lockheed Electronics and Avionics Division (LEAD), Los Angeles, Calif.

S. J. Jatras

He joined Lockheed in 1956 as a staff scientist in the Missiles and Space Division, and served that division as business administrator for the research branch and manager of the flight controls and guidance division before becoming assistant to the director of research. He transferred to LEAD when the new electronics division was formed March 2, 1959, and has served as executive assistant to the general manager and director of marketing at LEAD.

Before joining Lockheed, he was vice

(Continued on page 84A)



(Continued from page 82A)

president and chief engineer of Midwestern Instruments, Inc., Tulsa, Okla.

Mr. Jatras received the B.S. degree in electrical engineering from Carnegie Institute of Technology and the M.S. degree in electrical engineering from Massachusetts Institute of Technology. He also has been a research engineer at M.I.T., instructor in electrical engineering at the University of Massachusetts, and telephone engineer with Stromberg-Carlson.

He is a member of Sigma Xi, and the American Institute of Electrical Engineers.

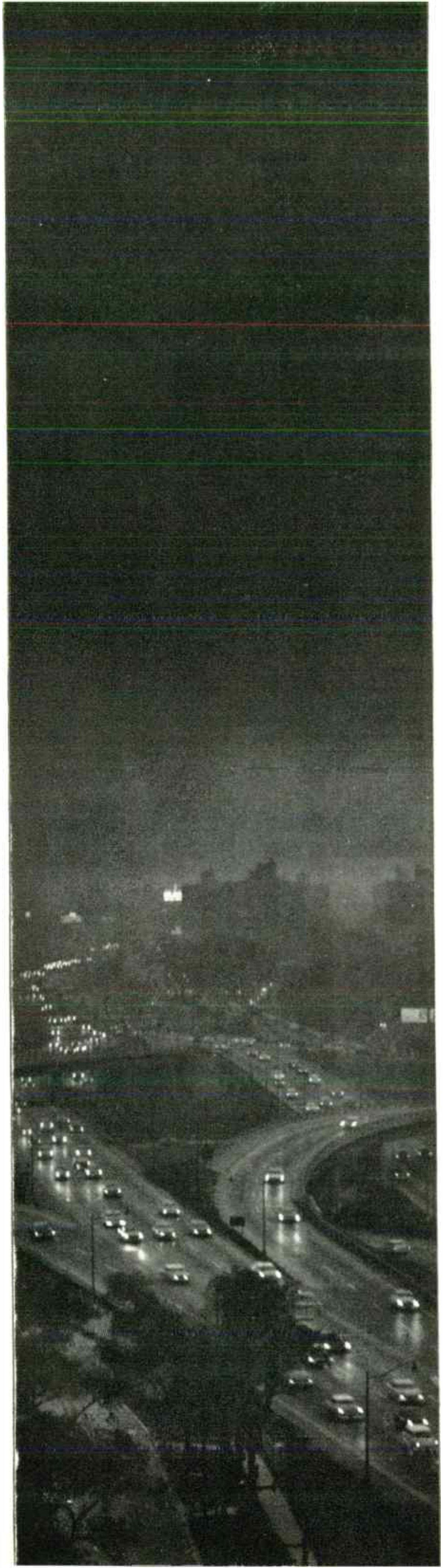
❖

Kermit B. Karns (A'49–M'53) has been named chief of the Navigation Aids Electronic Engineering Section in the Air Navigation facilities Division of the Federal Aviation Agency's regional headquarters, Kansas City, Mo.

K. B. Karns

He joined F.A.A.'s predecessor agency in 1937 as an airway keeper. In 1947 he was named an electronics maintenance technician in charge and in 1955 was rated an electronic engineer,

(Continued on page 86A)



One of a series

Resolving the driver-car-road complex

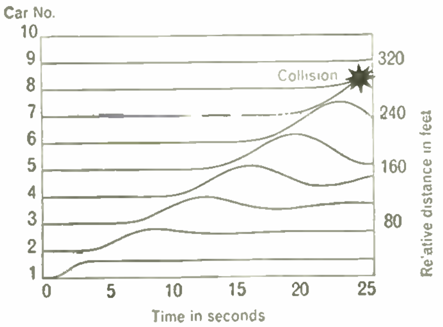
The manner in which vehicles follow each other on a highway is a current subject of theoretical investigation at the General Motors Research Laboratories. These studies in traffic dynamics, coupled with controlled experiments, are leading to new "follow-the-leader" models of vehicle interaction.

For example, conditions have been derived for the stability of a chain of moving vehicles when the velocity of the lead car suddenly changes — a type of perturbation that has caused multiple collisions on modern superhighways. Theoretical analysis shows that the motion of a chain of cars *can be stable* when a driver accelerates in proportion to the relative velocity between his car and the car ahead. The motion is always unstable when the acceleration is proportional only to the relative distance between cars. Experimentally, GM Research scientists found that a driver does react mainly to relative velocity rather than to relative distance, with a sensitivity of reaction that increases with decreasing distance.

Traffic dynamics research such as this is adding to our understanding of intricate traffic problems — what causes them, how they can best be resolved. The study is an example of the ways GM Research works to make transportation of the future more efficient and safe.

General Motors Research Laboratories

Warren, Michigan



Relative positions of 10 hypothetical cars after lead car goes through maneuver. Amplitude of instability increases, resulting in a collision between 7th and 8th cars.

IRE People

(Continued from page 81A)

based on his research and development work.

Mr. Karns is a member of the American Institute of Electrical Engineers.

❖

The promotion of **Fred L. Katzmann** (S'52–A'53–M'58) to manager of circuit design for the instrument engineering department of Allen B. Du Mont Laboratories, Inc., has been announced by Robert W. Deichert, engineering director of the company's oscilloscopes and other scientific instruments.

F. L. KATZMANN

In his new post, Mr. Katzmann will be responsible for circuit conception, design, and development for the complete line of Du Mont's scientific instruments.

Associated with Du Mont since 1952, he has held high engineering posts in the company's instrument operations. Prior to that he served with the U. S. Army as an electronics engineer, after serving two years at Ft. Huachuca, Ariz. with the Signal Corps.

Mr. Katzmann was graduated from the College of the City of New York with the B.E.E. degree. He also attended graduate schools of CCNY and at the University of Arizona. He is presently taking specialized post-graduate work at Stevens Institute.

❖

Howard Katzman (M'58) has joined Granger Associates, Palo Alto, Calif., to become project manager in the firm's countermeasures and transmitter development program.

From 1952 through 1955 he was with Stavid Engineering Co., Plainfield, N. J., where he worked on radar system parameters and the design of radar transmitters and waveguide components. From 1955 until joining Granger, he was employed by Lockheed Missiles and Space Division. There he was involved in the design of equipments used for programming various functions in satellite vehicles. He was also a project engineer on a semi-automatic checkout equipment for telemeter systems.

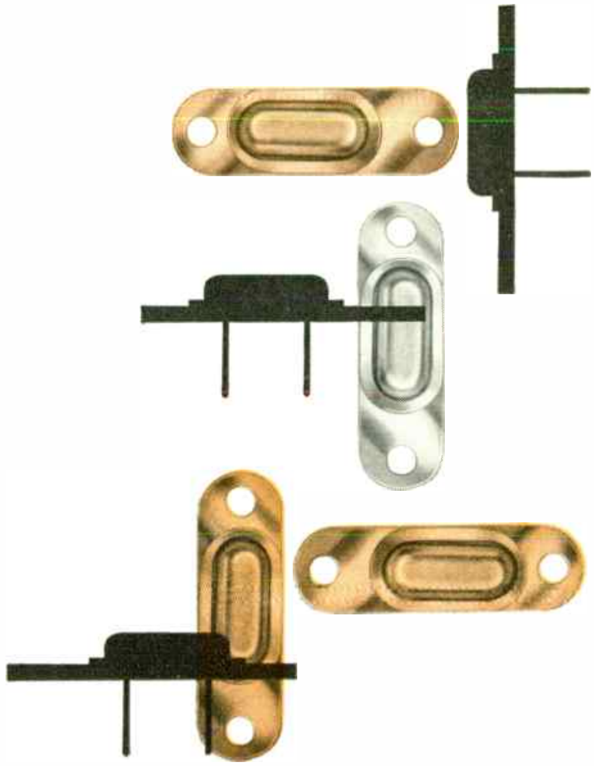
Mr. Katzman received the B.S. and M.S. degrees in electrical engineering from Newark College of Engineering.

Mr. and Mrs. Katzman and their three children reside at 648 Hudson Drive, Santa Clara, Calif.

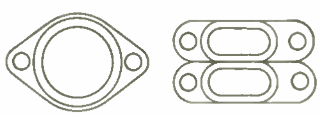
❖

Paul H. Kreager (A'38–M'45) has been named Field Operations Manager of the Instrument Division of American Electronics, Inc. His duties will include supervision of contracts administration and the

(Continued on page 88A)



CLEVITE'S NEW
SPACESAVER
TRANSISTOR



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THREE AMPERE SWITCHING TYPES

TEST	CTP 1728	CTP 1735	CTP 1729	CTP 1730	CTP 1731	CTP 1736	CTP 1737	CTP 1733
Min BVcbo @ 2 ma (volts)	40	60	80	100	40	60	80	100
Min BVceo @ 500 ma (volts)	25	40	55	65	25	40	55	65
Min BVces @ 300 ma (volts)	35	50	65	75	35	50	65	75
Max Icbo @ 90°C @ Max Vcb (ma)	10	10	10	10	10	10	10	10
Max Icbo @ 2 V (μa)	50	50	50	50	50	50	50	50
D. C. Current Gain @ 0.5A	30-75	30-75	30-75	30-75	60-150	60-150	60-150	60-150
Max Veb @ 3.0 A (volts)	1.5	1.5	1.5	1.5	1.5	1.5	1.5	1.5
Max Vce (sat) @ 3.0A, 300 ma (volts)	1.0	1.0	1.0	1.0	0.8	0.8	0.8	0.8
Min fae @ 3.0 A (kc)	20	20	20	20	15	15	15	15
Max Thermal Resistance (°c/w)	2.5	2.5	2.5	2.5	2.5	2.5	2.5	2.5

Compared with present power transistors of similar ratings, the new Clevite *Spacesaver* gives you important new advantages.

Better Switching — Its low base resistance gives lower input impedance for the same power gain and lower saturation resistance, resulting in lower "switched on" voltage drop. Its lower cut off current means better temperature stability in direct coupled circuits (such as regulated power supplies) and a higher "switched off" impedance.

Better Amplifying — Improved frequency response leads to higher audio fidelity, faster switching and improved performance in regulated power supply applications.

Better Mounting — The *Spacesaver's* simple rectangular configuration and low silhouette make it adaptable to a wide variety of mounting requirements where space is at a premium. In aircraft and missile applications, its low mass (half present type) improves shock and vibration resistance of lightweight assemblies.

A DIVISION OF
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Phone for data and prices.

CLEVITE TRANSISTOR

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(*Continued from page 86A*)

training and supervision of all manufacturers representatives.

Before joining the Instrument Division, his last position was as Western Division Contracts Manager for Kearfott Company, a position he held for three years. His other experience includes four years as Contracts Administrator, Bendix Aviation Corporation, Radio and York Divisions, on guided missile component programs for the Armed Services, as well as Sales Manager for Bendix Radio with responsibility for contracts administration and negotiation, sales promotion, customer and public relations.

Mr. Kreager attended Wittenberg College and technical schools in Dayton, Ohio. He is a member of the Aircraft Owners and Pilots Association.

❖

Paul Gottfried (SM'58) and **Leonard L. Schneider** (M'58), partners in the firm of Reliability Engineering Associates, have announced the move of their headquarters to a new and larger office at 3740 Dempster Street, Skokie, Ill. They were formerly directors of the extensive reliability testing and component part screening programs conducted by Inland Electronics Corp.

❖

Dr. Anthony D. Kurtz (M'57) will be responsible for research, development, and production activities in semiconductor materials and devices in his new post as General Manager and Executive Vice President of Kulite Semiconductor Products, Inc. For the past three years he had been Director of Research of the Semiconductor Division of Minneapolis Honeywell Regulator Co. In this position he was responsible for the advanced device research and development efforts for Honeywell.

A. D. Kurtz

Prior to joining Honeywell, he was associated with Clevite Transistor Products as Senior Engineer in charge of diffused device development. Before that, he spent three years as a staff member of Lincoln Laboratory at M.I.T., where he was engaged in research on crystal perfection in semiconductors and solid state diffusion.

Dr. Kurtz received the B.S., M.S., and Sc.D. degrees from the Massachusetts Institute of Technology in the fields of physics and physical metallurgy. He is a member of the American Physical Society, the American Society of Metals, and Sigma Xi. He has published numerous articles in the fields of solid state physics and semiconductors.

❖

(*Continued on page 92A*)

Only a microwave engineer who has extensive experience with Travelling Wave Tubes and Backward Wave Oscillators can fully appreciate this latest advancement in the power supply art.

Look at these *exclusive* features...

- *built-in* delay line sweep over the entire range from 150 to 3600 volts
- *built-in* Automatic Gain Control
- *built-in front panel* switching for grid or anode modulation
- *built-in* digital readout for delay line supply
- *built-in* dual output jacks for parallel tube operation or external metering

PLUS automatic sequential application of filament, grid and collector, delay line, and anode voltages... each with its own *front panel adjustments*.

Naturally, there is automatic safety overload protection in the anode, delay line, and collector current circuits. The best news (of course) is that the PRD 813 BWO/TWT Power Supply is available FROM STOCK.

For the full story on the PRD 813, contact your nearest PRD representative or write:

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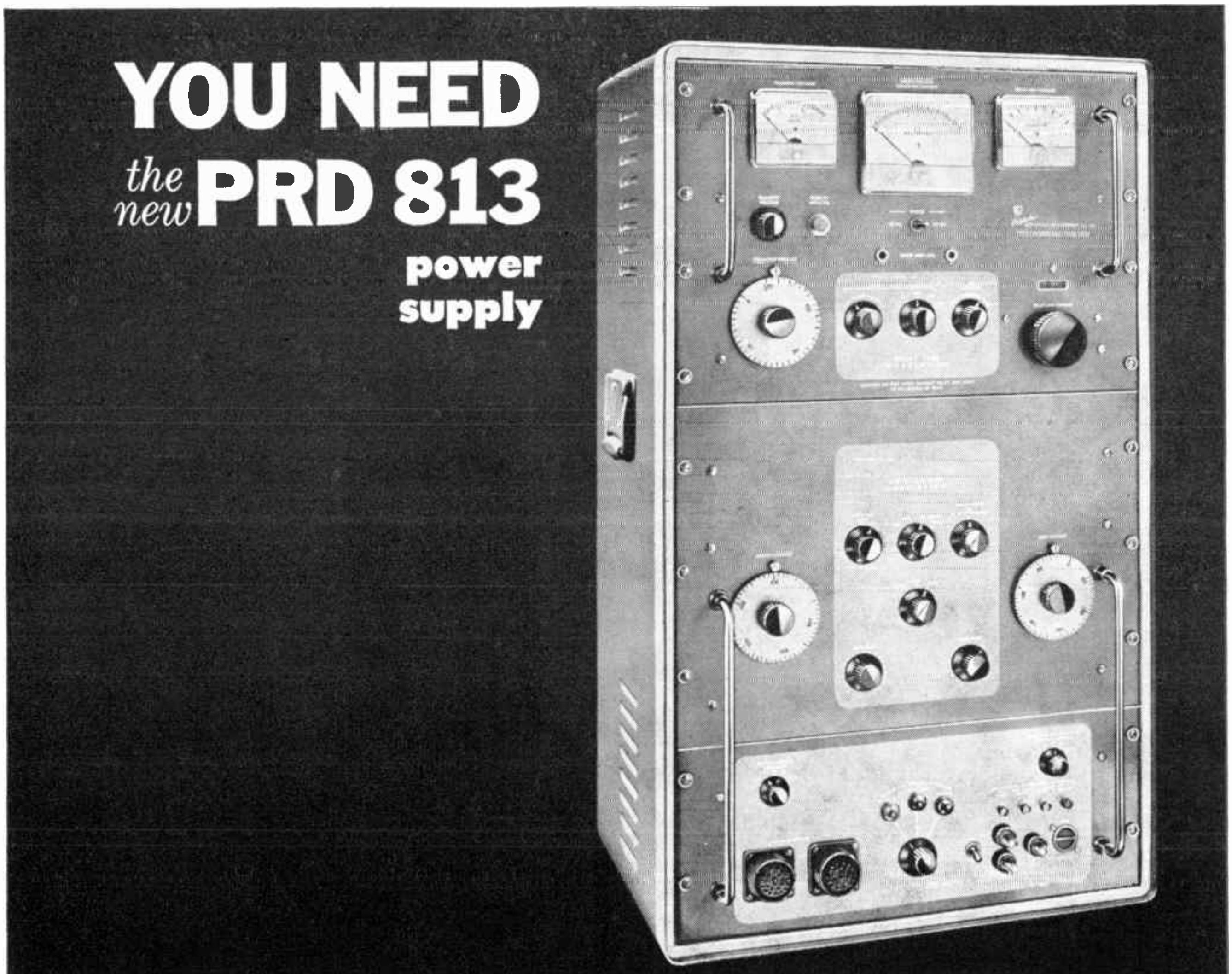

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HIGH-SIGNAL VOLTAGES

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540 CHANNELS

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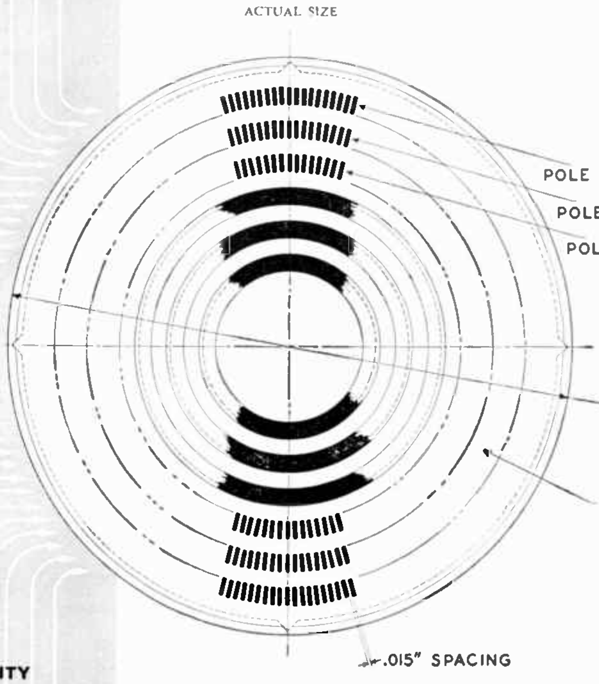
LIQUID FLOW

LOW-SIGNAL VOLTAGES

RADIATION

major advance in design:

SUPRAMICA® 555 ceramoplastic commutator plates



MYCALEX SWITCH
TM MODEL CP 499

combine: up to 540 channels... long life ...low noise... <u>in a 3 inch diameter</u>

The unique properties of SUPRAMICA 555 ceramoplastic make possible an equally unique standard of performance in commutator switching plates.

Total dimensional stability prevents contact loosening. Thermal endurance of the plate up to 700°F permits long, reliable operation. 1600 hour performance at 600 rpm *without* cleaning or adjustment is normal for MYCALEX commutators of this type. In tests, an additional 4,000 hours were easily obtained after simple brush cleaning. At 1,800 rpm, 200 hours of continuous operation are normal without cleaning or adjustment.

A new low in noise level—less than 1 millivolt, when switching 5 volts into a 150 ohm load—allows the sampling of transducers with peak output as low as 10 millivolts and noise level of 10-20 microvolts, without the use of pre-amplifiers.

And we are now able to include up to 540 rectangular contacts in one three-inch diameter plate.

A letter or telephone call will bring you complete data on MYCALEX commutator plates, switches and matched brush assemblies. Our engineers are ready to design and manufacture plates and complete switches to meet your individual specifications.

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IRE People

(Continued from page 88A)

The appointment of Paul A. Lang (M'58) as mechanical engineering manager of the West Coast Division of Military Electronic Operations for Allen B. Du Mont Laboratories, Inc., has been announced by E. F. Phillipi, Jr., West Coast Division manager.

P. A. LANG

He will assume the responsibility of heading the West Coast Division's mechanical engineering activities which include research and development as well as product design of mechanical and electro mechanical equipment.

He received his education in mechanical engineering at Rutgers University. After spending several years in New Jersey on engineering design and development projects, he moved to Los Angeles, Calif. in 1952, where he became associated with the West Coast Division of RCA. There he supervised a mechanical engineering group which developed several electro-mechanical products in the field of radar and display equipment.

Prior to joining the Du Mont staff, he spent twelve years in engineering research and development, having been personally involved in developing complex instrumentation for satellites, tracking antennas, radar equipment, display devices, precision electro-mechanical products and special purpose machinery.

His most recent position has been with the Space Technology Laboratories, Inc., of Los Angeles, where he headed the mechanical-engineering section which developed the various antenna feeds for the world's largest movable parabolic antenna system located at Jodrell Bank, England. On this assignment, he worked very closely with Professor A. C. B. Lovell of the University of Manchester, designer of the antenna. Also, while at Space Technology Laboratories, Mr. Lang developed TV cameras, tracking antennas, a considerable amount of instrumentation, and special devices, carried by the payloads connected with the "Pioneer" satellite programs.

Mr. Lang was issued numerous patents in the above fields and has written several technical papers and articles.

He holds membership in the American Society of Mechanical Engineers and the I.S.A. He is an amateur radio station operator and he also holds a private airplane pilot's license.

(Continued on page 104A)

1960 Radio Engineering Show
March 21-24, 1960
New York Coliseum

SIZE 8

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400 Cycle: Many for 125 C operation . . . Higher for special applications

Many Immediately Available From Stock in Small Quantities

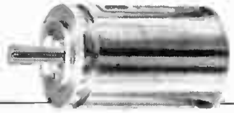
SIZE 8 SYNCHROS

Highly Stable. Minimum Error Variation from −55°C to +125°C

OSTER TYPE	CLASS	INPUT VOLTAGE	INPUT CURRENT AMPS	INPUT WATTS	OUTPUT VOLTAGE	PHASE SHIFT (° LEAD)	ROTOR RESISTANCE (OHM)	STATOR RESISTANCE OHMS	Z_{ro} OHMS	Z_{so} OHMS	Z_{rss} OHMS	NULL VOLTAGE (MV)	MAX. ERROR FROM E.Z. (MIN.)
4253-01*	LZ-CT	11.8	.087	.21	23.5	9.0	157.0	24.0	212+j722	28+j119	263+j69	30	±7
4269-01*	Diff	11.8	.087	.21	11.8	9.0	35.0	24.0	37+j139	28+j124	47+j13	30	±7
4273-01**	XMTR	26.0	.100	.54	11.8	8.5	34.0	12.0	48+j255	12+j45	82+j31	30	±7
4277-01*	HZ-CT	11.8	.030	.073	22.5	8.5	316.0	67.0	500+j1937	79+j350	594+j182	30	±7
4261-01**	Resolver	26.0	.043	.39	11.8	15.0	162.0	22.0	208+j612	34+j159	243+j77	30	±7

*Stator as Primary **Rotor as Primary

SIZE 8 SERVO MOTORS



OSTER TYPE	RATED VOLTAGES	Z = R + j X	IN. OZ. STALL TORQUE	RPM NO LOAD SPEED	WATTS PER PHASE	GM. CM. ROTOR INERTIA	LENGTH IN. MAX.	WEIGHT OZ.	T/I RATIO RAD/SEC²
5004-01	26V 26V	288 = 226 + j 176 294 = 238 + j 174	.15	6200	2.0	.47	0.863	1.2	22,500
5004-02	26V 36V	288 = 226 + j 176 526 = 409 + j 332	.15	6200	2.0	.47	0.863	1.2	22,500
5004-03	26V 40V	288 = 226 + j 176 715 = 582 + j 415	.15	6200	2.0	.47	0.863	1.2	22,500
5004-09	26V 40V	230 = 190 + j 131 519 = 399 + j 332	.20	6200	2.5	.47	0.863	1.2	30,000

SIZE 8 MOTOR TACH-GENERATORS

OSTER TYPE	RATED VOLTAGES	Z = R + j X	IN. OZ. STALL TORQUE	RPM NO LOAD SPEED	WATTS PER PHASE	GM. CM. ROTOR INERTIA	LENGTH IN. MAX.	WEIGHT OZ.	T/I RATIO RAD/SEC²	GENERATOR VOLTAGE	INPUT WATTS	OUTPUT VOLTS PER 1000/RPM
6204-01	26V 40V	230 = 190 + j 131 519 = 399 + j 332	.20	6000	2.5	.65	1.728	2.5	21,800	26	2.5	.25
6204-03	26V 26V	230 = 190 + j 131 230 = 190 + j 131	.20	6000	2.5	.65	1.728	2.5	21,800	26	2.5	.25

Oster

The Size 8 400 Cycle Servo Motor Tach Generators listed above have 150° max. cont. frame temperature. 110 MA input current, ±5° phase shift and Null Voltage (Total R. M. S.) of 15 millivolts.

OTHER PRODUCTS INCLUDE:

- Resolvers
- Computers
- Indicators
- Servo Mechanisms
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- DC Motors

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Call or write for illustrated bulletins and complete information on standard terminals, or send sketches and details for quotations on custom types you require.

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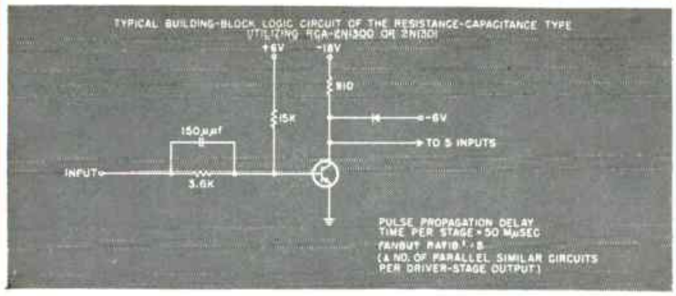
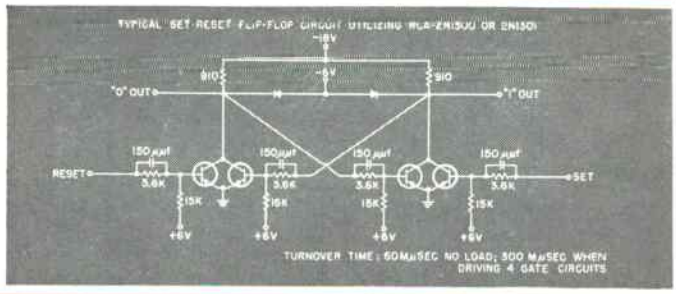
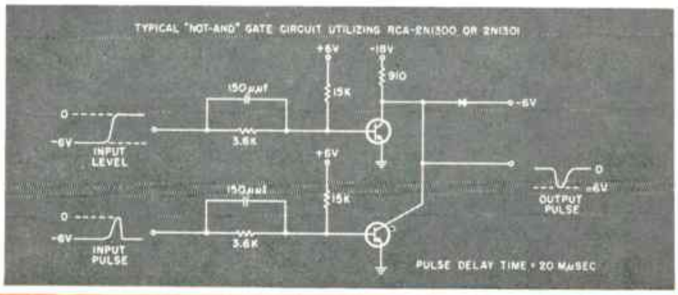
feature

HIGH POWER DISSIPATION
FAST-SWITCHING TIMES
AT LOW COST







AVAILABLE NOW... IN QUANTITY

RCA TYPE	Maximum Ratings—Absolute-Maximum Values							Characteristics: Common-Emitter Circuit, Base Input—Ambient Temperature = 25°C		
	Collector-to-Base Volts	Collector-to-Emitter Volts	Emitter-to-Base Volts	Collector Ma.	Transistor Dissipation Milliwatts			Minimum DC Current Transfer Ratio		Gain-Bandwidth Product▲ Mc
					at 25°C	at 55°C	at 71°C	at collector ma = −10	at collector ma = −40	
2N1300	−13	−12	−1	−100	150	75	35	30	—	40
2N1301	−13	−12	−4	−100	150	75	35	30	40	60

▲For collector ma = −10 and collector-to-emitter volts = −3

RCA's Germanium P-N-P Mesa Transistors 2N1300 and 2N1301 combine low-cost and quantity availability with these major benefits for designers of switching circuits:

- high power dissipation—150 milliwatts maximum at 25°C, 75 milliwatts maximum at 55°C
- fast switching times—made possible by high frequency response and low total stored charge
- rugged Mesa structure—with an extremely small base width to insure top performance at high frequencies
- high current transfer ratio—permits high fanout ratios (number of paralleled similar circuits per driver-stage output)
- high breakdown-voltage and punch-through voltage ratings—result of the diffusion process
- high current ratings—improves overall system speed
- especially well suited for use at pulse repetition rates up to 10 Mc
- rugged overall design—units have unusual capabilities to withstand severe drop tests and electrical overloads
- electrical uniformity—a result of the diffused-junction process used by RCA in the manufacture of Mesa Transistors

Contact your RCA Field Representative for prices and delivery. For technical data, see your HB-10 Semiconductor Products Handbook, or write RCA Commercial Engineering, Section A-35-NN, Somerville, N. J.

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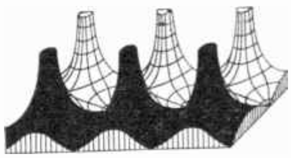
Poles and Zeros



Out of Focus. Preliminary data indicate that freshman classes of our engineering colleges have again declined in numbers, for the third successive year. The tentative enrollment figures for this fall show a seven per cent decline from the 1958 class entering engineering. This may indicate that with 1961 we begin an era of successively decreasing engineering graduations—at a time when our political scientists and economists predict we will be facing the greatest challenge of our history: the demands for production and advance in standard of living created by an expanding population, compounded by an ideological battle then approaching a climax.

We realize that graduation predictions are difficult, since there is no constant ratio between graduations and numbers of freshmen four years earlier. Much of our output now comes via one or two years at a junior college, or a regional university. This increases the number of advanced standing transfers-in, but we are also showing a declining ability to retain students entering an engineering college directly as freshmen.

There are those who interpret the enrollment decline as a direct result of raised entrance standards and increased emphasis on mathematics and the basic sciences, thereby claiming that the supply of engineers is being cut off. This is countered by the claim that engineering colleges in the past graduated too many who were technicians by modern definition (witness past low average salary levels); that what we are doing today is directing such students very properly to the technical institutes, thereby allowing the universities to concentrate on the education of the techno-scientist of the future. There are many who maintain that if the trend to more basic science for the engineer is continued, then we will merely be turning out poor counterfeits of the scientist—countering this are those who point out that unless we do educate engineers for tomorrow's scientific problems we will lose the battle to preserve our civilization.

Much of the above reflects the internal debate within the engineering profession on the direction in which engineering education should move. It is a result of a conflict of philosophies—is the engineer an applicator in private practice, bending the materials and forces of nature to the needs of his customers, or is he an applied scientist, employed to discover new knowledge and to apply that knowledge in the solution of the problems of an industrial age?

Both of the above views may actually result in useful educational processes, but resolution of the basic question is not for this space or audience. It seems possible, however, to discuss an associated but smaller question: Why are the high school graduates not turning to engineering for a career in the expected numbers?

We would like to propose that at least a portion of the difficulty lies in a failure of the communications channel to the high school—so that the change in the engineer, resulting from his employment in research and industry, has not been transmitted to young Johnny and his guidance counselor. To Johnny, the engineer may all too often be pictured as dressed in corduroys and knee boots, sighting through a transit, or referring to a handbook.

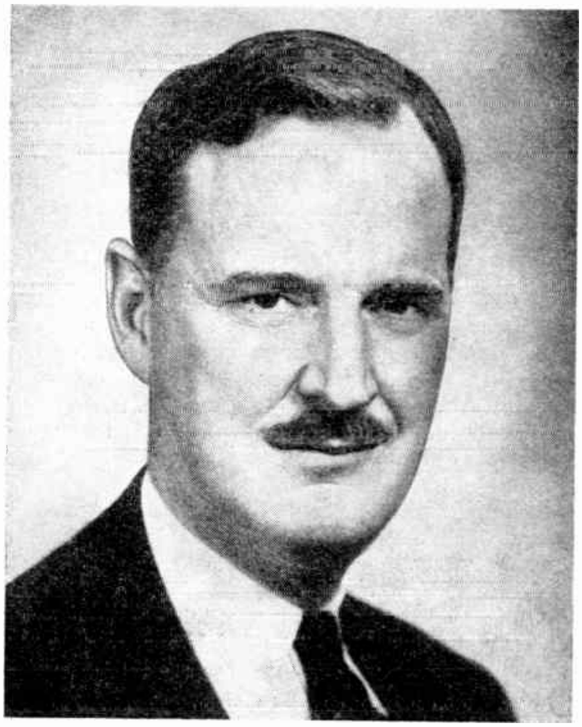
Young Johnny finds it difficult to fit this image of an engineer into a world of electronics, jets, and space navigation —a world which the press infers is really that of the scientist. No one has told Johnny that today an engineer may be dressed in a business suit, seated at a desk, calculating the pattern of a space antenna after appropriate deference to a computer in the next room.

Further, the guidance counselors seem confused by a barrage of information concerning an engineering shortage, while at the same time reading of engineering layoffs. No one has bothered to tell the counselors that many of those laid off were engineers only by their own claim, and were technicians in reality. We must also remember that the guidance counselor never had an engineering class in college; instead they heard how difficult engineering was. All these factors make it easier, and possibly safer, for Johnny to *not* take engineering, and so he may be advised. Johnny hears of achievements ascribed to science—the satellite transmits weather information—he believes that science offers an intellectual challenge in seeking the new—but engineers perform by empiricism and rote.

That Johnny is avoiding the misunderstood field of engineering in considerable numbers is shown by other recent enrollment statistics. Reduction in engineers seems offset by campus increases in physical science and mathematics; in fact the enrollment there has increased 10.4 per cent in one year, three times the gain to be expected due to general college enrollment increase.

The rapid rise in electronics enrollment further supports the argument, since Johnny can see a place for electronics in rockets and space. Since 1950, electrical engineering has become the major field and its enrollment has increased from 21.6 to 29.4 per cent of total engineers, while mechanical engineering has gained only 0.4 per cent and civil engineering has declined 2.7 per cent.

To summarize: Johnny seems convinced of the future in scientific things. Has our field attempted to make him aware that engineering is really applied science? What must be done, and by whom, to correct the image of the engineer at the high school level? Can we correct, or must we live down, the false impressions created by the engineering shortage publicity?—J.D.R.



Ronald L. McFarlan

President, 1960

Ronald L. McFarlan was born on March 8, 1905 in Cincinnati, Ohio. He attended the University of Cincinnati and subsequently the University of Chicago, from which latter institution he received the Ph.D. in physics in 1930. His next two years were spent as a National Research Council Fellow in physics at Harvard University, and were followed by three more years as an instructor at Harvard.

Dr. McFarlan's industrial career has included positions as chief physicist of the United Drug Company, and later the B. B. Chemical Company, and also as director of research for the Bulova Watch Company. For about eight years he was executive assistant to the director of equipment engineering, Raytheon Manufacturing Company, and is presently a consultant to the DATAmatic Corporation and the Raytheon Manufacturing Company.

Dr. McFarlan's early professional work was in the fields of X-ray diffraction and scattering, ultraviolet spectroscopy, and electronic instrument design and navigation. He has been associated with the management of projects including electronic digital computers, radar, automatic guidance and control, microwave communication, sonar echo ranging, and depth sounding equipment. He is the holder of several patents, and the author of a number of articles in the electronic and optical fields.

Dr. McFarlan is a member of Sigma Xi, the American Physical Society, the American Chemical Society, the Institute of the Aeronautical Sciences, and the American Society of Naval Engineers. He has served on the IRE Board of Directors since 1957 and the Executive Committee since 1958. He has been Vice-Chairman and Chairman of the Boston Chapter of the Professional Group on Engineering Management, Chairman of the Boston Section, Chairman of the Professional Group on Education, and a Senior Member of the IRE since 1951.

Scanning the Issue

Radio Transmission by Ionospheric and Tropospheric Scatter (Joint Technical Advisory Committee, p. 4)—At this writing the International Radio Conference, meeting in Geneva under the aegis of the International Telecommunications Union, is drawing to a close. The most important of all meetings in the communications field, this world-wide conclave has been in session since August of 1959 to review and revise as necessary the ITU regulations governing the international allocation and utilization of the radio spectrum which were formulated at the Atlantic City conference of 1947. Today we find our radio communication environment substantially changed from that of 12 years ago, not only by the influences of many technical advances but also by the growing problems of spectrum congestion. The authors of the report presented in this issue are no strangers to problems affecting the use of the radio spectrum. Formed in 1948 by the IRE and the Electronic Industries Association to assist the FCC, the Joint Technical Advisory Committee has achieved an extraordinary world-wide reputation for providing government and other bodies with disinterested and expert technical advice on frequency allocation and utilization matters. The present report is in essence a supplement to the widely known book "Radio Spectrum Conservation," published by the JTAC in 1952. It deals with two major advances in the communication art which have emerged since 1952 and which have an important bearing on the Geneva deliberations—namely, ionospheric scatter and tropospheric scatter. The report is divided correspondingly into two sections. Each section summarizes the present body of knowledge concerning propagation theories, experimental data, design practice, fields of application, and factors affecting frequency allocation. The JTAC felt this study to be so important that special arrangements were made with the IRE to rush 800 preprints of the report to Geneva delegates to the International Radio Conference well before its adjournment. It stands as the most authoritative and, at the same time, the most readable summary of scatter communication techniques that has yet been published.

The Ninth Plenary Assembly of the CCIR (Herbstreit, p. 45)—In order for the above-mentioned ITU Conference in Geneva to arrive at the proper decisions, engineering information of the type exemplified by the preceding JTAC report is needed on a host of technical questions covering essentially the whole radio engineering art. The principal source of this information is the CCIR—the International Radio Consultative Committee. This body is a permanent organ of the ITU. Its function is to study technical and operating radio problems of international interest and to recommend solutions for those problems. The CCIR held its Ninth Plenary Assembly in Los Angeles last April. Working through 14 Study Groups, the Assembly considered many new developments, from stereo broadcasting to space vehicle communications, which bear on the utilization and conservation of the radio spectrum. This paper, prepared especially for the PROCEEDINGS, summarizes the work of the 14 CCIR Study Groups at Los Angeles, important work which formed the basis for many of the decisions reached later at the Geneva conference.

Crosstalk Due to Finite Limiting of Frequency-Multiplexed Signals (Cahn, p. 53)—This investigation is concerned with the important problem of the effects due to clipping of a frequency-multiplexed signal as a result of amplifier saturation in a radio transmitter, particularly when the number of channels in the system is large. The results show that the amount of crosstalk caused by clipping is surprisingly low and that an optimum clipping level may be specified. This work will be of especial interest to, and should be understood by, communications engineers engaged in building and operating frequency-multiplexed digital communication systems, whether for telemetry or teletype. As an added asset, the general audience will find the presentation quite easy to follow, with the mathematical derivations in the appendixes.

IRE Standards on Methods of Measuring Noise in Linear Twoports (p. 60)—Few topics are as basic and broadly applicable to the radio communication art as noise is. For that reason it was decided to publish this material as a separate Standard rather than as one section of a larger Standard as originally planned. It defines and specifies the various ways in which the noise performance of transducers may be rated and then describes suitable methods of measuring noise performance. It is a companion to the paper that follows.

Representation of Noise in Linear Twoports (IRE Subcommittee on Noise, p. 69)—The IRE Subcommittee that prepared the foregoing Standard on noise measurements has prepared an excellent tutorial paper to go with it which provides, for those who need it, the theoretical background necessary for a full understanding of the Standard. The Standards Committee felt that a valuable service would be rendered by making this paper readily accessible to all.

Multipole Representation for an Equivalent Cardiac Generator (Geselowitz, p. 75)—The heart is a major center of electrical activity. It is run by current impulses which trigger each heartbeat. These impulses give rise to potential differences on the body surface which can be measured and recorded by an electrocardiograph to provide a partial picture of the functioning and condition of the heart. To provide a more nearly complete picture would require obtaining the potential distribution over the entire body surface, an impractical procedure. Researchers have therefore endeavored to develop a picture of the heart in terms of a hypothetical equivalent heart generator capable of producing the observed body surface potentials, instead of in terms of the surface potentials themselves. This paper is concerned with the key problem of finding a suitable equivalent generator. It provides information that may help to resolve this fundamental question and should certainly stimulate further work in this important area. In addition, the theoretical problem treated here concerning multipole sources in a conductor will interest many engineers outside the medical electronics field.

The Effect of a Cathode Impedance on the Frequency Stability of Linear Oscillators (Kohn, p. 80)—This paper sheds some new light on the perennially important problem of oscillator stability. It has been assumed in the past that the frequency instabilities contributed by the tube were mainly due to tube capacitances. The author shows that the growth of an interface layer between the cathode sleeve and the oxide coating causes the cathode impedance to change during lifetime to the point where this becomes an equally important, and previously overlooked, factor in the design of highly stable oscillators.

Multiple Diversity with Nonindependent Fading (Pierce and Stein, p. 89)—Diversity reception is now a widely used technique for overcoming the effects of fading on HF and scatter communication circuits. Previous analyses have assumed that fading of the several received signals occurs in a randomly independent fashion. In a number of situations, however, practical limitations prevent separating the channels sufficiently in space or frequency to insure that fading will indeed be independent. A considerable number of people have tried to analyze the nonindependent fading case recently but have had to give up on it because of its difficulty. The authors of this paper have succeeded. The results will be of interest to a broad group of workers concerned with many types of communications systems, and of direct utility to the small but highly significant group responsible for the design and evaluation of such systems.

Scanning the Transactions appears on page 126.

RADIO TRANSMISSION BY IONOSPHERIC AND TROPOSPHERIC SCATTER

A Report of the
JOINT TECHNICAL ADVISORY COMMITTEE
JTAC

Foreword

THIS report on "Radio Transmission by Ionospheric and Tropospheric Scatter" has been prepared as a supplement to "Radio Spectrum Conservation," an earlier comprehensive report which was completed by the Joint Technical Advisory Committee and published by the McGraw-Hill Book Company, Inc., in 1952. The purpose of "Radio Spectrum Conservation," as stated in the Introduction thereto, was "to analyze and evaluate the current uses of the radio spectrum and to formulate constructive suggestions for the future."

Soon after publication of the report "Radio Spectrum Conservation," two new methods for beyond-the-horizon extended-range radio transmission began to emerge as major advances in the communication art. The importance of these new radio techniques, now usually described as "ionospheric scatter" and "tropospheric scatter," led the Joint Technical Advisory Committee in February, 1955, to establish an Ad Hoc Subcommittee on Forward Scatter Transmission. It was the task of this Subcommittee to compile factual data on the new scatter techniques and to prepare a report on "Radio Transmission by Ionospheric and Tropospheric Scatter" to supplement the book "Radio Spectrum Conservation."

In 1955, the Subcommittee, in cooperation with the IRE Professional Group on Antennas and Propagation, sponsored publication of the October, 1955, Scatter Propagation Issue of the PROCEEDINGS OF THE IRE. This special issue aided greatly in consolidating and disseminating authoritative information on scatter transmission—and remains a major and lasting contribution to the technical literature.

Recently, the Subcommittee completed preparation of the present report which, hopefully, will serve usefully to augment "Radio Spectrum Conservation" (particularly Chapter 2—*Propagation Characteristics of the Radio Spectrum*), as well as to summarize those unique features of ionospheric and tropospheric scatter which now must be considered in planning for future efficient utilization of the radio spectrum.

The members of the JTAC Ad Hoc Subcommittee on Forward Scatter Transmission were: J. B. Wiesner, *Chairman*, W. G. Abel, L. G. Abraham, D. K. Bailey, H. H. Beverage, K. Bullington, J. H. Chisholm, H. V. Cottony, R. C. Kirby, W. E. Morrow, Jr., K. A. Norton, W. H. Radford, J. F. Roche, T. F. Rogers, and R. J. Slutz. Messrs. R. M. Davis, Jr., R. G. Merrill, V. R. Eshleman, and A. D. Wheelon assisted in the preparation of Sections 2 and 3 of Chapter 1—*Ionospheric Scatter Transmission*. Compilation of this report was completed under the guidance of W. H. Radford.

I. Ionospheric Scatter Transmission*

1. Introduction

THE large body of knowledge which we now have concerning the downward scattering of low-angle VHF radio energy from the lower ionosphere[1] began to be acquired in the early 1950's. Sufficient information was obtained by late 1952 or early 1953 to support the conclusion that this mode of radio wave propagation is capable of supporting a communication circuit having certain unique properties. Soon thereafter, this relatively new propagation mode was utilized for important military point-to-point communication circuits in the North American Arctic and North Atlantic Ocean regions, as shown in Fig. 1-1; more recently, a network has been established by NATO extending ionospheric scatter communication from Norway through Western Europe and to Turkey. A one-way experimental circuit for civil communication has been used between The Netherlands and Italy, and expansion to full duplex operation is planned. Plans are already far advanced for its exploitation in several other areas. Certain specialized applications for air/ground communication are under consideration, but it is fair to say that point-to-point services represent virtually all the potential application.

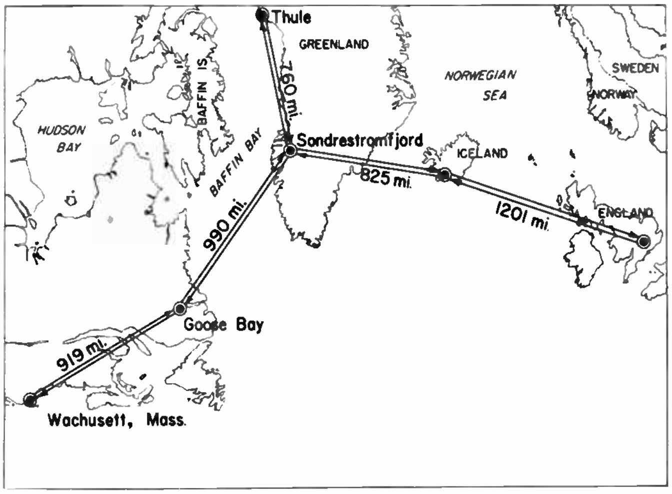


Fig. 1-1—North Atlantic VHF ionospheric scatter system—USAF.

Two basic methods have been employed to establish a useful communications circuit with the use of D-region scattering. One method employs the continuously present but weak and variable background scatter signal level for continuous transmission at speeds typical of HF services; the other stores the traffic at the terminals and subsequently transmits it in a discontinuous manner at high speed when a large "burst" of signal energy, arriving via scattering from a meteor induced ionization trail, closes a duplex circuit. By far the greatest application has been made of the continuous transmission technique to date, and it is this technique and the continuous propagation mode which are under consideration in this report.

From the viewpoint of the communication engineer, the understanding of certain aspects of D-region scattering is incomplete at this time. Although a few basic studies of this phenomenon were performed as long as three decades ago, experiments of quantitative value have been undertaken only within the last decade. Such experiments, because of the very weak and variable nature of the scattered field strengths, require the use of large and expensive equipment for rather long periods of time. Yet, such experiments must be conducted and evaluated with attention to detail over long periods in order to minimize the cost of the eventual communication plant design. Sufficient information is certainly now at hand to permit adequate circuit designs to be made for many purposes.

The major characteristics of communication circuits utilizing the continuous ionospheric scatter mode may be summarized as follows: utilizing a large effective radiated power at one or two frequencies in the 30–60 mc range, large receiving antenna arrays, sensitive receivers, and modern modulation, detection, coding methods, a duplex point-to-point multiplex teleprinter circuit can be reliably maintained throughout the distance region of some 700–1200 statute miles. A few of these circuits can be placed in tandem. Of particular note is that this reliability is obtainable throughout the auroral regions where the effectiveness of conventional high frequency ionospheric propagation is often most seriously disrupted.

A VHF ionospheric scatter circuit is similar to a HF sky-wave circuit in that a distance of 1000 miles or somewhat more may be covered. Large antennas and transmitter powers are required; the received signal level varies over many orders of magnitude. The traffic handling capacity is relatively low compared with a typical 300-mile tropospheric service (though comparable with good HF services). The influence of multipath and interference must be dealt with in design for its effective use. Unlike an HF circuit, essentially single-frequency or at the most dual-frequency operation is permitted. An individual path length is limited to 1300–1400 statute miles in most locations. Atmospheric noise does not play nearly as important a role, especially in

* Original manuscript received by the IRE, October 19, 1959.

[1] Much of the literature has alternatively referred to ionospheric scatter as from the *lower ionosphere*, the E-*region*, or the D-*region*. The term D-*region scattering* has been adopted for use throughout this report; the height of scattering has been measured, at least in temperate latitudes, to lie in the D-region below about 95 km down to about 70 km.

arctic regions, and all but the most intense polar absorption effects at the lower frequencies have little influence on circuit reliability when a proper circuit design is employed. Occasional circuit interruptions have been encountered at frequencies in the 30- to 40-mc range during such events, and on experimental paths at higher frequencies which used broad-beam antennas.

The D-scatter mode resembles in some respects the UHF tropospheric mode, as regards single-frequency operation reliability even in arctic locations, long individual path distances, and the necessity for sophisticated diversity methods to obtain high short-time reliability. However, the need to keep the scattering angle small restricts its use to paths longer than 600–700 miles. At shorter ranges, tropospheric scatter is more advantageous, as suggested by Fig. 1-2.

Contemporary VHF ionospheric scatter circuits operate in the 35–50-mc region and use antennas having 20–25 db gain (relative to an isotropic antenna in free space). The antennas are so sited as to have the radiation pattern main lobe directed to a scattering region about 85 km high at midpath over the great-circle route. A transmitter power of 10–50 kw is relied upon, along with dual space-diversity reception and efficient modulation detection methods, to provide the equivalent of about a dozen 60-wpm multiplex teleprinter channels having a character error of no more than 1:1000 for about 99 per cent of a year's hours.

It is important to appreciate that estimates of circuit reliability and capacity made here are, as in the case of long-distance tropospheric circuits, related specifically to the level of technical development already achieved and exhibited in the use of VHF ionospheric scatter. Current research on the scatter mode of radio wave propagation and improved terminal equipment may be expected to lead to different and more economical requirements for antennas and power.

As regards development of civil international fixed services, it should be noted that the International Radio Consultative Committee (CCIR) of the International Telecommunications Union has already adopted official technical questions and study programs leading to recommendations for engineering practice and standards of international connections of fixed services using ionospheric scatter propagation.

No attempt to summarize general long-distance radio communication design features, requirements, and characteristics is made in what follows, since the literature is both rich and advancing in this field; rather, only the aspects of circuit and component design peculiar to the exploitation of the D-region scatter mode are considered.

A very brief representation is given in Section 2 of the theories seeking to explain the observed propagation, and Section 3 summarizes the principal experimental data affecting engineering consideration of D-

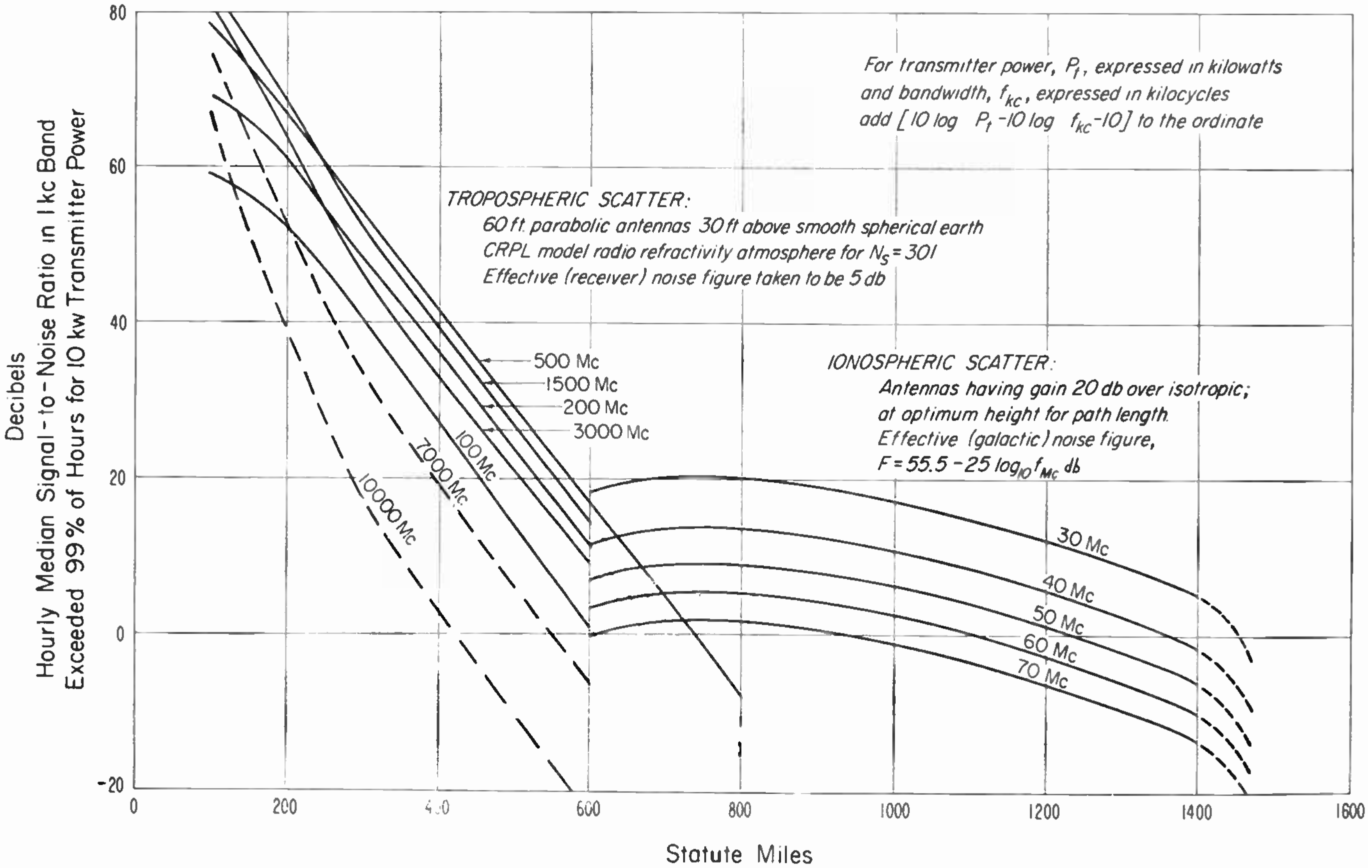


Fig. 1-2—Comparison of distance and frequency dependence of ionospheric and tropospheric scatter.

scatter for communication. Section 4 deals with current design practice.

2. Theories of Scattering from the D-Region

2.1 Introduction

Scattering from the D-region is observed at well above regular ionospheric reflection frequencies as a continuous, rapidly fading signal, with bursts of energy superposed. The continuous signal has been variously attributed to scatter from irregularities in ionization in the D-region, partial reflections from gradients of ionization, and overlapping reflections from the ionized trails of many small meteors. The superimposed bursts are reflections from large meteor trails of sufficient density. This section mentions briefly theories proposed to explain the continuous signal on the basis of turbulent irregularities and meteoric ionization, as well as suggestions concerning the sources of ionization in the D-region. Theories of scattering from irregularities in electron density involve statistical analysis of the density fluctuations and of the parameters of the turbulence. Continuous scatter from meteor trails involves integration over a given volume of trails, as well as the assumption of a distribution of sizes, numbers, and radiants of small meteors. Partial reflections from gradients of electron density are analyzed by electromagnetic theory. Partial reflection theory (advanced by Feinstein and Salzburg) is not discussed further, since an extension of this theory seems to be required to account for the propagation as observed.

2.2 Ionization in the D-Region

General evidence for assuming a solar origin for the E-layer ionization is given by Appleton. The maximum ionization density (about 10^5 electrons per cubic centimeter) varies by 50–60 per cent with the sunspot cycle. Solar eclipses show that the ionizing radiation travels with the velocity of light, and soundings made during eclipses verify the solar dependence. Rocket soundings show that the agent is probably soft X rays. Similar solar dependence is associated with the D-region, which has a typical electron density of from 100 to 1000 electrons per cc.

The cause of ionization in the D-region in temperate latitudes is generally taken to be photoionization of nitric oxide according to

$$NO + h\nu(>9.5\ eV) \rightarrow NO^+ + e,$$

by Lyman α radiation (λ 1216 Å), which penetrates effectively to a height of 70 km. Atmospheric motions, causing fluctuations in NO concentrations, account for variations in the D-region ionization. Various radio methods have long shown a sudden increase in ionization at very low heights, down to about 65 km, during a solar flare; rocket measurements have shown an increase in Lyman α and X rays during such times.

Experimental results, including rocket measurements, over the past two decades have given ample evidence of stratification of the D-region, *i.e.*, the existence of discrete horizontal layers of ionization below the E-region, down to about 50 km. Further, it is clear that the region exhibits motions at all levels, and that influence of winds and turbulence is an important physical aspect of the D-region. Velocities change with height, sometimes abruptly, so that sharp wind shears occur.

Probably the most consistent feature of radio observations of the D-region (at frequencies from VLF through VHF) is the presence, day and night, of a reflection height of about 85 km at temperate latitudes. The height appears to be associated with the temperature minimum (mesopause) and is the same height at which sharp maxima are found for twilight sodium D-radiation, the occurrence of noctilucent clouds, and the twilight scattering of sunlight from dust. Radio reflections are usually very weak, and several reflection heights are observed simultaneously. Individual heights below the 85-km region are observed to remain constant for periods of many hours but may be different on different days. Heights below 80 km are usually observed in the daytime only, and most often in winter, on several consecutive days.

At high latitudes, additional complex ionization processes, such as high-energy particle bombardment, often dominate in the D-region. Even the height distributions of electron density are not well known for a wide range of disturbed conditions prevailing in arctic regions.

2.3 Scattering from Irregularities; Turbulence

The initial experiments to observe ionospheric scattering were supported, and to some extent suggested, by the turbulence theory for tropospheric scattering. Turbulence can cause irregularities of electron density in an ionized medium through pressure fluctuations associated with turbulence, or by randomization of a mean gradient in electron density or temperature. A medium containing such irregularities in electron density can give rise to incoherent scattering of radio waves.

a) Electromagnetic Scattering by Irregularities: A useful expression for frequency and geometric dependence of *transmission loss* p_r/p_a is given for antenna gain kept constant with frequency (*i.e.*, scaled antennas):

$$\frac{p_r}{p_a} \sim l^2 f^{n_s} \left(\sin \frac{\theta}{2} \right)^{n_s - 1}$$

where

p_r = the radiated power,
p_a = the available power from the receiving antenna,
l = the distance from the scattering volume to the receiver,
f = frequency,
θ = angle through which the scattering takes place,
n_s = frequency exponent for constant gain (scaled antennas).

b) Turbulence Theories: The spectrum of turbulent electron irregularities is the physical parameter of central interest for scatter propagation. More particularly, one is interested in the spectrum at wave numbers $k=4\pi/\lambda \sin \theta/2$, which lie in the following range for typical VHF scatter paths:

$$0.2 \text{ (meters)}^{-1} < k < 1.3 \text{ (meters)}^{-1}.$$

On the other hand, the wave number k_s, corresponding to smallest eddy size at which diffusion viscosity effects become important, may lie within the significant range of interest.

Three distinct physical theories have been developed in recent years to describe the irregularities of electron density established in the turbulent velocity field of neutral ionospheric gases:

1) theory of pressure fluctuations
2) Obukhov-Corrsin mixing theory
3) Villars-Weisskopf mixing theory.

The theory of pressure fluctuations has essentially been abandoned as a major source of ionospheric irregularities. The two mixing theories agree that a random convection mixing process characterizes the production of irregularities. The Obukhov-Corrsin theory is patterned on the Kolmogorov theory of the velocity spectrum. It assumes that the input and inertial ranges do not overlap. The Villars-Weisskopf theory notes the gradient of ionization established in ionospheric layers, and considers turbulent motions of the neutral carrier transfer electrons from low to high density points on the gradient's profile, and vice versa. These intruding cells appear as fluctuations in electron density and scatter accordingly.

In any case, because the scattering process acts like a narrow-band filter, emphasizing the uniquely important wave number k, it is quite clear that the spectrum of irregularities $S(k)$ is the basic description of electron density fluctuations required of the turbulence theories in order to predict electromagnetic wave scattering.

2.4 Scattering from Meteoric Ionization

Tens of billions of meteoric particles enter the upper atmosphere each day. They produce long (average of 25 km) trails of ionization at a mean height of about 93 km, with nearly all of the trails being formed between 80 and 120 km. The initial diameter of a trail is less than one meter, but it subsequently expands due to normal molecular diffusion. After perhaps half a minute, the rate of expansion may increase rapidly as a result of eddy diffusion. The initially straight trails are also contorted into serpentine shapes by wind shears and turbulence.

There are several ways that meteors may contribute to the scattering mechanism in ionospheric scatter propagation. 1) Intense signal bursts are caused by trails produced by the larger meteoric particles. These signals usually fluctuate or fade as the dense trail is distorted by wind motions. 2) Weak overlapping signals may be scattered from the straight trails produced by numerous small meteors. 3) Weak continuous signals may be scattered from a large number of both dense and tenuous trails which have become very distorted and mixed into a spaghetti-like configuration. 4) Meteoric dust may collect at the thermopause and provide there a layer that can be ionized easily by solar radiation. Turbulent mixing could then produce inhomogeneities of ionization which scatter the VHF waves. Some success has been achieved in the determination of the relative importance of 1) and 2) as compared with the signal components scattered from other sources. However, the signals described in 3) and 4), if they exist, may be indistinguishable from the signals which may be scattered from inhomogeneities produced by turbulent motion in a region of non-meteor-produced ionization.

a) Scattering from Individual Trails: The radio energy scattered from newly formed meteor trails is highly directive. We say the reflection is specular; the incident and reflected ray-paths make equal angles with the trail axis. As the trail expands and contorts, the echo strength decreases and the width of the scattered beam increases. The decay of echo strength is approximately exponential for trails whose line density of ionization is less than about 10^{14} electrons per meter (low-density trails). The strength becomes very small before appreciable beam-broadening can occur. The decay of echo strength is slower and less regular for the high-density trails which have line densities in excess of 10^{14} per meter. If the echo persists for the order of 10 seconds, appreciable energy may be scattered in all directions. But the concept of echo persistence implies a threshold of detectability. After the time interval above, there may be minute amounts of energy scattered in all directions, even from low-density trails.

While a meteor trail is being formed, some energy is reflected from the newly created ionization near the meteoric particle. The motion of the reflecting point causes a Doppler "whistle" when this energy is added to the continuous scatter signal. This whistle descends in pitch while the particle approaches the point of specular reflection, and rises as it recedes. After the trail is formed, it may be translated by winds and large-scale turbulence. This motion of the trail also produces a small frequency shift of the scattered energy.

b) The Total Meteoric Signal: In order to find the total effect of scattering from a large number of meteor ionization trails, we must consider not only the characteristics of individual echoes but also the distributions in density and orientation of the trails. Meteoric particles exist over a wide range of mass, with the more massive particles being less numerous. The more massive particles produce trails of higher line density. The number of trails having more than a certain line density of electrons is approximately inversely proportional to this line density. For instance, there are approximately ten times as many trails formed with line densities greater than 10^{13} electrons per meter than with 10^{14} electrons per meter.

From these considerations, it has been calculated that the median signal level supported by overlapping echoes from meteor trails should vary as f^{-7} for scaled antennas, or as f^{-5} on a per unit aperture basis. This frequency dependence corresponds roughly to the average of the measured median values. The transmission loss for a continuous scatter signal supported by meteors has been estimated from the best available data on the average influx rate of meteors and found to be in order-of-magnitude agreement with the measured values.

More meteor trails are formed during the morning hours than during the evening hours because of the motion of the earth in its orbit about the sun. The bunching of meteors toward the morning, or leading side of the earth, is perhaps the strongest factor controlling the number and directions of arrival of meteoric particles.

Beam swinging experiments provide the best evidence for the relative effects of specular scattering from meteor trails and scattering from other sources. For off-path narrow-beam antennas and for broad-beam antennas, the specular meteoric component appears to be the dominant signal source. For on-path narrow-beam antennas most of the signal may result from scattering at turbulent layers, especially during the day. These These layers could include nonspecular meteors or ionization produced by meteoric dust.

2.5 Table of Theoretical Frequency and Angle Dependence

Frequency and Angle Dependence* of Transmission Loss Given by Theories of Scattering from Ionospheric Irregularities (Turbulence), Partial Reflections, and Integration of Meteor Reflections

Theory	Inertial Range $\frac{4\pi}{\lambda}\left(\sin\frac{\theta}{2}\right) \ll k_s$		Dissipation Range $\frac{4\pi}{\lambda}\left(\sin\frac{\theta}{2}\right) \gg k_s$	
	n_s	References†	n_s	References†
Turbulence				
1) Pressure fluctuations	19/3	Villars and Weisskopf [31]	13	Villars and Weisskopf [31] Wheelon [33]
2) Villars-Weisskopf mixing	7	Villars and Weisskopf [32] Wheelon [33], [34] Gallet [20]	15	Villars and Weisskopf (unpublished)
3) Obukhov mixing	17/3	Obukhov [25c] Corrsin [13] Silverman [29] Bolgiano [8] Batchelor [6]	29/3	Batchelor [6] Batchelor, Howells, and Townsend [7]

	n_s	References
Partial Reflections	5	Feinstein and Salzburg [19]
Continuous Meteor Propagation	7	Villard, Eshleman, Manning, and Peterson [30] Eshleman and Manning [17]

* n_s is used because of general applicability to transmission loss without regard to mechanism; for scattering cross section σ, $n = n_s - 2$.
† Numbered references are as listed in the Bibliography for Section 2.

3. Experimental Observations of Propagation Characteristics

3.1 Measured Signal Characteristics and Their Dependence on Frequency

a) System Loss: The hourly values of system loss[2] discussed in this section correspond to hourly median values of the randomly fading signal.

In the frequency band between 25 and 108 mc, ionospheric scatter signals have been recorded over paths of 1000–2000 km with values of system loss ranging between about 140 to 210 db. The level depends on many factors, including frequency, type of antenna and orientation, geographic position, season and hour.

1) Diurnal, seasonal and annual variability: Fig. 1-3 shows the diurnal variation of the scatter signal near 50 mc in three different seasons: winter, summer, and equinox. The curves represent transmission over paths of about 1200 km with narrow-beam (6°) rhombic antennas aimed along the Great Circle, at both middle and high latitudes. Fig. 1-3 gives the levels of signal intensity which were exceeded 10 per cent, 50 per cent, and 90 per cent of the time during the month.

A very general characteristic of the scatter signal seems to be a diurnal minimum between about 1900 and 2100 hours local time at the path midpoint. In middle latitudes the diurnal maximum occurs in the daylight hours and is broader in summer than in winter. At high latitudes the diurnal maximum is broader and

[2] "System loss" is defined as the ratio of power available at the receiving antenna to the power delivered to the transmitting antenna, excluding transmission line losses. CCIR Recommendation as described by K. A. Norton, "System loss in radio wave propagation," Proc. IRE, vol. 47, pp. 60–62; September, 1959.

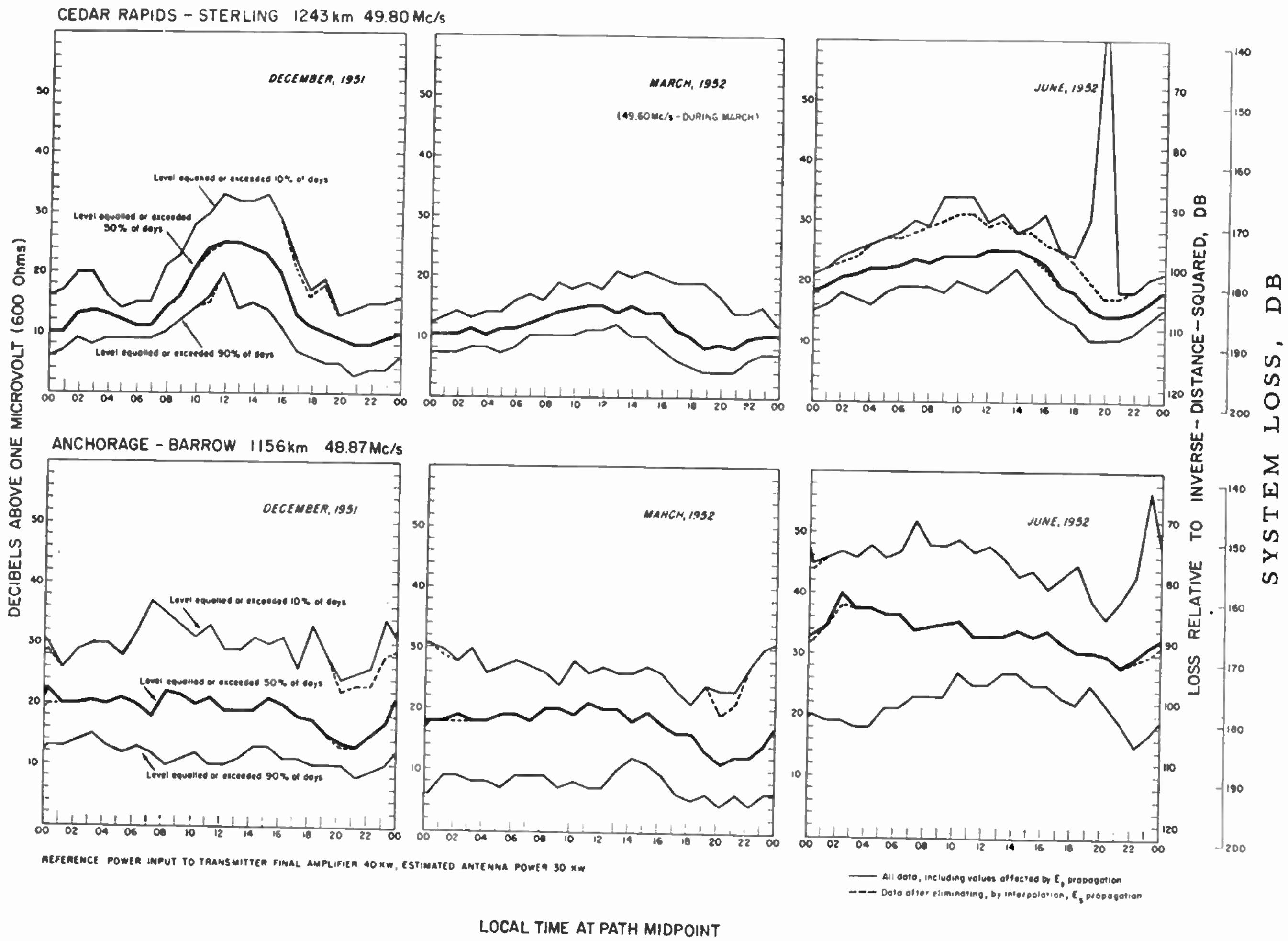


Fig. I-3—Diurnal variation of system loss.

tends to center earlier in the day. System loss values vary considerably from day to day at a given hour. The variation is greater in arctic regions than at temperate latitudes. The median value for a given month varies less from year to year than it does between adjacent months of the same year.

As shown in Fig. I-4, the seasonal variation of signal intensity at mid-latitudes has a strong summer maximum at all hours of the day. Daytime levels also show a secondary winter maximum. At high latitudes a very strong summer maximum is observed.

If attention is confined to the weakest received signals, those exceeding 95 per cent of the hours of a month, the level is fairly uniform through winter and equinox but rises to a maximum during summer.

In Fig. I-4, some year-to-year variation is apparent, though this is small compared to seasonal variation within a year.

2) *Frequency dependence:* Experimental studies of dependence of transmission loss on frequency have given values for n_s in

$$\frac{p_r}{p_n} \sim f^{n_s}$$

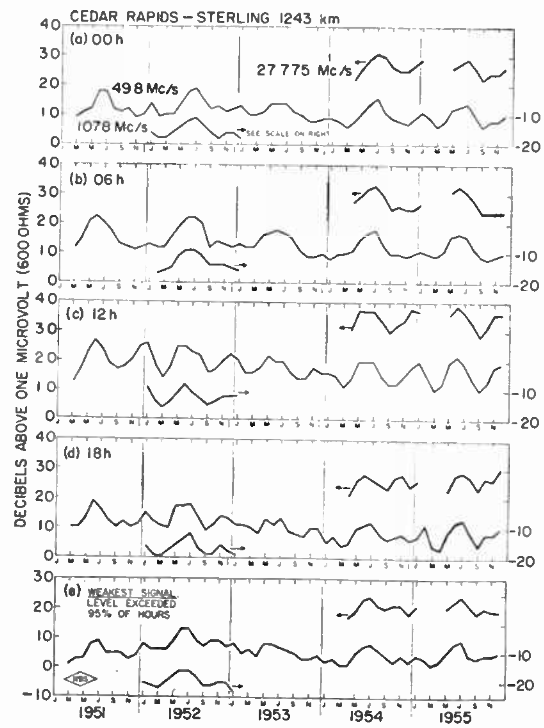


Fig. I-4—Seasonal variation of signal intensity.

for constant gain antennas (antennas scaled with frequency).

The results of early studies indicated that the exponent is itself a function of frequency. A recent improved experimental study gives results indicating that the exponent is uniform over the frequency range 30–108 mc for a particular path and time. The results of this five-frequency experiment are illustrated in Fig. 1-5, showing how well the observed values of signal intensity (in decibels) *vs* log frequency fit a straight line.

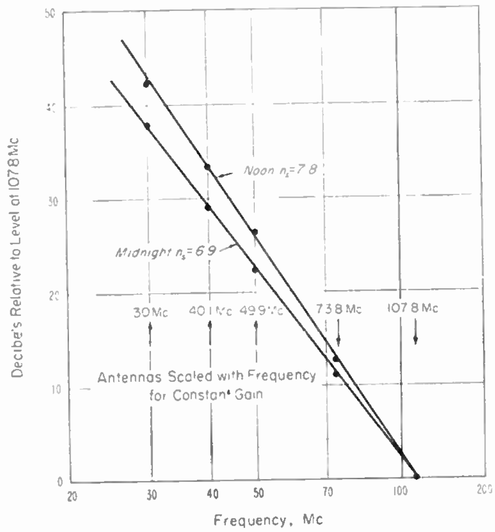


Fig. 1-5—Dependence of signal intensity on frequency for 10 days of June, 1958 on Long Branch, Ill. to Boulder, Colo. path.

The value n_s varies with time of day and season from about 6 to 10. A value of $7\frac{1}{2}$ has been taken as representative for system performance calculations.

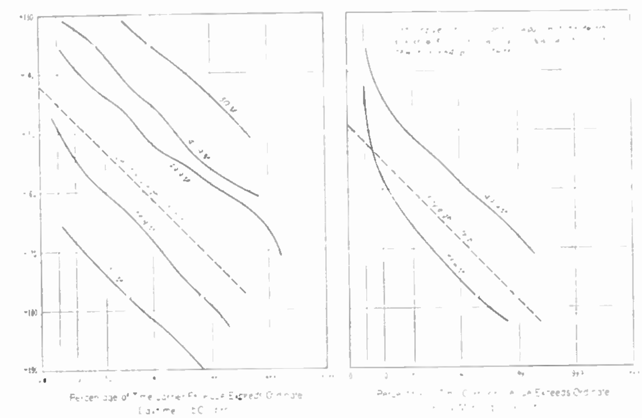
The exponent is generally higher in the daytime than at night. Seasonally, the exponent is relatively high in the months around equinox, low in summer, and probably low in winter. Substantial changes in the exponent can occur from one hour to the next. A change of 1.0 is common and a change of 2.0 occurs occasionally.

A limited amount of data indicates that the frequency dependence of received power in the Arctic is approximately the same as at middle latitudes. The measurement of the frequency exponent at high latitudes is complicated by conditions of high absorption that often prevail. During these conditions, the signal at low frequencies is attenuated more than at high frequencies, giving rise to apparently low exponents. When this influence was removed, the median exponent for a circuit in Alaska is found to be comparable to that observed on a temperate latitude path.

The diurnal variation of the frequency exponent, and perhaps its seasonal variation as well, are related not only to the variation of the parameters of turbulence, but also to the diurnally and seasonally varying relative influences of turbulence and meteoric ionization. Absorption effects must also be considered.

b) Short-Time Signal Statistics: High-speed recordings of the fading characteristics of the signal have been analyzed for amplitude distribution and fading rates of the envelope. Studies of phase stability have also been reported. (See Section 3.6).

Recognizable bursts of a signal from meteor reflections are found every few seconds superposed on a background of continuously fluctuating signal, presumably containing the effects of scattering from irregularities as well as the overlapping effects of many small meteors not individually recognizable. Observed amplitude distributions are illustrated by Fig. 1-6.



(a) (b)

Fig. 1-6—Typical amplitude distributions of ionospheric scatter signal at frequencies from 30 to 108 mc, obtained simultaneously on Long Branch, Ill. to Boulder, Colo. path (May, 1959) using narrow-beam antennas (20 db gain over isotropic).

During most of the day the envelope is approximately Rayleigh distributed; during early morning hours the influence of meteor bursts is evident in the distributions. Fading rate is found mainly in the range of about 0.2 to 5 cps for a 50-mc carrier frequency, using a narrow-beam on-path system. Expressing fading rate as proportional to frequency raised to some exponent, values are observed for the exponent from about 0.75 to about 1.25. Fading rate varies from about 0.3 to 1.0 cps at 30 mc and from about 1.0 to 2.0 cps at 108 mc. Fading rate is much higher for broad-beam or off-great-circle systems than for on-path narrow-beam systems.

An important additional feature superposed on the fluctuating signal is the frequent occurrence of Doppler-shifted meteor "whistles" as illustrated in Fig. 1-7. These components are produced by reflections from the heads of ionization trails and by scattering from the rapidly lengthening trails during their formation. Observations of the distribution of Doppler frequency shifts have been made on a VHF path at 50 mc and are illustrated in Fig. 1-8. The maximum theoretical expected Doppler shift is 6 kc at 50 mc. Although the average energy contained in the whistle components is small, occasional errors can be caused in narrow-band frequency shift systems unless special precautions are taken, as discussed in Section 4.

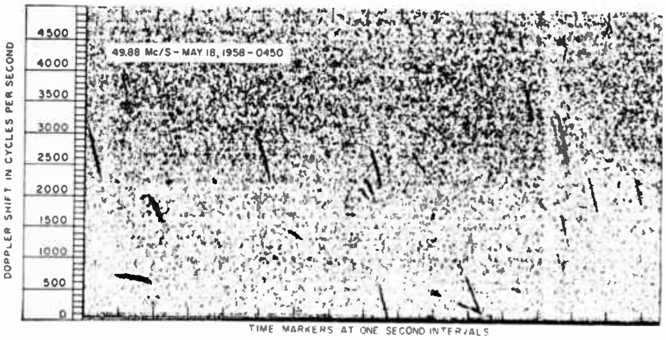


Fig. I-7—Spectra of meteoric Doppler-shifted signals over a 20-second interval.

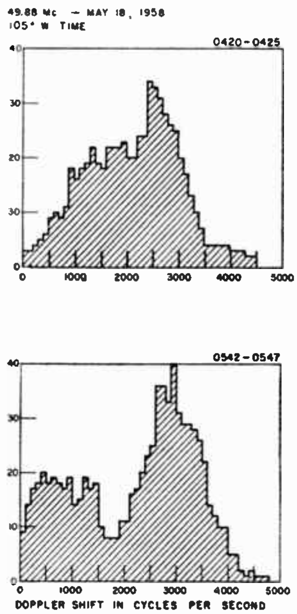


Fig. I-8—Total meteoric Doppler-shifted components observed for selected time segments.

c) Space Correlation: It has been found that correlation of signals received at antennas spaced along the great circle path is much greater than that for antennas spaced perpendicular to the path. Measurements obtained at 50 mc using narrow-beam transmitting antennas showed that the correlation coefficient falls to 0.5 at only 3.5 wavelengths normal to the path, while a separation of 40 wavelengths corresponds to the 0.5 coefficient along the path. The correlation of signals received at spaced antennas depends to a considerable extent on the antenna beamwidths.

d) Frequency Correlation: The extent of correlation of fading across a frequency band affects the analogue modulation bandwidth which may be usefully employed and the design of frequency diversity techniques. Using narrow-beam (6°) antennas at 50 mc, the correlation coefficient is observed to fall to about 0.5 for frequency separations of 6 kc during the early morning hours; during midday the correlation remains higher, *e.g.*, about 0.65 for a frequency separation of 6 kc, falling to 0.5 at about 7 kc separation. The correlation would be poorer for broad-beam antennas or off-great-circle transmission.

3.2 Signal-to-Noise Ratio

Galactic noise is an inherent limitation in all VHF scatter systems. At times, other types of noise, such as from thunderstorms and precipitation or man-made noise, limit the operation of scatter circuits. These difficulties occur for only a relatively small proportion of the time, however, and a fairly accurate estimate of the reliability of a given communications system at temperate and arctic latitudes can be made from a knowledge of the distribution of the signal-to-galactic-noise ratios. Special consideration is indicated for areas where local thunderstorm prevalence is especially high.

Fig. I-9 shows the cumulative distributions of the observed signal-to-noise ratios at 50 mc on three different paths over the period of one year. The curves are for a transmitter power of 30 kw and a bandwidth of 2 kc.

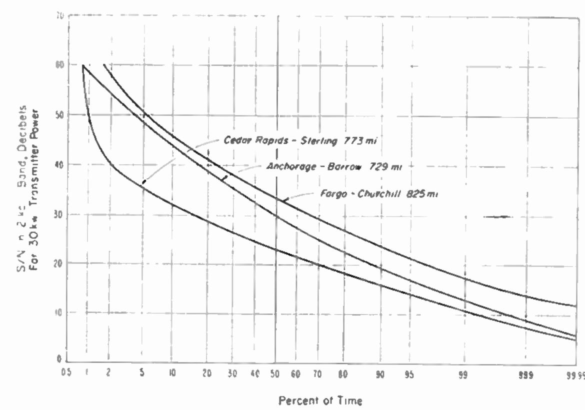


Fig. I-9—Cumulative distribution of hourly median signal-to-noise ratio through a year for representative ionospheric scatter paths at 50 mc using narrow-beam antennas having 20 db gain over isotropic.

3.3 Geometrical Factors and Antenna Directivity Studies

a) Pulse Delay Times; Heights of Scattering: A knowledge of the ionospheric heights at which signals are returned by the scatter mode is of prime importance to antenna design, distance dependence studies, and to various theoretical investigations; the determination of the heights of scattering has received considerable attention.

The earliest estimates of scattering heights were made comparing the reception at three sites at different distances along a line from the transmitter. Calculations of distance dependence of transmission loss were made for a series of model ionospheric scattering heights; by comparing the predicted rate of decrease with distance with the observed rate, the scattering height could be deduced. It was found by this method that predominant daytime heights were between 75 and 80 km, and nighttime heights were between 85 and 90 km. These observations were made during the spring season.

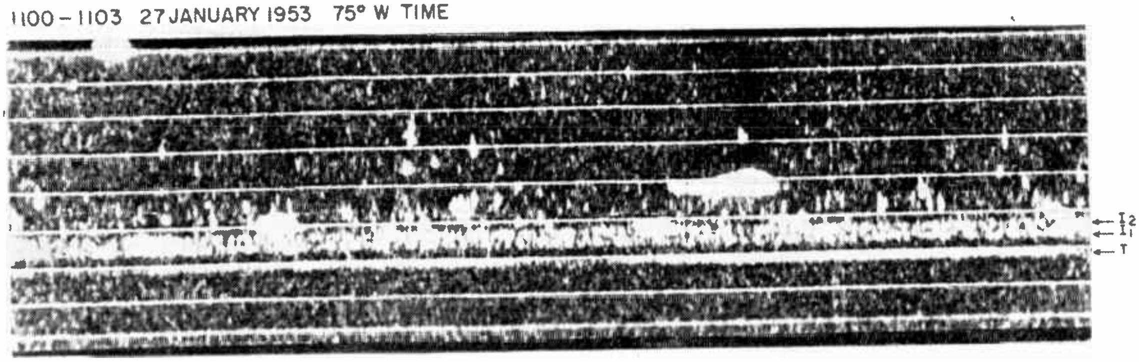


Fig. 1-10—Range time record of pulses received over an 811-km path. T is tropospheric signal; I_1 and I_2 are ionospheric signals with relative delays corresponding to 70 and 88 km.

In an experiment based on the comparison of the delay times of tropospheric and ionospheric pulses received over an 811-km oblique path, the results indicated that at night, scattering took place at 85 to 90 km. Some scattering from this stratum continued in daylight hours, but the most effective region for daytime scattering was a stratum at 60 to 75 km. These observations were obtained in November, December, and January. Fig. 1-10 is a typical record obtained from such observations, showing returns from two ionospheric strata, the tropospheric return, and returns from meteor bursts delayed from zero to several hundred microseconds relative to the continuous signal.

A summertime experiment designed to test the heights given above resulted in a range of heights from 72 km at midday to 92 km at night, when adjusted for experimental errors. It was concluded that daytime scattering probably takes place from two different heights simultaneously. The upper stratum at 85–90 km is probably an enduring phenomenon, while the lower stratum, at about 75 km, is a daytime phenomenon, probably resulting from solar influence.

Three additional experiments have been concerned with finding the daytime scattering heights. In one of these, based on the observation of round-trip pulse delay times at two seasons, winter and fall, the median adjusted scattering height between the hours of about 0800 and 1700 was 87 km. The path used was one of 624 km from Sterling, Va. to Round Hill (South Dartmouth), Mass.

In the second experiment, round-trip delay times measured over the 1295-km path from Long Branch (Havana), Ill. to Boulder, Colo., yielded a median scattering height during the daytime of 85 km.

Analysis of antenna height gain observations over an extreme range path were interpreted in terms of effective scattering height, indicating heights of 85–90 km at night and lower heights during daytime. Seasonal variation was indicated, showing that the higher ionospheric heights especially prevailed during equinoctial seasons, at all hours, and that the lower heights were primarily a summer effect. Antenna height gain is important at low heights, but this effect becomes less pronounced when the antennas are elevated to normal heights. The questions of optimum antenna heights, and tolerances about the optimum, are dealt with in Section 4.

b) Distance Dependence: Ionospheric scatter propagation is most effective at ranges of the order of 1000 to 2000 km. At shorter distances, the received signal intensity is more and more attenuated by the increasing scattering angle. At greater distances, the extreme useful range is limited quite sharply by the earth's curvature. The ionospheric scattering height and effect of tropospheric bending are factors in distance dependence. Antenna height requirements increase with increasing path length; extreme range observations have been made to 2300 km for point-to-point systems, where natural elevations of greater than 2000 feet were available at the terminals.

As path length is increased, the portion of the scattering stratum visible in common from the transmitter and receiver approaches the horizon. As antenna heights are correspondingly elevated in order that the first ground reflection lobe can illuminate the common volume, multiple ground reflection lobes are formed. If the free space vertical beamwidth of the transmitting antenna is large compared to the vertical beamwidth of the ground reflection lobe, as is usually the case, power is wasted in radiation in the higher ground reflection lobes. Thus the total power incident on the common volume, *i.e.*, that integrated over the first ground reflection lobe, decreases with increasing path length. Calculations have been made of the field illuminating the scattering volume from elevated antennas as a function of distance, taking into account the effects of tropospheric refraction, divergence at the spherical reflection surface, and parallax. Integration over the scattering volume, using these calculations, has been carried out to interpolate between measurements made at different distances, with the results shown in Fig. 1-2. This curve gives distance dependence for ionospheric scatter transmission between ground stations at which the antennas are at optimum heights. The extreme range data were based on measurements of transmission loss made over a one-year period

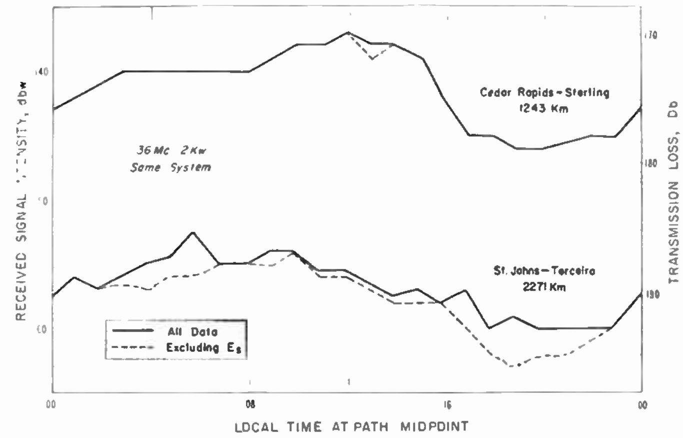


Fig. I-11—Comparison of signal intensities received at distances of 1243 and 2271 km.

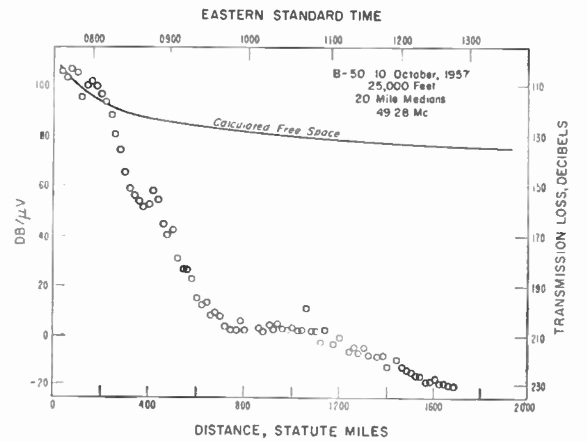


Fig. I-12—Aircraft measurements of ground transmissions at 50 mc showing line of sight, tropospheric, and ionospheric scatter signals.

at 36 mc over a 2270-km path from Newfoundland to the Azores. The same system was compared on the 1243-km path from Cedar Rapids to Sterling, as shown in Fig. I-11. Diurnal and seasonal variation of distance dependence are both observed, and these variations have been attributed to variations of scattering heights. The midday and summer maxima of signal intensity on the extreme range path are less evident than on shorter paths, suggesting that these maxima are associated with the lowest scattering heights which are cut off by the earth horizon from the scattering volume.

Observations of ground transmissions from high-altitude aircraft have indicated recordable signal levels to beyond 1500 miles, as shown in Fig. I-12.

c) Realized Gains of Antennas; Path Antenna Power Gain; Azimuth of Arrival: The term "realized gain" is used here to denote the ratio of the signal intensity observed with given transmitting and receiving antennas to that observed using simultaneously a reference antenna system, usually dipoles. Referred to two isotropic antennas, the realized gain of the system is the path antenna power gain, G_{pp}.

It has been observed that the realized gain of an an-

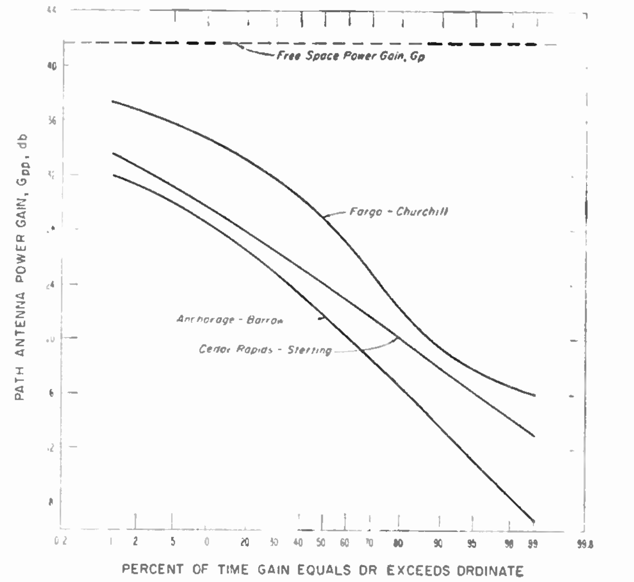


Fig. I-13—Realized antenna gain for narrow-beam (6°) rhombics aimed "on path." (Distributions of half-hourly medians).

tenna at one end of a path is dependent on the directivity of the antenna at the other terminal. A measured distribution of values of path antenna power gain G_{pp} are given in Fig. I-13 for narrow-beam rhombic antennas, with horizontal beamwidth of 6° between half-power points aimed along the great circle between transmitter and receiver points.

The path antenna power gain of antennas for scatter propagation depends on the angular extent and direction of the effective scattering volume in relation to the beamwidth and aiming of the antennas. For narrow-beam antennas aimed along the great circle, path antenna power gain tends to be greatest at midday. It is known that scattering centers (presumably mainly meteoric ionization) off path are more efficient than the path midpoint during many hours of the day, especially at night when signals are characteristically weak. Under these conditions, the narrow-beam antennas directed along the great circle yield substantially less than their maximum advantage over broad-beam antennas. Aiming the narrow-beam antennas off path in the direction of the effective scattering centers gives an advantage in average signal over the great-circle transmission. The subject of realized gain cannot be discussed adequately without emphasizing the effects of off-path transmission, as obtained by "split-beam" or "slewed-beam" transmission illustrated in Fig. I-14. An experiment was carried out using the beam arrangements shown in Fig. I-14(d); the median signal intensities recorded for the three path arrangements are shown in Fig. I-15 showing that for many hours of the day, off-path transmission gives several db less hourly median loss than on-path transmission. The fading is more rapid for the off-path transmission; further studies need to be made to specify very well any advantages to be obtained from off-path operation.

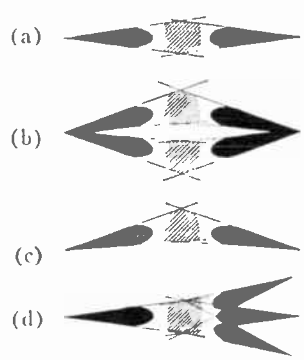


Fig. I-14—Beam arrangements. (a) "On-path" great-circle transmission. (b) "Split-beam" transmission. (c) "Off-path" or "beam-slewed" transmission. (d) Experimental arrangement to test geometrical arrangements of (a)-(c) using essentially common volume of ionosphere.

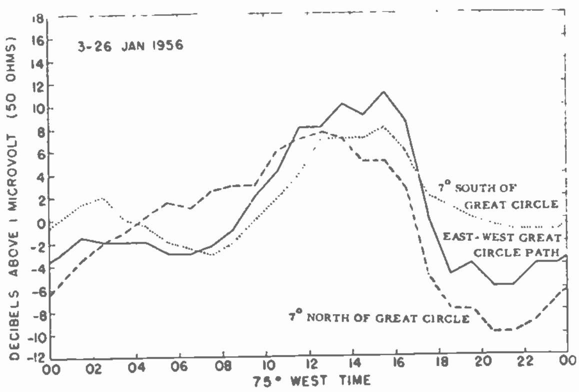


Fig. I-15—Diurnal variation of hourly median signal intensities at 50 mc observed simultaneously along the great-circle path from Cedar Rapids to Sterling and on two off-path receivers.

3.4 *World-Wide Variation of Signal Characteristics*

Experiments have been carried out using uniform recording techniques in various parts of the world to observe geographical dependence of signal characteristics, mainly transmission loss. The main data available were obtained at 50 mc using narrow-beam (6°) antennas aimed along the transmitter-receiver path, over paths of approximately 1300-km length. Paths at arctic, subarctic, temperate, and equatorial latitudes were studied in North and South America, and two additional paths in the European-Mediterranean area established that there were no gross longitude effects.

Except for the transequatorial path, the monthly median values of transmission loss for the paths observed were within 5 db of the median value for all paths; the levels exceeded for 95 per cent of the hours of the month were much more uniform, probably within ± 3 db of the central value.

Within these limits, the signal intensities on the arctic paths were statistically higher than those observed on the temperate latitude paths. The European-Mediterranean paths (England to Spain and Libya to Italy) showed statistically slightly lower signal intensities than the U. S. control path, but also within the limits cited above. The differences shown among the paths may in part result from the diverse path orientations, insofar as aspect sensitive meteor reflections and perhaps field-aligned irregularities may be involved in the total signal.

An IGY experiment conducted in South America in the area of the magnetic equator revealed that signal intensities there are much higher than in the U. S. The background scatter signal is of roughly the same magnitude, but the normal signal is obscured a large portion of the time by strong enhancements (as much as 50 db) which appear to be of the nature of sporadic E. Very rapid fading is often characteristic of the enhancements. There is some evidence that circuits operating outside the zone of the magnetic equator but still in tropical latitudes would encounter signals lower than at temperate latitudes.

It appears from the latitude variations that a geomagnetic influence is evident in the signal levels of scatter propagation.

3.5 *Other Geophysical Effects*

Meteoric and other sources of ionization in the D-region have been discussed in Section 2, and interpretation of some of the angle of arrival observations has been mentioned in Section 3.3 as regards the relative influence of meteoric and direct solar effects. This section deals with some observed relationships to geomagnetic and solar events and indices.

No very direct relationship of VHF transmission loss has been observed with indices of total ionization in the lower ionosphere such as f_0E (E-region critical frequency) or HF attenuation at temperate latitudes. Normally, such correlations are probably masked by meteoric effects and variation in the parameters of turbulence; abnormal events such as solar flares which enhance D-region ionization for short periods certainly show effects on the VHF signal. At arctic latitudes, more satisfying correlations are observed with geomagnetic indices and HF attenuation.

a) Flares and SID's: The behavior of the received signal on VHF scatter circuits during SID's varies with the frequency. The increased ionization created by the radiation from the flare appears to have two effects on the signal: intensity-enhancement and increased absorption. At lower frequencies, where absorption effects are strongest, the net result may be a marked decrease in signal level during an SID. At the higher frequencies the result is an increase of the signal level. At intermediate frequencies scarcely any net effect is seen in some instances; it may be supposed that the absorption and enhancement are nearly balanced. Fig. I-16 shows the differential effect during an SID on signals transmitted over a 1295-km path in the U. S.

On the other hand, on a parallel system which used broad-beam (65°) antennas for transmitting and receiving, attenuation was noted simultaneously at all frequencies up to 75 mc—the highest used.

b) Aurora and High-Latitude Absorption; Correlation with Indices of Ionospheric Absorption and Disturbance:

other observations, one would expect that the error rate begins to increase markedly near 1000 bits/second.

The short delay multipath limitations are imposed by strong off-path meteor reflections delayed by up to about 2 msec relative to the continuous background signal; the distribution of delay times depends on antenna beamwidth. Measurements using the broad-beam Yagi antennas showed an upper limit of about 150 bits/second keying speed during the worst periods.

2) *Long delayed multipath:* During times of high solar activity and high prevailing F2 layer reflecting frequencies, distant ground backscatter can be propagated by the F2 layer, arriving in the back or side lobes of the receiving antenna delayed tens of milliseconds after the D-region scatter signal arrival. In design of special modulation techniques to avoid the very destructive self-interference from this source, it is important to know the range of delays encountered. Observations at 30 and 40 mc are summarized in Fig. I-18, which shows delays typically from 12 to about 60 msec, and rarely as much as 80 msec. Special provision can be made in modulation systems for eliminating errors from this source, such as use of long pulses, frequency stepping or sweeping, or the use of more complex keying conditions than in usual two- or four-frequency FSK systems. Difficulties from this source can also be alleviated by operating at frequencies above the F2 MUF so that no reflection occurs, or by design of antennas for adequate suppression of radiation outside the main lobe (as illustrated by Fig. I-21).

c) *Comparison of Frequency and Amplitude (SSB) Modulation for Radiotelephone Service:* Commercially available frequency modulation and single-sideband equipment have been used for voice communication. Tandem transmissions of up to three hops have been simulated by retransmitting tape recordings of the received signals over the same path; actual tandem voice communication links have been used experimentally on the USAF North Atlantic system.

The frequency modulation equipment employed has used a deviation ratio of one. If the received signal is substantially greater than the noise, the theoretical performance of the FM system as compared to single-sideband is 1.76 db in favor of FM. (For transmitter average power for FM equal to the peak envelope power for single-sideband.)

Alternate transmissions of FM and single-sideband were made with test recordings at various times of the day and night. The received signals were tape recorded, and intelligibility tests were made later with a group of observers. The frequency modulation transmissions give more nearly "solid copy" at the higher signal-to-noise ratios than did single-sideband; however, at low levels the intelligibility is better with single-sideband. The single-sideband signal, though very intelligible at high signal levels, was not as pleasing to the ear because of the more pronounced effect of fading. This effect also showed up in the tandem runs as a more rapid deterioration of the single-sideband transmissions.

Voice tests using broad-beam antennas have not been reported. The more rapid fading and the poorer bandwidth correlation observed with broad-beam antennas imply greater signal distortion.

Under conditions of long delayed backscatter multipath observable at the lower frequencies, single-sideband shows a definite advantage; in fact, the FM system becomes unintelligible under the more severe multipath conditions.

3.7 Initial Operational Experiences

Fig. I-1 shows the North Atlantic VHF scatter system operated by the USAF, from its western terminal in Massachusetts, via Labrador, Greenland and Iceland, to England. The system was designed initially to provide four-channel time division multiplex teleprinter service, and has recently been equipped for 16-channel operation. The operational reliability of this system is illustrated for three of the links in Fig. I-19. Fig. I-19 illustrates the time lost on the Sondrestromfjord-Goose

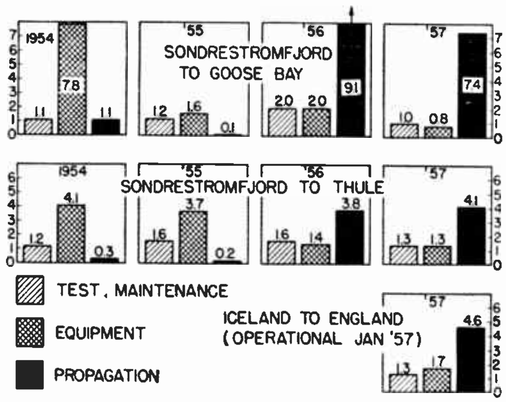


Fig. I-19—Lost time, North Atlantic VHF scatter circuits.

Bay and Sondrestromfjord-Thule paths for the years 1954 through 1957. It also shows the operating time lost on the Iceland to England path for the year 1957. The solid black blocks represent propagation outages, including weak signals, noise, and multipath. During the first couple of years, the lost time due to propagation was very little and equipment outage was dominant. The transmitters in use were early modifications of standard military high-frequency units. The time lost in 1956 and 1957 was actually mainly due to long delayed multipath, the effect of F2 propagation of ground backscatter from great distances, prior to use of any corrective measures for this difficulty.

4. Design of Ionospheric Scatter Communication Circuits

4.1 Introduction

Previous sections contain a summary of the basic propagation phenomena; here we briefly present a summary of those characteristics which have a distinct bearing upon VHF scatter communication circuit design.

a) Transmission Loss and Distance Range: The transmission loss expected to be exceeded by no more than 1 per cent of a year's hourly median values as measured with appropriately sited and elevated half-wave dipoles is, roughly, 200 db at 50 mc; using high-gain (20 db/isotropic) antennas, the value is approximately 190 db. This value varies only a few decibels over the distance range from 600 to 1200 miles if antennas are at appropriate height for the path length. The distance dependence function, based on ground measurements, is shown in Fig. 1-2 for ionospheric and tropospheric propagation. As distance decreases below about 600 miles, the scattering angle increases more rapidly with decreasing distance and the path loss correspondingly increases; at 500 miles, the ionospheric scatter mode loss can be perhaps 8 db greater than within the 800–1000-mile region of minimum loss; at ranges less than about 600 statute miles, adequately designed tropospheric systems can be more advantageous than ionospheric scatter systems. Beyond the 1100–1200-mile region, the signal strength decreases as the curvature of the earth begins to impose restrictions on the common scattering volume in the D-region and on the antenna terminals. Antenna height design is such as to aim the first ground reflection lobe at approximately 85-km ionospheric heights, as discussed later. The diurnal, seasonal, and geographical variations of transmission loss, and cumulative distributions of hourly values of signal-to-noise ratio for representative 50-mc installations were outlined in Section 3.

The working value of signal-to-noise ratio is generally estimated from such statistical descriptions gathered over long periods, indicating the expected year-round median path loss and signal-to-noise ratios, and the levels expected to be exceeded during 90 per cent, 99 per cent, (etc.) of the hours. The distributions depend upon antenna beams and orientation, specific site noise conditions, frequency, etc.; allowance for these and other factors must be made for any specific circuit design.

b) Frequency Dependence: Frequency dependence and its time variation were discussed in Section 3. As a practical matter, using scaled antennas, transmission loss may be taken to vary as $f^{7\frac{1}{2}}$.

In the absence of man-made disturbances, the limiting noise background at VHF is predominantly cosmic in origin, at least for mid- and high-latitude locations; the received noise power varies as approximately $f^{-5/2}$. Thus signal-to-noise ratio varies as f^{-5}, appreciably less than dependence of the signal power alone.

c) Fading: For design purposes, the amplitude fluctuations of the fading signal may be taken to be Rayleigh distributed; at 50 mc the power spectrum is concentrated below about $\frac{1}{2}$ to 5 cps, except for fading associated with aurora or reflections from meteor trails which may range from several tens of cycles to kilocycles (discussed in Section 3.1). Fading rate varies with frequency as $f^{3/4}$ to $f^{5/4}$. The fading rate and coherence-bandwidth depends also on path length and antenna beamwidths. For frequencies in the 30–60-mc region, a 1000-mile path length, and antenna half-power beam angles of some 10°, frequency selective fading can be noticed between two frequencies spaced by a few thousand cycles, and a correlation coefficient of 0.5 will generally be reached in the region of 5–10 kc, neglecting effects produced by "sputter" and Doppler components due to meteor trails.

d) Interference, Multipath and Noise: A previous section surveyed the various interference and multipath limitations.

In addition to the cosmic noise background at VHF, atmospheric noise is observed at times. It can originate either in electric storms in the vicinity of the receiving terminal or, within line-of-sight at greater distances, within the main antenna beam or major side lobes. When other short-time deleterious influences such as F-layer propagation, meteor- and auroral-induced multipath and signal level fluctuations are effectively eliminated (by appropriate frequency, antenna, modulation and diversity selection), and an adequate signal-to-cosmic-noise ratio is provided, atmospheric impulse noise remains as the most serious factor in maintaining circuit reliability. The occurrence of such noise is dependent upon electrical storm activity which, in turn, has pronounced geographical and seasonal variations. Relatively little atmospheric noise is expected in high latitudes, but precipitation static caused by charged snow or sleet particles striking a large antenna structure can prove troublesome.

4.2 System Design

a) Frequency Utilization: The most useful range of frequencies for ionospheric scatter applications is from 30 to 60 mc. Several considerations affect the choice of optimum frequency for a given service. Transmission loss increases with increasing frequency, so that the lower part of the band, *i.e.*, below about 40 mc, is preferred to minimize power requirements. At such lower frequencies, however, the frequent propagation of strong signals to great distances by the F2 layer greatly extends the interference range during years of maximum solar activity during seasons of highest F2 MUF; this calls for carefully planned and restricted frequency assignments, as at HF, or changing frequency during interference conditions to frequencies above the F2 MUF with consequent increased power requirements or reduced channel capacity compared to operation at the lower frequencies.

In *arctic latitudes*, polar disturbances sometimes cause intense absorption at frequencies in the lower VHF band. As a consequence of both the absorption effects and long-range interference possibilities, arctic circuits requiring the highest order of reliability must have a frequency available above about 45 mc. This can be used either occasionally as required, or for continuous service, sacrificing the power reduction realizable at lower frequencies. Single-frequency operation as low as 30 mc is suitable for services which can tolerate occasional interruption (a few times per year during years of maximum solar activity), providing frequency assignments can be made to avoid long-range interference and jamming is not a consideration. Modulation and/or antenna design for both telephone and telegraph services must take account of the long delayed multipath encountered at the lower frequencies.

At *temperate latitudes*, on the other hand, absorption has not been observed to be a serious factor even as low as 30 mc. Thus many services, especially civil services which do not normally take antijamming precautions, can operate continuously and reliably at a single frequency anywhere in the band. This is providing, of course, that suitable frequency assignment plans are made, as is required at HF, to avoid long-range mutual interference by F2 propagation; effective modulation techniques and antenna design can satisfactorily eliminate the multipath "self-interference" effects accompanying operation at frequencies below the F2 MUF. Amplitude modulation speech intelligibility is not seriously affected by this source of self-interference.

The conflicting considerations to be weighed, therefore, are the higher power requirements at the higher frequencies in the band vs the complexity and restrictions of frequency allocations to avoid long-range interference at frequencies below the F2 MUF.

At 50 mc, approximately 5 kw per 60-wpm teleprinter channel is required (without error correction) for a typical system (over a propagation path characteristic of continental U. S.), for satisfactory service 99 per cent of the hours in the year, for element-error rates of 1 per 10,000. Error detection techniques presently in use on some ionospheric scatter links can reduce the power requirement for a given error rate by a factor of 7 or 8, or alternatively, can greatly reduce the error rate for a given power. Intelligible voice transmission requires about 10 kw per channel at 30 mc for service 99 per cent of the hours. (At arctic latitudes, frequencies higher than about 45 mc must be used to assure continuity through disturbances.)

b) Terminal Equipment: Design of the terminal equipment is directed toward achieving reliability and efficiency. While such goals are, of course, of general importance for the communication engineer, they have a particular significance in the design of an ionospheric scatter circuit. The scatter mode offers the radio engineer his first opportunity of achieving reliable fixed-frequency radio communication over distances as great as some 1200 miles without the use of intermediate repeaters, even in regions subject to magnetic and auroral disturbances which can shut down HF circuits. In order to realize this inherent capability, however, a very high basic transmission loss must be overcome, and system design must take into account the special fading and multipath characteristics.

1) Modulation, detection, multiplexing: The path losses are such that relatively low values of signal-to-noise ratio must be accepted for reasons of economy of transmitter power. The modulation and detection method chosen must be near optimum in order to achieve the best possible reliability. Relative to the high initial costs of site buildings, roads, power plant, and RF equipment including antenna system, the cost of using the best modern modulation and detection techniques is comparatively minor, and the circuit capacity-reliability product can be improved greatly by the use of such techniques. As previously discussed, other factors influencing the choice of modulation-detection-multiplexing methods are: multipath delays encountered during auroral displays sometimes identifiable as "sputter," long-distance F-layer backscatter, Doppler shifts, and delay resulting from off-path propagation via ionized meteor trails.

The large path loss which must be overcome for reliable operation of a VHF ionospheric scatter circuit has, to date, generally restricted its use to multichannel teleprinter operation and single-channel voice operation.

The frequency shift keying (FSK) method of teleprinter operation has found favor in scatter applications over the keyed-carrier method for a number of reasons. It is inherently capable of an efficient detection and, with the use of good limiting circuits in the receiver, is less susceptible to the influence of intense fading. Also, the use of a sufficiently wide frequency shift obviates the difficulty caused by meteoric Doppler shifts. (Recently, detection schemes not requiring a wide shift to avoid the Doppler shift errors have been devised and tested, and show some promise.)

Multipath delays limit the minimum permissible individual element length which can be used. The most serious delays arise from F-layer backscatter where delay times of several tens of milliseconds have been observed; not even the slow keying speed associated with automatic 60-wpm teleprinter transmission is proof against such long delays. Generally, appropriate antenna design and siting, modulation techniques, and proper frequency selection can reduce or remove this difficulty.

In northern latitudes, especially in regions near the auroral belt, delays several milliseconds long can be caused by multipath propagation via auroral ionization regions. While antenna directivity will reduce the occurrence and length of delay due to this type of multipath, which sometimes manifests itself as fast-fading ("sputter"), a requirement for highly reliable teleprinter operation will generally place a limitation of perhaps a few milliseconds on the shortest length permissible.

Off-path reflections from meteoric ionization will also give rise to multipath delays of up to several milliseconds. Unlike the auroral multipath propagation conditions which occur only spasmodically over the year and are dependent upon path location, the meteoric multipath delays are generally observable throughout the year and are, to a first approximation, independent of path location and orientation. Experimental system observations using narrow-beam (6°) antennas have shown that these delays limit digital transmission speeds to about 500 bits/second at the worst times.

Finally, even in the absence of these forms of multipath, delays of as much as a few tens of microseconds would be expected because of scattering in depth from more than one stratum or from a thick stratum within the D-region.

In order to realize very low error rates therefore, even under conditions where F-layer interference is absent, a minimum element length must be used as protection against the persistent meteoric delay component. While a specific value for this minimum element length cannot be generally stated—it will depend upon the path length, antenna directivity, and frequency—a value of some 2–4 msec appears to be indicated during the worst conditions for a beamwidth of 6° to 12°.

Errors associated with Doppler-shifted meteor echoes must also be avoided. Experimentally observed distributions of Doppler shift were given in Section 3. At 50 mc, the maximum Doppler shift may be as great as 5 kc. The present technique employed to avoid errors from this source is to transmit a 6-kc shift; receivers use two distinct filters centered about the two carrier-shift positions each having a bandwidth sufficient only to pass the required pulse energy. A basic filtering limit is reached, for long pulse systems, in the short-time random carrier frequency fading spectrum introduced by the scatter propagation mechanism. (Recently, experimental trials have shown some promise for a technique devised to avoid Doppler-shift errors using a minimum frequency shift required by the keying speed.)

In addition to narrow-band filtering, the use of such nearly (theoretically) optimum modulation-detection techniques such as matched filter binary transmission have also found favor in order to yield the highest possible detection sensitivities for binary information transmission. The use of very narrow-band filtering and matched filter detection techniques, however, demand that commensurate attention be paid the requirements for very accurate frequency control for the former, and synchronization time control for the latter. High-stability oscillator-clocks should be used to provide the frequency and/or time stability required, and automatic group-phase correction should also be used.

The phase fluctuations in the medium are excessive for successful use of presently available phase-keying systems using 20-msec elements.

Error correcting codes may be used in the original message or feedback error correcting technique mav be used to achieve very low character-error rates, or alternatively, to reduce power requirements.

The usual requirement for the simultaneous transmission of several teleprinter channels requires that a decision be reached as to the optimum method of multiplexing the several channels upon a single carrier frequency.

For a relatively few channels, time division appears to possess certain advantages. Extensive multiplexing by frequency division should employ a technique for avoiding Doppler-shift errors which does not require excessive interchannel spacing, in order to avoid excessive total bandwidth requirements. The frequency division method requires that lower average power be assigned to each such carrier channel in order that peak power limitations not be exceeded, whereas a time division system may operate at peak power for all channels.

The limitation to the use of time division multiplexing systems is reached when the individual element length approaches the limit set by the various transmission multipath delays mentioned above. If we assume, roughly, a 20-msec element for a single 60-wpm teleprinter channel and a 2-msec minimum element length before the possibility of multipath error becomes large, about a dozen such channels appear to be a reasonable limit for effective use of time division multiplexing. Should more channels be desired, more than one such time division group of channels could be frequency multiplexed upon the carrier frequency, with the subcarriers for each time division group being sufficiently separated to eliminate Doppler difficulties. Alternatively, multiple frequency shift keying may be employed with an increase of available average power per channel where there is a peak power limitation.

Where a fixed circuit capacity system is not mandatory, a variable capacity time divison multiplex system offers a distinct advantage; by monitoring the prevailing multipath and S/N conditions, the element length can be adjusted approximately and the maximum usable capacity can be made available at all times. This, of course, require that the receiver bandwidth be adjusted in step with the keying speed, and may in practice further require some en route "storage" of messages.

The over-all transmission bandwidth required is somewhat larger than that usually employed in HF. As many as a half dozen frequency multiplexed bands of time division multiplexed teletype channels can be accommodated in a bandwidth of less than 100 kc.

The ionospheric scatter mode at present appears to be fundamentally most suitable for moderate capacity binary information transmission, *i.e.*, teleprinter or data, but circuits can be designed to provide a useful voice channel, using high power (approximately 10 kw at 30 mc and 90 kw at 50 mc for operation 99 per cent of the hours of the year at temperate latitudes). Present techniques cannot provide (with reasonable power) a voice circuit consistently exhibiting so-called "commercial" quality. However, for many applications an acceptably intelligible speech service can be provided.

Of the various modulation methods experimented with to date for voice transmission on a single circuit, single-sideband modulation might tentatively be adjudged best at the lower frequencies where long delayed multipath is encountered. Narrow-band phase modulation is at least as effective at the higher frequencies where only signal-to-noise limitations (no long delayed multipath) are encountered and is definitely superior for tandem links. Lack of phase correlation across the band limits the optimum deviation to about ±3 kc at 50 mc; this bandwidth limitation is approximately inversely proportional to carrier frequency. At the lowest level of signal-to-noise ratio, the smaller bandwidth requirements of SSB and the absence of a detection threshold for low S/N appear to give it an advantage in intelligibility in a noisy channel. In addition, the meteoric Doppler-shift whistles are somewhat less noticeable on SSB than on FM or PM. The carrier frequency stability normally provided on a scatter circuit for other reasons (see above) is more than sufficient for single-sideband voice operation. The average modulation level may be raised by the use of a compandor or by amplitude limiting; this can result in improved intelligiblity for a given peak modulation power level. In tandem links, the effect of noise addition from link to link is less for FM signals above threshold, and for systems involving three links the advantage of FM is marked. The problem of AGC does not arise with FM. Compandors are also effective in the audio channel of FM systems.

The transmission quality of a voice circuit, however, would normally be expected to be limited for an appreciable fraction of the time, especially on repeater circuits, unless a transmitter power of 50–100 kw (at 50 mc) and a high-gain antenna are employed. In addition, distortion of analog waveforms will accompany auroral multipath; in auroral regions, the distortion associated with "sputter" conditions might occur some 50 hours/year. Comparative studies of SSB vs FM have not been made under such severe distortion conditions, but it is very likely SSB would remain the more intelligible.

Fig. I-20 shows typical spectrum occupancy for ionospheric scatter transmission. Fig. I-20(a) shows a SSB voice in the channel with FSK teleprinter transmission; Fig. I-20(b) shows FM voice instead of SSB. Fig. I-20(c) shows a four-frequency FSK usage, employed on some USAF circuits; this also can represent a conventional "twinplex" arrangement, or four keying condition FSK system.

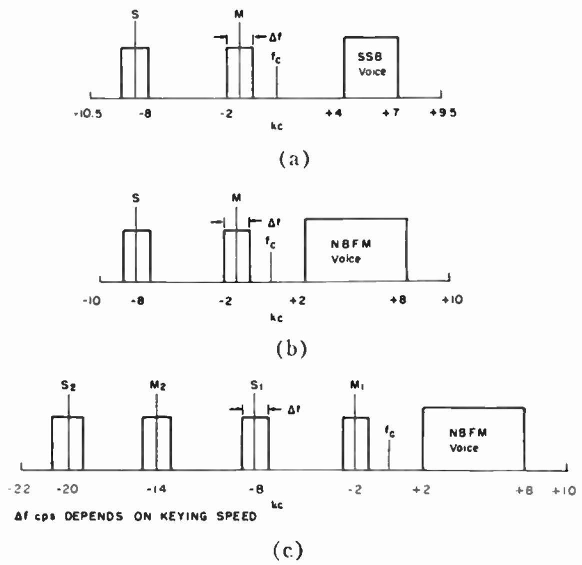


Fig. I-20—Typical present spectrum occupancy for ionospheric scatter transmission.

Space diversity is usually employed and is discussed in Section 4.2-c with antennas. Polarization diversity is effective for ionospheric scatter only if both polarizations are transmitted, as with frequency diversity. Some applications of frequency diversity can be very effective.

2) *Receivers:* The above section has indicated broad design objectives for modulation and detection techniques applicable to circuit design. Next are discussed more specific features of transmitter and receiver design essential for FSK teletype and SSB voice transmission on VHF scatter circuits.

The minimum detection bandwidth should be used in order that the detection sensitivity be as high as possible. In the limit, this bandwidth is determined only by the sum of the bandwidth required for optimum pulse detection and the fading power spectrum. The important components of the fading spectrum should require an over-all bandwidth of the order of a few cps at 50 mc; use of a bandwidth of a few cps should be adequate for all but very rare occurrences—"sputter" multipath, for instance.

Short-time instabilities in the generation of the carrier frequency of any subcarrier frequencies should, therefore, be restricted to a few cps at 50 mc. A carrier stability figure of a few parts in 100 million/day can be achieved fairly easily with present techniques. Special receiver front-end filtering and other efforts to eliminate receiver responses at frequencies other than that desired will also be of assistance in the protection of the sensitive ionospheric scatter receivers from outside interference. This subject deserves considerably more attention than is usually given conventional communication receivers, because of the problem of receiving a weak desired signal in the presence of possibly strong interfering signals at other frequencies.

The receiver noise figure should be kept down to a few decibels even though cosmic noise levels are substantially higher than this (perhaps by 6–12 db depending upon path location, orientation, and the antenna beamwidth) most of the time. Abnormal D-layer absorption (associated with SID or polar blackout phenomena) at times can cause a reduction in the signal level especially at the lower frequencies discussed here. The cosmic noise level will also be lowered simultaneously by this absorption and, at least for the smaller absorption values, the resulting S/N will not decrease as rapidly as the signal decrease alone, provided the receiver noise figure is low enough not to be limiting.

The gain and time constant characteristics of the receiver should be such as to permit consistently reliable operation in the presence of intense and, at times, rapid variations in signal level. During sporadic-E propagation conditions, for instance, signal levels can increase more than 50 db above that normally experienced with pure scatter propagation, and meteor bursts 20–40 db above the continuous background signal level will often occur.

If a voice capability is desired, the AGC requirements must be carefully considered for amplitude modulation systems because of the rather rapid and intense short-time signal level fluctuation. Adequate solution of this problem as far as the SSB system is concerned is closely connected with the question of diversity operation. Increasing the number of effective diversity antenna/receivers can greatly reduce the intensity of such fluctuations, but maintaining the proper noise levels and gains among a large number of sensitive receivers can be difficult. This intense short-time fading, which can exhibit marked frequency selective characteristics over a bandwidth of a few kilocycles, presents a particularly interesting problem in attempting to obtain satisfactory AGC for SSB voice operation.

3) *Transmitters:* Amplification over the bandwidth must be characterized by good linearity. This is required in order to minimize the cross-talk between any frequency-multiplexed teletype channels or to minimize distortion for SSB voice operation.

Attention must be paid to the reduction of spurious and harmonic signals. Considering both the high power levels and large antennas employed, even signal levels as much as 40 to 60 db below full carrier output can constitute a potentially serious interference hazard to the other services sharing this portion of the radio spectrum. Particular care must be taken to minimize spurious signals and noise at the frequencies lying near or within the pass band(s) of return-path and repeater receivers located close to the transmitter. Carrier frequencies should be separated by several mc for duplex service, especially in those cases where a single antenna is to be used both for transmitting and receiving ("duplex" operation).

A design technique favored in the search for good long-term transmission reliability on both ionospheric and tropospheric scatter circuits is that of using two or more transmitters in parallel. During regular maintenance periods or breakdown of one of the individual transmitters, the total power output will be reduced by only a few db, and no outage time will normally result from a switchover.

High efficiency is important in the final stage(s) of the transmitter, especially for those remote transmitter locations where the cost of transporting the power generator fuel is a large part of the total fuel cost. For the same reason, the power output level should be easily adjustable to permit its reduction during prolonged periods of relatively low path loss.

The receivers used in a space diversity array must have a sufficiently high order of frequency stability to match the transmitter performance, and this frequency stability must be maintained for two widely separated and readily available frequencies if dual-frequency operating concept referred to earlier is employed.

4) *Frequency changing arrangements:* Two-frequency operation may be used (as discussed in Section 4.1) to avoid long-range F2 propagation, or absorption blackout in arctic regions, by use of the higher frequency at required time, and to reduce power requirement normally by use of the lower frequency. During years and/or seasons of generally low MUF for a particular circuit, a frequency near 30–35 mc could be employed, thereby taking advantage of the associated relatively low path loss; at those times when the MUF is expected to approach or exceed the frequency used, a change could be made to a frequency in the 45–60-mc region. It is not expected that such changes in operating frequency would take place many times even over a period of time as long as a year, but the transmitter design would have to anticipate operation in two frequency regions separated by nearly an octave and would have to provide means whereby such a large change could be made not only promptly and simply, but without appreciable losses being introduced by those RF components usually considered to be frequency sensitive.

5) *Spare equipment:* As in the case of many tropospheric scatter and line-of-sight UHF and SHF circuits, the employment of adequate spare equipment and "running backup" methods can result in a large increase in circuit operating time, and parallel operation of several of the major equipment items is in order when a long-term reliability in excess of 95 per cent is required.

Finally, any requirements for multiple repeater operation must be given very careful attention in the early over-all circuit design studies.

c) Antennas: Perhaps the outstanding characteristic of VHF ionospheric scatter propagation with which the communications engineer must contend is the inherently large path loss. Even with modest information capacities, this loss requires the use not only of large transmitter powers and sensitive receivers, but also the employment of antennas capable of efficient illumination of the appropriate scattering region(s) and the collection of the scattered energy. Interference potential, fading rates, minimum usable pulse lengths, and correlation bandwidth are also established largely by the antenna beamwidth and direction of aiming. To date this has generally meant the employment of antennas with plane wave gains approximately 20 db relative to half-wave dipoles at the same height; in the lower portion of the VHF region, this requires large structures.

The design of these large antennas reflects the engineer's concern with regard to multipath and interference. A very small main-lobe beamwidth will confine the major fraction of the radiated power to a small volume and will minimize the relative incidence and magnitude of meteoric and auroral off-path delays. For these reasons, and to reduce the vulnerability of the sensitive receiving system to long delayed backscatter and/or interference during sporadic-E and F-layer propagation conditions, very low side and back lobe levels are advantageous. It is desirable that these low side and back lobe levels be maintained independent of the polarization of incident fields. These latter characteristics have been more readily achieved with reflector types of antenna designs and are reasons why such designs are presently preferred over the less expensive rhombic design used initially.

The transmitting antenna must also handle average transmitter powers of up to perhaps 100 kw without arcovers that could prove disastrous in duplex operation, permit dual frequency operation if necessary, and be of sufficient mechanical strength to maintain its electrical characteristics even under the extreme conditions of wind, snow, and ice encountered in some locations.

d) Azimuthal Beamwidth and Direction: For services requiring the maximum transmission bandwidth and maximum resistance to interference and multipath, narrow-beam high-gain antennas are appropriate. Azimuthal beamwidths of 6° to 12° are used. The pattern of a corner reflector antenna designed for ionospheric scatter service is shown in Fig. I-21; the suppression of response outside the main lobe was designed to be at least 40 db by grading the currents in the driven elements to minimize, in the lower VHF band, the effect of interference and long delayed backscatter. (At higher frequencies, say above 45 mc, the requirements are much less critical for responses outside the main lobe.)

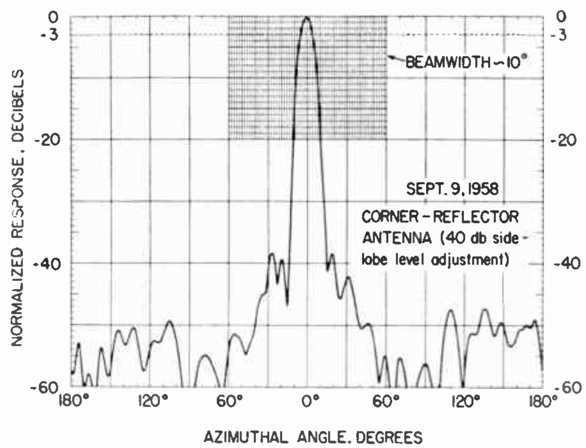


Fig. I-21—Measured pattern of corner reflector antenna designed for 40 db suppression of radiation outside main lobe.

An advantage can be obtained during most of the day by slewing antenna beams a few degrees to one side or the other of the great circle path between transmitter and receiver, to enhance the contribution of aspect sensitive meteoric ionization as discussed in Section 2. Though this technique is not yet in regular operational practice, the experimental studies mentioned in Section 3 (illustrated in Figs. I-14 and I-15) have established the potential advantages. It is especially important that transmission losses can be reduced by this technique during the hours of the day when weakest signals are encountered. Corner reflector antennas have been designed for beam slewing at least 10° off path which maintain the low side lobe levels as comparable to Fig. I-21. Further studies need to be made of optimum angles for beam slewing as a function of diurnal and seasonal variation of meteoric influx, and of error rates for off-path transmission as compared to on-path transmission.

It should not be inferred that only narrow-beam antennas are useful. They are appropriate for services designed for maximum transmission capacity and efficiency. There are applications, however, where broad-beam antennas can be very useful; more elements can be combined in diversity to achieve signal-to-noise performance comparable to the larger antenna installations providing the keying speeds (pulse lengths) are limited in accordance with the multipath limitations of the greater scattering volume. For typical 5-element Yagi antenna systems (60° horizontal beams) the minimum usable pulse length is about 6 msec; for a time division multiplex system having four 60-wpm channels, four such receiving antennas in diversity provide about the same binary error rate as a high-gain narrow-beam system using two receiving antennas in diversity. Fading rates are much more rapid with the broad-beam system, but this is not necessarily a disadvantage with adequate diversity combining. The more severe selective fading (poorer bandwidth correlation) observed on the broad-beam systems would undoubtedly introduce more distortion in voice channels than is characteristic of the narrow-beam systems, but no direct comparative observations have been reported.

1) Beam elevation; antenna height; siting: The hoped-for results of excellent antenna design and construction can be seriously compromised by a poor site selection, since radiation characteristics in the lower VHF region are markedly influenced by terrain features both near the antenna site and within the radio horizon along the propagation path. A sufficient expanse of level ground used to provide good lowest lobe radiation pattern formation in the direction of the appropriate scattering volume. These measures should insure the most effective "coupling" with the scattering volume and the realization of as much of the image antenna plane wave gain as is possible in scatter field reception.

Since the radiation pattern formed using horizontal polarization is not influenced to a very great extent by variations in the local earth (or sea) constants, its use appears preferable where constancy of radiation angle and intensity is desired over the seasonal cycle.

A corollary to the observed failure to realize, for a large fraction of the time, antenna gains comparable to the plane wave gains is that the additional gain resulting from ground reflection will not be fully realized, even though ideal sites are employed at the transmitter and receiver. Nevertheless there are demonstrable advantages in siting the antennas with respect to a ground surface so that the plane wave ground-reflection lobe patterns are well formed. In fact, if either antenna is poorly sited, the beams may only partially intersect in the height region where maximum scattering occurs.

The optimum antenna height is such that the lobes from the transmitting and receiving antenna intersect throughout the scattering volume in a way which maximizes power transfer to the receiver. This means that the lobes are aligned slightly short of the path midpoint, symmetrically. In computing patterns in the lower ionosphere for elevated antennas over spherical earth, allowance must be made for tropospheric bending, spherical divergence, near-horizon diffraction, and parallax effects associated with the finite distance to the scattering volume. Fig. I-22 illustrates optimum antenna height requirements as a function of distance at 50 mc.

An earlier published curve (Ref. 3) for antenna height, which has been used in practice, is shown for comparison with a more recently computed curve for optimum height. The figure also shows the "lobe alignment height," i.e., the height required to aim the axis of the first ground reflection lobe at an 85 km height at path midpoint; the "lobe alignment equivalent" height indicates the tolerance for height compromise applicable to antennas of narrow azimuthal beamwidth. The height of narrow beam antennas may be reduced to values indicated by this curve with no loss relative to the "lobe alignment" height. The optimum height gives a gain of a decibel or so over "lobe alignment" height depending on a number of factors; the optimum height curve applies for antennas having a broad free space beamwidth relative to the beamwidth of the ground reflection lobe. For very narrow free space vertical beamwidths the "lobe alignment" height more nearly applies as optimum.

In siting an antenna intended to function effectively for small angles of arrival and departure, it is insufficient merely to provide a suitable site on which to perform the construction. Nor is it sufficient simply to have an unobstructed horizon in the desired azimuth for the desired angle of departure or arrival. When the angle of departure or arrival is small, the ground for a considerable distance in front of the antenna plays a critical role in formation of the lobe pattern. The problem of ground-reflection lobes and general requirements for a smooth first Fresnel zone has been given considerable study, particularly with respect to the ground radar siting problem.

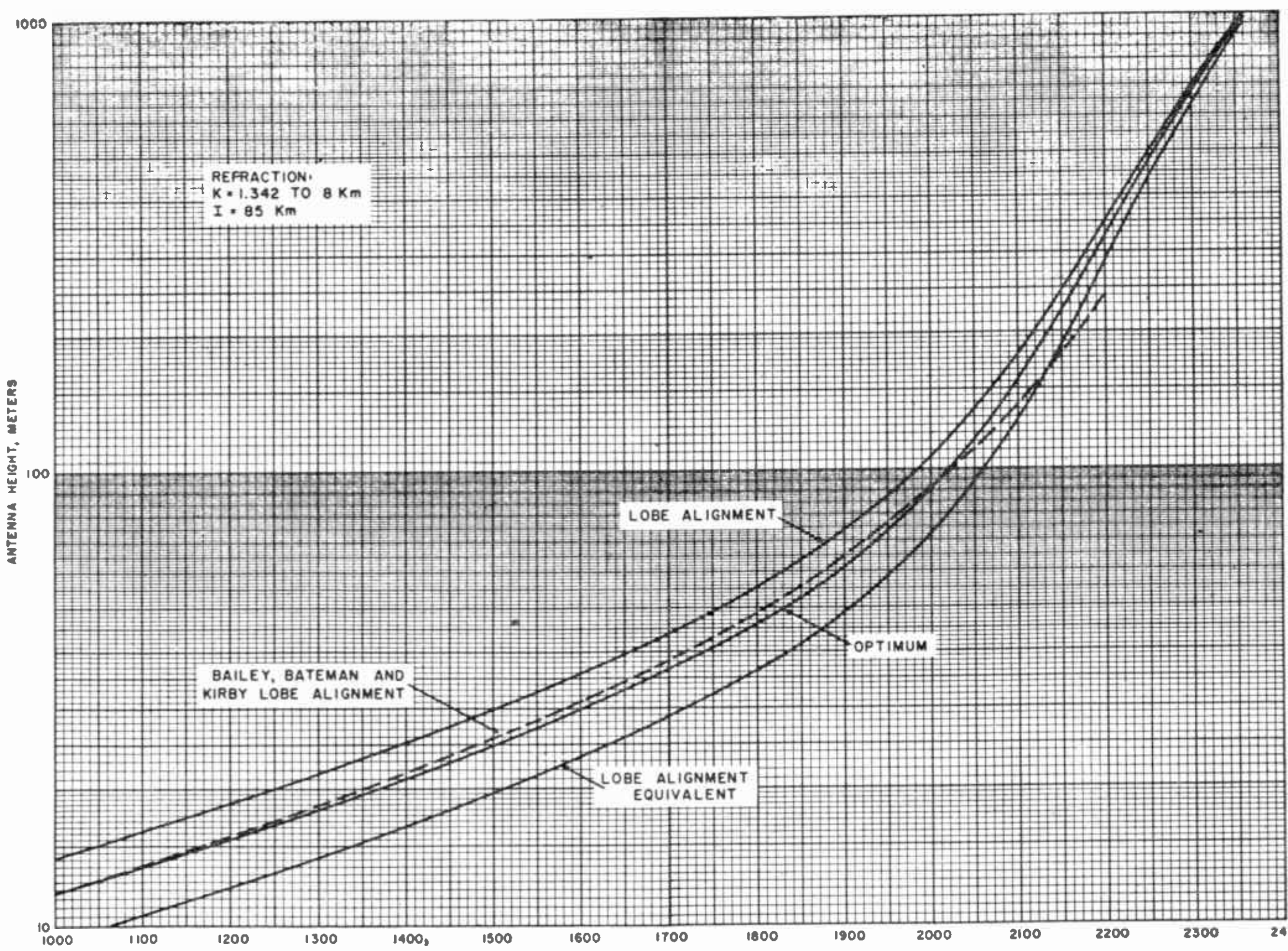


Fig. I-22—Lobe alignment, optimum, and lobe alignment equivalent antenna heights, 50 mc

Variations in the tropospheric refractive index can have an important influence upon received signal level for very low radiation angles (1° or less). Sufficient horizon clearance should be provided for the conditions of substandard refractive index gradient likely to occur in the colder and drier seasons, rather than simply the average conditions expected over the year. For most middle- and low-latitude paths, especially those located near the edge of large bodies of water, the possible influence of superrefraction and discrete layers must be carefully considered for very low angles of radiation.

In siting antennas, various aspects of interference must be given specific and careful attention. Employment of the very high effective radiated power needed to provide a large capacity-reliability product on an ionospheric scatter circuit can give rise to large line-of-sight, diffraction, and tropospheric "scatter" fields out to rather great distances from the transmitter. The interference potential of these fields can be reduced by locating high-power RF terminal and repeater equipment in relatively remote locations and, where possible, siting the antennas so that screening angles are presented to azimuth sections other than the main beam direction.

2) *Diversity:* Effective diversity is of vital importance on a scatter circuit. Of the various methods which may be employed to secure the benefits of diversity operation, multiple antenna arrays spaced normal to the propagation path have been the most favored. Space diversity is preferred to frequency diversity, both in the interests of spectrum conservation and in order to realize the additional gain derived from not having to split the transmitter power among the channels used. Relatively small lateral spacing is required, since a correlation coefficient of even as high as 0.4 will yield nearly the full diversity advantage available with zero correlation. Such a value can usually be obtained, especially during the weaker signal periods, and if moderate antenna azimuth beamwidths are employed, with an antenna spacing of about 5 wavelengths; the exact value depends also upon the distance.

While early circuits employed dual diversity with spacings of 10 wavelengths or greater, consideration is now being given to the use of larger numbers of closer spaced diversity antennas. FSK system using a 6-kc shift also permit use of frequency diversity between mark and space channels.

3) *Other:* Use of more sophisticated antenna systems such as split or multiple beam antennas, or of antennas capable of scanning and selecting the optimum azimuth arrival angle have been suggested, and further work needs to be done to develop the practical advantages of such techniques.

5. Fields of Application

Compared with cable, microwave, or tropospheric relay systems, long-distance radio communication by the ionosphere is at best a poor way of meeting modern trunk communication requirements. But where cable or relay terminals are unavailable, or vulnerable as in military systems, there is a really genuine requirement for long-range radio services. The emphasis in trunk service is on high capacity for data transmission and multichannel telephone service. But there is a need for many services of more modest capacity, such as command channels, air traffic control channels, and many civil telecommunication links where four to sixteen channels of teleprinter and/or a single voice channel are representative requirements.

A well defined role is emerging for ionospheric scatter as simply the most reliable radio technique available for the fixed services in the range beyond tropospheric propagation, *i.e.*, from about 700 to about 1200 miles. Furthermore, though its exploitation is confined to the lower VHF band where frequency allocation problems are difficult, it offers the possibility of some measure of relief from the vastly greater congestion of frequency assignments in the HF portion of the spectrum. The utility of ionospheric scatter transmission is ordinarily limited to the fixed service.

Most of the applications of D-scatter communication to date have been military ones. The impact which the VHF mode has already had upon radio communication in the Arctic and North Atlantic has been considerable. The North Atlantic Treaty Organization has recently extended a network from Norway through central Europe to Turkey. Inter-island civil air communication requirements might be well served by its use. At least one civil telecommunications application is known: an experimental one-way link has been operated between The Netherlands and Italy, and it is understood that plans are underway to expand this service to full duplex operation.

Further studies of the propagation characteristics and improvement of applicable communication techniques can be expected to lead to lower cost and more efficient systems, as well as to new useful applications. Ultimately, the utility of this, as well as other long-range radio techniques, may have to be measured against that promised by line-of-sight microwave links using earth-satellite relay.

6. Factors Affecting Frequency Allocation

6.1 *Frequency Bands of Interest*

The most useful range of frequencies for ionospheric scatter applications is from 30 to 60 mc. Section 3.2 discussed frequency dependence of the transmission loss and signal-to-noise ratio, and Section 4.2 discussed factors affecting choice of operating frequency.

In summary, it appears that ionospheric scatter services would find it advantageous to develop their frequency utilization in three categories:

1) Two frequency operations using 30 to 45 and 45 to 60 mc ranges. Actual choice of frequencies within these ranges depends on geographical location. For

example, near the continental United States, the higher frequency would most usefully be near 50 mc. This category would comprise services requiring the highest obtainable reliability which are designed for minimum power requirement for normal operation, but can use higher power, or reduced transmission capacity, at the higher frequency when needed to assure continuity through the occasional periods of intense absorption, or through periods of interference by long-range F2 propagation. (IRAC has considered a policy which would require the scatter services to operate above the F2 MUF so as to avoid interference. If the higher frequency of the pair is above the values given by Fig. 1-23, the F2 interference potential will be conservatively limited to less than 1 per cent of the hours of the worst season during years of highest solar activity.)

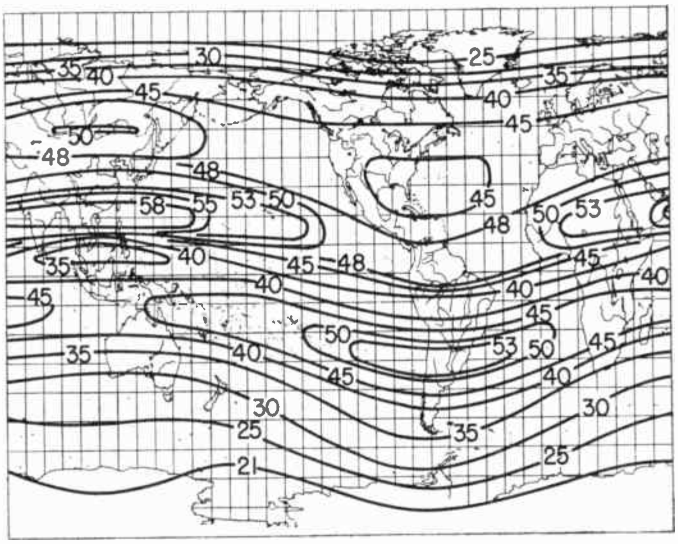


Fig. 1-23—F2 4000 MUF exceeded during 1 per cent of hours December solstice; sunspot maximum No. (150).

2) Single-frequency services above about 45 mc, comprised of services requiring highest reliability, but which are of limited channel capacity or designed for the high power required. Antennas and modulation techniques can be simpler than for single-frequency operation below 45 mc.

3) Single-frequency services in the lower part of the band (say 30 to 45 mc) can obtain at temperate latitudes reliable operation at these frequencies, using for telegraph and facsimile services special modulation techniques and/or antennas designed to suppress self-interference from F2 propagated ground-scattered multipath echoes. Intelligibility of amplitude modulated speech is not seriously affected by this multipath interference and requires no special provisions. Of course, continuous single-frequency operation at these lower frequencies implies the possibility of long-range interference by F2 propagation, as at HF; allocations and assignments would likewise have to be internationally coordinated and protected.

The propagation modes most significant in long-distance interference between scatter services and other services are sporadic E and F2. To avoid sporadic E interference, ionospheric scatter services should have their transmitting and receiving terminals geographically separated from other circuits or services by at least 1800 miles. Fig. 1-23 shows a contour map of F2 maximum reflecting frequencies exceeded for 1 per cent of the time during the winter season of highest MUF's at sunspot maximum. A circle of 2000 km radius centered on the scatter station gives the frequencies at which interfering propagation over 4000-km paths can occur 1 per cent of the time. The percentage of time is smaller for paths longer or shorter than 4000 km.

6.2 Bandwidth Requirements

Ionospheric scatter transmission has been used mainly for multichannel teleprinter service, and, to a limited extent, single-channel voice service, using such baseband arrangements as are illustrated in Fig. 1-20. The factors affecting channeling arrangements and required bandwidth were discussed in Section 4.2.

It appears that 40-kc blocks would provide adequately for the transmission capacity of most ionosphere scatter services using present techniques—smaller bands would provide for many services using only teleprinter transmission, the actual width depending on number of channels and the multiplexing arrangements. Future developments may be expected to reduce the necessary bandwidth or to provide for expansion of transmission capacity within the band.

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II. Long-Range Tropospheric Transmission*

1. Introduction and Summary of Principal Features

A CONSIDERABLE amount of experimental evidence has been collected in the past ten years indicating that dependable radio transmission over tropospheric paths extending far beyond the radio horizon can be obtained at frequencies in the VHF, UHF, and SHF ranges. Although the propagation characteristics of such paths are not yet fully understood, it has been found that entire radio communication systems can be reliably established with repeater spacings which are many times greater than the line-of-sight distances ordinarily employed at these frequencies. While various names have been applied to this type of radiowave propagation, it is perhaps most generally known as "tropospheric scatter." Its discovery opens up new possibilities for the use of radio, particularly in the field of fixed point-to-point communications, and can be expected to have considerable impact on the problem of spectrum utilization and frequency allocations.

When VHF and higher frequencies were first put to use, it was generally believed that propagation to distances beyond the horizon was governed essentially by the laws of diffraction over a smooth spherical earth surmounted by a stable homogeneous atmosphere. The smooth earth theory predicts an exponential decrease in signal strength at a rapid rate at and beyond the radio horizon. The useful range of these frequencies, therefore, was believed to be limited to distances not much greater than line of sight. However, as more powerful transmitters, higher-gain antennas, and more sensitive receivers were developed, an increasing amount of evidence was obtained that signal levels well beyond the horizon were much higher than diffraction theory predicts, and these signals exhibited considerable fading. At first, these "anomalous" beyond-horizon fields were attributed to the influence of such sporadic tropospheric radio-meterological situations as superrefraction, guidance in ducts, or reflection from elevated layers, and to the influence of the ionosphere at lower VHF, but as the data accumulated it became apparent that the transmission effects were much more persistent than such meteorological and ionospheric conditions permitted. Accordingly, numerous experimental investigations were initiated after World War II—first to determine the characteristics of these peculiar fields, and then to explore their usefulness in a new type of transmission.

The continuing experimental investigations have been accompanied by a considerable amount of theoretical work. At this writing, the generally accepted theories ascribe the propagation, apart from D-region ionospheric scattering, to the effects of small-scale irregularities in the dielectric constant of the atmosphere, brought about by atmospheric stratification and turbulence. The existence of such irregularities is readily confirmed by observations with the microwave refractometer. These irregularities or "reflectors" are usually large compared to the radio wavelength, but their sizes and relative motions vary in a random manner and over large ranges. The effect of such irregularities in the medium is consistently to scatter or reflect small, but significant, amounts of power out of the main beam of the radio transmitter, some in the direction of the radio receiver. The output of the receiver represents the integrated effect of signal components arriving from the assemblage of all appropriately oriented scatterers or reflectors.

There is not yet complete agreement among the various theoretical models proposed to explain long-distance tropospheric propagation, and further experimental and theoretical radio-meteorological studies of this very complex problem are in progress. However, on the basis of the most generally accepted scattering theory, it now appears possible to make quantitative predictions of many of the principal propagation features. In addition, as a result of the great amount of experimental work conducted over the past ten years, a substantial amount of empirical data are now available to provide a basis for circuit and systems engineering. In the past several years, a number of communication systems, both military and experimental, have been put in operation using frequencies ranging from a few hundred up to about 5000 mc. The circuits are characterized by the use of transmitter powers up to 50 kw, free-space plane wave antenna gains as high as 40–45 db, and space-diversity reception. Circuit parameters of this order are necessary because the losses on typical tropospheric scatter paths, although much smaller than the values given by smooth earth diffraction theory, nevertheless are 50 to 100 db, or more, greater than the corresponding free-space values.

In the present stage of the art, single-hop circuits can be designed to carry 10 or 12 circuits over distances of some 500 miles, the maximum range depending on the grade of performance desired. Increased circuit capacities, of course, entail reductions in the permissible link length. Long multirepeater systems with capacities of 100 or more telephone circuits are practicable with repeater spacings of the order of 100 to 150 miles, with quality and reliability entirely adequate for many military applications. The capacity, reliability, and quality required of long, multilink, commercial service circuits, the greater cost control often required in their installation and operation, and detailed considerations of

* Original manuscript received by the IRE, July 16, 1959.

traffic patterns may, however, require a somewhat more conservative view of their commercial attractiveness. Low-capacity high-reliability air-to-ground communications can be designed for 200–300 miles beyond the radio horizon. In spite of the inherent multipath nature of the propagation medium, there is evidence that reasonably satisfactory transmission characteristics can be obtained for standard black and white television signals at distances of 150–200 miles, when appropriate antenna and diversity techniques are employed.

2. Review of Tropospheric Propagation Theories

2.1 General

The propagation of VHF, UHF, and SHF radio waves over paths within the radio horizon is now quite well understood, and the performance of line-of-sight circuits can therefore usually be predicted with reasonable quantitative accuracy. A great deal of study has been directed toward the influence of atmospheric refractive index changes upon propagation characteristics. The proper placement of antennas under all anticipated propagation conditions is mandatory in order to hold the path loss to an acceptable minimum for the required operating time. Also, variable multipath conditions can produce rapid and deep short-time fading, and place a limitation on the propagation medium flat "bandwidth." Radiowave energy is also diffracted into shadow regions beyond surface obstacles on the bulge of the earth. The free-space loss law is a geometrical one; the smooth sphere diffraction is an exponential one. As a result, the diffracted component weakens rapidly with distance beyond the line of sight, especially at the higher frequencies.

2.2 "Anomalous" or Sporadic Long-Distance Field Strengths

At distances well beyond the line of sight, field strengths are often measured which appreciably exceed the value calculated upon the basis of a standard atmosphere and the application of rigorous diffraction theory. These stronger fields, which occur from time to time, can usually be explained on the basis of departures of the tropospheric refractive index height profile or peculiar path geometry conditions.

A change in the slope of the refractive index profile can result in superrefractive conditions in which the average ray curvature is increased and the effective radio horizon is extended. Marked inversions in the profile can permit energy to be propagated well into the shadow region, either by reflection from the resulting layer or, particularly at the shorter wavelengths, by being guided in an atmospheric or surface based duct. Also, the path loss between antennas on either side of a sharp ridge or mountain can be much less than that usually expected under such conditions if the antennas are suitably located to take advantage of multiple-ray knife-edge diffraction.

The ionosphere can also make its influence known at VHF and even UHF. Energy can be propagated to very great distances via sporadic E clouds, turbulent layers, or meteor-induced ionization columns located near midpath at altitudes of 70–110 km; via auroral ionization from great heights and, at times, locations away from midpath; and via normal F-layer propagation (including ground backscatter) at lower VHF during abnormal solar conditions.

2.3 "Persistent" Long-Distance Field Strengths

Experiments conducted at distances well beyond the normal radio horizon, primarily during the years from 1935 to 1950, fostered the suspicion that even when all due allowance had been made for the influence of the sporadic mechanisms outlined above, there remained a weak, but remarkably persistent, field-strength component. Such a conclusion was based upon a number of data sources from different parts of the world, over various distances and covering some two orders of magnitude of frequency variation (40–4000 mc), all of which indicated that the average fields far beyond the radio horizon were much higher than could be explained on the basis of a simple application of diffraction theory.

The basic diffraction theory assumes a smooth spherical earth and a homogeneous atmosphere. Attempts were made to reconcile the measured field-strength values with those calculated on the two further elaborations of earth surface roughness and variations in the slope of the atmospheric refractive index profile. The results were not encouraging.

In the early 1950's, these persistent long distance fields came to be called "scatter" fields. The rapid and intense short-time fading characteristic of at least the lower average field-strength values naturally brings to mind the concept of a multiple-source scattering propagation mechanism. Other possible propagation mechanisms have been advanced by theoreticians since that time, but the word "scatter" has continued to find favor as a generally descriptive term to describe the phenomenon of persistent long-distance VHF-UHF-SHF tropospheric propagation.

2.4 Outline of the Major Theories

a) Turbulent Scattering Theory: The theory which has been given the most attention, and has seen the widest acceptance to this time, has been that which hypothesizes the weak scattering of radio waves by turbulent departures from the mean tropospheric refractive index. This theory assumes the existence of refractive-index time and spatial fluctuations within a common scattering volume defined by the solid-angle intercepts of transmitting and receiving antenna radiation patterns. By defining the spatial autocorrelation function in terms of a characteristic scale of turbulence, equations can be derived which express the power scattered forward and downward out of the transmitting beam into the receiving beam (see Fig. II-1). Available direct airborne measurements of the gross features of the refractive-index variations in temperate latitudes indicate that they are characterized by a spectrum of scales extending at least over the range from a few centimeters to several kilometers, and that the intensity decreases, on the average, exponentially with height.

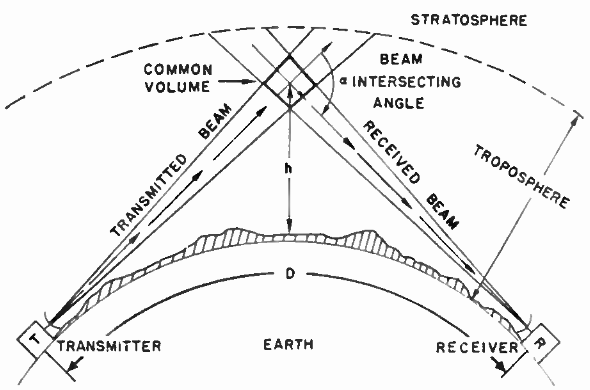

Fig. 11-1—Path geometry of a beyond-the-horizon UHF circuit.

In order to make quantitative predictions of scatter propagation characteristics, the theorist must concern himself with the spectrum of the turbulent variations. Two basic theories have been advanced to date: one, the Obukhov mixing theory, assumes fluctuations being injected into the spectrum at very large "blob" sizes; the other, the mixing-in-gradient theory, suggests that turbulent convection transfers water-vapor agglomerations between various heights, thus continuously changing the small-scale refractive index. Of the two, the latter theory appears to be somewhat the more favored, based upon a comparison of the observed-to-predicted scatter-field frequency dependence.

Adopting the mixing-in-gradient hypothesis, the theory of tropospheric scatter has recently been extended to account for the fields measured at distances beyond some 300–400 miles. At these greater distances, the common scattering volume normally would have to be located in the lower stratosphere. Fluctuations in water-vapor content cannot contribute significantly to the refractive index fluctuations at such heights, and only the dry atmosphere density variations are called upon to explain the weaker longer distance fields. To make quantitative predictions of such long distance propagation characteristics as the average path-loss dependence upon distance and the diurnal, seasonal, climatological, and frequency variation thereof, general and detailed knowledge of the lower atmosphere is required. For instance, the variation of average path loss with distance depends crucially upon the rate at which the intensity of the refractive index fluctuations decays with height. After careful normalization of such factors as frequency, antenna aperture-to-medium coupling loss, surface refractive index, atmospheric gaseous absorption and path height, and surface distance and angular distance, good agreement can be obtained between the observed and calculated median transmission loss out to distances of some 600 statute miles, when the assumption is made that the intensity of the turbulence decreases exponentially with height. Here, good agreement is defined as a 6-db RMS deviation of the long-term medium winter fields from prediction. A stratospheric model (as mentioned above) has been called upon to explain the increased rate of tropospheric radiowave attenuation measured at UHF beyond some 500–600 miles, as compared to the 100–300 mile region. Agreement between the calculated and measured absolute path losses in the 600–800-mile region can be attained with the use of a stratospheric scattering model, but there is no indication of a marked increase in fading intensity at the troposphere-stratosphere interface region near 400 miles which this concept also predicts.

The mixing-in-gradient theory predicts a first-power wavelength dependence of the scattering cross section and this appears to yield a better agreement than the Obhukov ($\lambda^{-1/3}$) with that observed on the average for UHF at intermediate distances. There is experimental evidence, however, that there is a variation in the wavelength dependence about this average.

A generalization of the basic scatter theory to admit of anisotropic turbulence has permitted a better agreement between calculated and measured values of the aperture-to-medium coupling loss observed with narrow beamwidth UHF-SHF antennas at intermediate distances. Some recent measurements could be interpreted as indicating that this loss may be frequency and distance dependent.

The multipath-limited propagation medium bandwidth predicted by various workers appears to depend upon the definition of useful bandwidth, the detailed characteristics of the scattering model, and the instantaneous signal level. For antenna radiation patterns large with respect to the cone of scattered rays, the prediction is of some 4–30 mc at 100 miles, with this value expected to decrease with the cube of the distance. The lower of these values would appear to be more correct when compared with measured values. Ratios of aperture to wavelength sufficiently large to produce radiation patterns smaller than the cone of scattering angles, of course, will reduce the multipath delay and increase the useful bandwidth correspondingly.

b) Turbulent Layer Scattering Theory: Airborne refractive index measurements have demonstrated that distinct refractive index stratification is present in the atmosphere for at least half the time in many locations. Airborne and shipborne field-strength measurements of field strength vs distance also very often give evidence of field-strength enhancements which can be (geometrically) related to layers. Accordingly, the basic scattering theory has been modified to predict the power expected to be scattered beyond the horizon from the turbulence believed associated with such layers.

Calculations based upon this theory yield a turbulent layer scatter component which may be as much as twice as great as the layerless calculation. The hypothesis of several elevated layers near the same height may be called upon to provide a sufficient power addition to agree with the field-strength enhancements measured some 100–200 miles beyond the line of sight.

c) Layer Reflections Theory: Another theory of long distance UHF propagation has been advanced which bears certain resemblances to the scattering theory. This theory holds that the cause of the fields is reflection from the many omnipresent elevated layers, formed by sharp gradients of refractive index, which are distributed randomly in position and orientation. No multiple reflections are considered, nor are the reflections considered to be correlated. While the two cases of layer dimension, very large and very small relative to the wavelength, are mentioned, the attention of the theorist is directed toward a model in which these two dimensions are commensurate.

This theory predicts a path loss which varies as the cube of the frequency; *i.e.*, in agreement with the mixing-in-gradient scatter theory and in agreement with the average measured dependence. It predicts a distance decay in absolute power loss of D^6 to D^7, depending upon the ratio of the beamwidth to the scattering angle. This is a somewhat more gentle decay than that measured on the average, and, in any event, does not predict the increased rate of path loss observed in the 500–600-mile region. The reflection theory also predicts an antenna-to-medium coupling loss which is in good agreement with the earlier measurements of this effect, but no predictions of propagation medium bandwidth have yet appeared.

It is interesting to note that the recent extensions of the early isotropic "blob" scattering theory to accommodate anisotropic, *i.e.*, elongated, and even elliptical, blobs suggests that the two theories may be drawing together, at least in respect to the fundamental physical picture of the irregular dielectric constant distribution in the lower atmosphere.

d) Normal Mode Theory: Another theory has been advanced which is fundamentally different from those discussed. It assumes that, because of the gradual decrease of the absolute value of the atmospheric refractive index with height, a small fraction of radiowave energy incident upon the atmosphere is coherently reflected downward from all heights. The work of the theorist here has required the establishing of descriptions of the average refractive index profile such as to meet the two-fold requirement of agreement with physical measurement over a large range of heights and mathematical tractability in the solution of the many, required, complex-normal mode equations and their eventual appropriate summation. Bilinear, trilinear, and inverse square descriptions of the profile have been used.

These different profiles all predict an absolute value of path loss some 50 db greater than free space for a 150-mile path between low antennas, which is not in great disagreement with the average value measured under these circumstances, and an exponential decay law. This latter prediction does not agree too well with the measured rate of decay over very long distances.

To date, this theory has not explicitly yielded predictions of a frequency dependence. Also, it is agreed that the effects of macroscopic turbulence must be introduced in order to explain other characteristics of these long distance fields.

2.5 Summary

The body of theoretical work directed toward the comprehensive quantitative description of beyond-the-horizon tropospheric propagation continues to grow. It has served the very useful purpose of directing the attention of the radioscientist to many very difficult and complex problems in the fields of small- and large-scale radio meteorology, the solution of which require obtaining new kinds of physical data concerning the detailed dynamic structure of the earth's lower atmosphere.

To this time the work of the theorist has been hampered to a very considerable extent both by the relative paucity of these data and, until lately, by not having available to him the results of careful and critical scientific radio-wave propagation experiments. As a result, theoretical developments have only recently advanced to the point where quantitative estimations can be made with sufficient exactitude and generality to be of great assistance to the communications engineer. The communications engineer must still rely upon his judgment in extrapolating, with the help of theoretical-empirical tools developed in the last year or two, what is by now a very substantial volume of statistical information to meet particular circuit or system design problems.

3. Experimental Observations of Propagation Characteristics

3.1 Techniques and Scope of Experimental Observations

In the last ten years, an extensive amount of experimental information has been obtained on the characteristics of radio waves that are weakly propagated by the troposphere well beyond the horizon. These experimental measurements have been accomplished in the VHF, UHF, and SHF portions of the frequency spectrum, and range from 100 to 800 statute miles beyond the radio horizon. They have been made possible largely by the high-powered transmitters and very large antennas which were developed during this period.

Much of the earlier data in the VHF band were derived from measurements of signals emitted by omnidirectional FM and TV stations. More recently, the potential application of this propagation phenomenon to useful long-distance communications systems required radio-wave propagation research programs which utilized high-powered UHF and SHF transmitters, together with large, highly directive antenna systems and sensitive receiving systems. Measurements have been made between many land sites, and between land sites and ships and aircraft. These programs have been designed, in general, to obtain data pertinent to the design of reliable communications systems and to investigate certain areas in which the existing theoretical models were either incomplete or inconsistent for accurate design purposes. Although the experimental results are far from complete in many details, considerable data have been obtained which are useful for the evaluation of: a) average attenuation rate vs range; b) variability of the signal levels as a function of time, distance, season, and climatology; c) frequency dependence; d) effective gain as a function of antenna aperture to wavelength ratio; and e) influence of the propagation medium on effective bandwidth and modulation techniques.

The summary of the experimental signal level data in this section is presented with reference to the free-space standard level for the typical sizes of antennas employed in the various experimental systems. The initial large loss incurred near and just beyond the horizon has made the use of large directive antennas mandatory for experimental measurements over the longer paths, particularly in the UHF and SHF range. As discussed later in this section, the ratio of effective gain to plane-wave gain of these larger antenna systems is not invariant with distance, frequency, and temporal changes in propagation conditions. Consequently, these data may include an increased loss caused by any reduction in effective gain of typical antennas; no attempt has been made here to separate these relatively small losses from the very much larger losses caused by the basic tropospheric propagation mechanism at long distances.

Since the signals propagated beyond the horizon fade rapidly, the signal levels are usually expressed in a statistical fashion. It is common practice to obtain the cumulative distribution of the varying signal amplitude over an hourly period and to state the signal level in terms of an hourly median value. This level represents a level equaled or exceeded for 50 per cent of the hourly period. The statistical distribution of the signal level is approximately a Rayleigh distribution for periods of a few minutes in the absence of duct propagation or aircraft multipath, and tends to approach a log normal distribution for hourly periods, at least for levels at and below the long time median. The hourly median values are used throughout the data discussed in this section as a standard measure for the signal level as a function of distance, season, frequency, and antenna aperture.

The surface distance between transmitter and receiver terminals is often an ambiguous basis for comparison of the data of signal level vs distance for the various experiments reported in the current literature because of the disparity in antenna heights. For the conditions of relatively smooth terrain and absence of local obstructions in the foreground of the antenna terminals, the various experimental data indicate good agreement when compared on a basis of the "effective distance" or "angular distance" beyond the horizon. The "effective distance" is determined by the distance between the points of tangency of the rays drawn from the centers of the transmitting and receiving antennas to the radio horizon. Over a rough earth the angular distance is the more suitable parameter; this may be computed by extending the great circle horizon rays until they intersect and computing the acute angle formed by these two ray paths.

3.2 Variation of Signal Level With Distance

The range of distances over which tropospheric circuits may be feasible is indicated by Figs. II-2 and II-3. They show the median signal level compared to free space as a function of distance. The data basic to both Figs. II-2 and Fig. II-3 were derived from the results of several representative and independent experimental studies. Those year-round data from which the curves in Fig. II-2 are derived were normalized for antenna heights of 50 feet above a smooth spherical earth; when

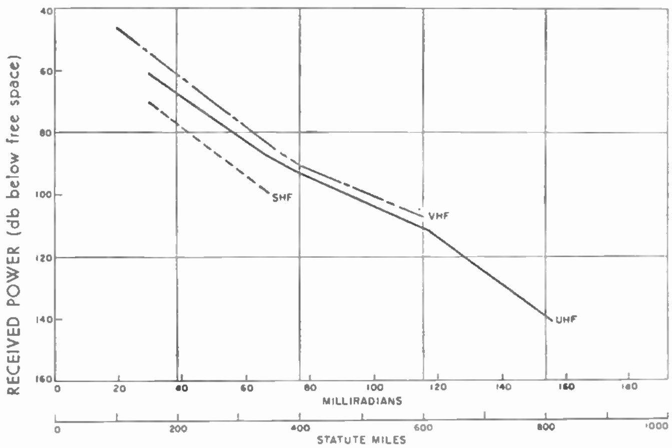

Fig. II-2—Variation of median signal level with distance; transmitting and receiving antennas 50 feet above ground.

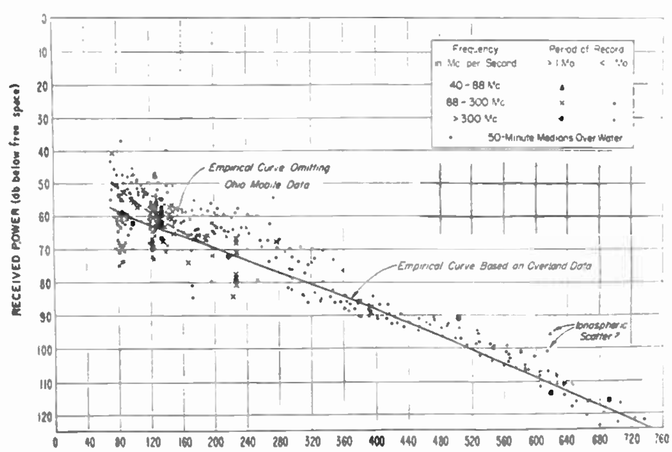

Fig. II-3—Variation of median signal level with distance for winter afternoon conditions.

normalized in this manner they are all in generally good agreement both in absolute signal level and relative trend, and clearly show the influence of frequency. The wintertime data in Fig. II-3 have not been normalized; of particular note is that they include points obtained on 80 randomly selected paths in the 100-mile region.

It should be emphasized that, in order to apply the data shown in Figs. II-2 and II-3 to the calculations of a particular tropospheric path loss, the attenuation encountered under free-space conditions must be added to the plotted values. Including this free-space loss, a median attenuation rate of about 20 db/100 statute miles is indicated for the 100–300 mile region. Curves fitted to the year-round data of Fig. II-2 indicate a decrease in this rate to 11–12 db/100 miles in the next 200–300 miles, followed by an increase again beyond 500–600 miles. The smooth curve in Fig. II-3 shows a good average fit to the winter afternoon data throughout the entire distance region beyond 200 miles, and indicates a total loss of some 11 db/100 miles in the 200–400 mile region, increasing to 12 db/100 miles beyond this.

3.3 Seasonal and Climatological Influence

It is well known that the variations in propagation by the troposphere to distances near and beyond the radio horizon are related to meteorological and climatological factors. The general trend of signal level as a function of climatological area may be inferred from observations made in arctic, temperate and subtropical areas. The yearly median value of signal for circuits operating at UHF in each of these areas plotted vs the annual mean temperature observed at weather stations near these circuits are shown in Fig. II-4. These data are representative of the signal level observed on paths 150–200 miles in length. There is some evidence that the variation of median signal level with annual mean temperature is decreased for paths many hundreds of miles in length.

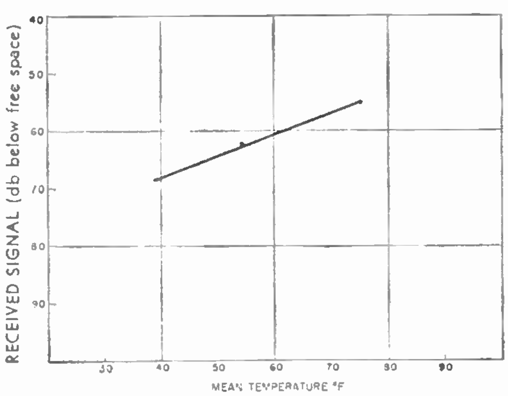

Fig. II-4—Variation of yearly median signal level with annual mean temperature of the path midpoint.

Since there is a marked seasonal variation in the monthly medians of observed signal level for a given path, use of data observed in the weakest periods is indicated for circuit-design criteria. Conversely, a knowledge of the expected signal in the strongest periods is necessary in determining the potential interference from a transhorizon circuit. The seasonal variability of median values is less pronounced as the length of the circuit is increased. The curves shown in Fig. II-5 indicate the seasonal variation of the monthly median of signal level for 188-mile and 618-mile paths, respectively. A difference of about 12 db between the maximum and minimum median values is indicated for the 188-mile circuit, while this difference is only 6 db for the 618-mile circuit.

The variability of hourly median signal levels within a month also decreases with distance. Distribution of hourly median values for summer and winter months at distances of 188 and 618 miles are shown in Fig. II-6. The difference between the 10- and 90-percentile values in decibels is shown in Table I.

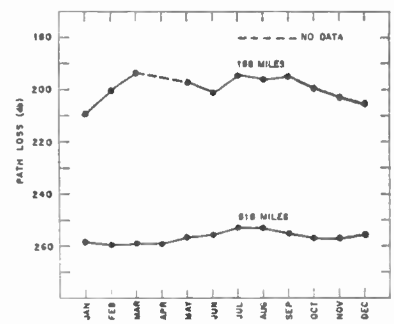

Fig. II-5—Seasonal variation of 400-mc data.

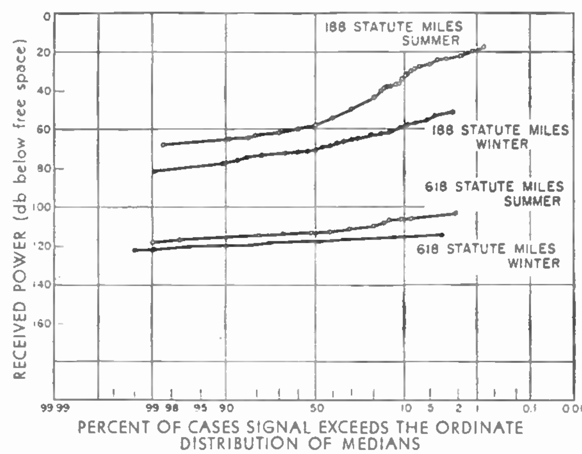

Fig. II-6—Distribution of hourly median values of received signal at 400 mc.

TABLE I

Distance	188 miles	618 miles
Winter	12 db	3 db
Summer	32 db	10 db

3.4 Frequency Dependence

A comparison of measurements made in the VHF, UHF, and SHF bands indicates a definite trend of decrease of signal level with increasing frequency. Precise determination of the frequency dependence is made difficult by the inability to account for the effects of the various beamwidths of the antennas employed throughout the various bands of frequency. Recent experiments have been completed which were designed to measure frequency dependence of the received signal level at 417 and 2290 mc. These experiments employed scaled antennas having equal beamwidths of approximately 5° at each frequency, and were operated simultaneously at the same height over a 200-mile path during a winter-to-summer period. In addition to the relative loss of 15 db at 2290 mc because of the λ^2 dependence in the effective absorption cross sections, an additional median loss of 7 db was measured at the higher frequency; this corresponds closely to a signal strength variation proportional to the first power of the wavelength and a path loss proportional to the cube of the frequency.

3.5 Performance of Large Aperture Antennas

The gain of a receiving antenna is usually defined as the ratio of the power received by the antenna from a plane wave incident upon the antenna at the angle of its maximum response to the power that would be received by an isotropic antenna from the same incident plane wave. In a completely general definition, the gain of an antenna varies as some function of the angular direction of incident plane wave. Therefore, any conventional antenna system will be subject to a "gain-loss," or an aperture-to-medium coupling loss, for plane-wave components arriving over directions other than the angle corresponding to the maximum gain as it is usually defined.

It is generally accepted that the mechanism of tropospheric propagation well beyond the horizon produces at a receiving antenna, in effect, an angular spectrum of equivalent plane components with random phase relations constantly changing in time. If the angular width of this spectrum of incoming radiation extends over a greater solid angle than that subtended by the receiving antenna pattern, the power collected will not increase as a linear function of the effective area of the antenna system.

An analogous condition will occur for the transmitting antenna. Under such conditions, an increase in effective area of the transmitting antenna will fail to decrease the path loss through the troposphere proportionally, since the narrower beam reduces the illumination of some portions of the intervening troposphere capable of scattering or reflecting radio waves toward the receiving antenna.

Measurements of the gain performance have been made for various sizes of antenna apertures at 2290 mc over paths 188 and 350 miles in length. These experiments included measurements of various combinations of antennas ranging in size from small pyramidal horns to parabolic reflectors 28 feet in diameter. The results of these experiments are shown in Fig. II-7 for the two

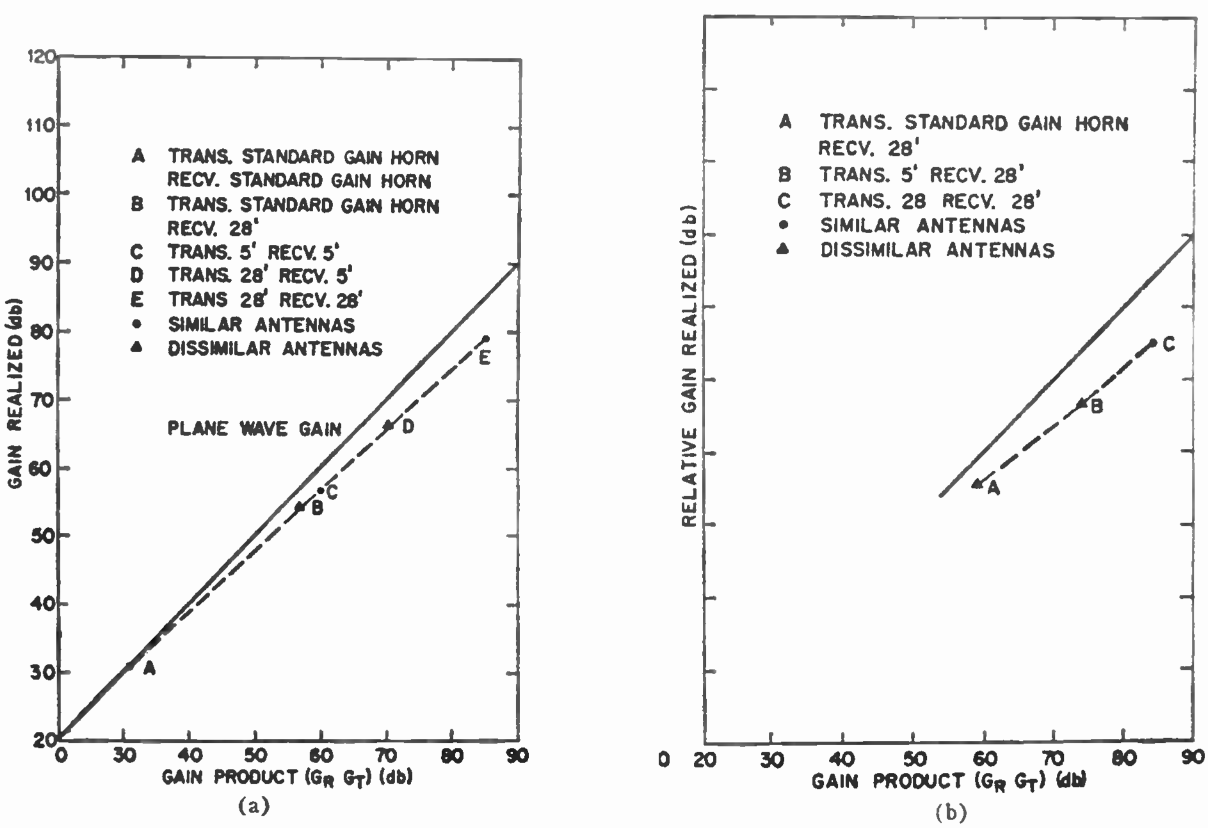

Fig. II-7—(a) Gains realized over 188-mile path at 2290 mc. (b) Gains realized over 350-mile path at 2290 mc.

distances. A progressive deterioration in realized gain with increasing distance is evident.

Comparative measurements over a 618-mile path at 413 mc using a small corner reflector and 60-foot diameter paraboloidal antennas, indicate that the total plane-wave gain of 71 db was approximately attained. However, these results are not consistent with the trend observed on 2290 mc at shorter distances, and suggest that the gain performance of very large aperture antennas is frequency dependent.

3.6 Fading

Long-range tropospheric transmissions are characterized by rapid fading. The distribution of the amplitude usually follows a Rayleigh distribution for periods of a few minutes, especially during periods of generally lower median signal level, tending to approach a log-normal distribution for periods of an hour. During enhancements caused by superrefraction, ducting, or layer reflection, these long period distributions do not follow a log-normal distribution at the levels where the signal exceeds the median.

Scaled beam experiments have indicated the effect of frequency on the fading rate of the received signals. Distributions of the rate of fading at 417 and 2290 mc are shown in II-8 indicating the median difference to be 9

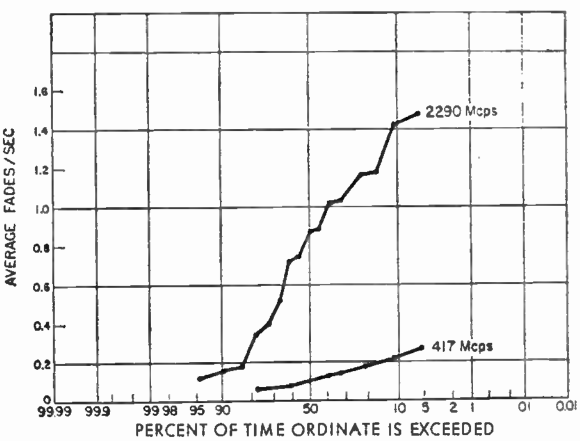

Fig. II-8—Observed fading rate at 417 mc and 2290 mc using equal 5° beam antennas.

to 1. Other experiments on lower frequencies not using scaled antennas have indicated a dependence nearer to the first power of the frequency. Power density spectra obtained from transforms of the autocorrelation function provide a more rigorous determination of the energy spectra of the fading signals. Experimentally-determined spectra indicate that the energy is contained in a band within a few cycles and that the spectrum width is not directly proportional to the operating frequency.

For some modulation techniques it is necessary to have information relating to the length of fades as well as the rate of fading caused by the mechanism. Analysis has indicated the length of fade at UHF is about 1 second at 6 db below the median, while fade durations of about 0.1 second were observed at SHF.

In addition to the fading characteristics of the tropospheric propagation mechanism, aircraft reflections can introduce rapid, intense, and relatively regular fading. This fading is caused by the beating between the Doppler shifted signal received directly from the moving aircraft and the background signal from the troposphere. The fading rate increases with transmitted frequency for a given aircraft velocity and orientation in the path. In practice, these fades may range from a few tens of cycles per second at UHF to hundreds of cycles per second at SHF; they can be very deep when the signal reflected from the aircraft is approximately equal in amplitude to the normal tropospheric signal.

Measurements made on a 188-mile path at 3670 mc indicate that some slight depolarization occurs in the troposphere; the cross-polarization component was, however, at least 12 db below the parallel polarization.

3.7 Bandwidth Limitations

The term "bandwidth," which has an explicit meaning in circuit applications to communications engineers, has been somewhat indiscriminately applied to the description of the communications capacity of the propagation mechanism of tropospheric scatter. In circuit applications, the transfer function $H(\omega)$ is usually complex and involves both the phase and amplitude as a function of frequency. The unambiguous meaning of these terms for physical circuits is based upon the tacit assumption that the parameters of resistance, capacitance, and inductance of such circuits to not vary in time.

There is general agreement that radio waves received well beyond the horizon are propagated by a tropospheric mechanism which produces, in effect, a multiplicity of equivalent plane-wave components arriving at the receiving antenna with random phase relationships and angular directions, both of which vary rapidly in time. A simplified picture of the nature of the received signal is that of the sum of a number of contributions arriving over paths slightly longer than the earliest arriving component in the great circle plane containing the transmitting and receiving antennas. These incremental delays are, however, many wavelengths at frequencies used for tropospheric scatter circuits.

An indication of the instantaneous transmission response has been experimentally determined by time delay measurements with very short pulses and frequency sweep techniques. In the former, multipath delays of a few tenths of a microsecond have been observed over paths 100 miles in length, while delays of up to 2 μsec have been observed over a 618-mile path. The antenna pattern beamwidth is the most important factor influencing multipath delays for beamwidths smaller than 1°, approximately; tropospheric irregularities predominate when beamwidths larger than this are used.

The results of measurements of the instantaneous transmission characteristics at 2290 mc obtained by sweeping the carrier over a 16-mc band is shown in Fig. II-9. The correlation coefficient between signals received on the center frequency and at frequencies offset from the carrier were determined. These measurements showed that at a separation of 4 mc the signals were almost completely uncorrelated over a 188-mile path with 1° antenna beamwidths.

In areas of dense aircraft traffic, multipath delays are caused by reflections from aircraft. In this case, the maximum differential delay is limited only by the geometry of the common volume of the antenna beams. This volume is determined by the antenna beamwidths and the length of the circuit; characteristic volumes could be expected to permit multipath delays of 0.1–10 μsec. Fig. II-10 has been prepared assuming that reflections can occur from aircraft locatd at points that are 3 db down on the antenna beams. Measurements of pulse distortion on circuits of various lengths and antenna beamwidths agree substantially with the values indicated in Fig. II-10. Reduction in multipath delays may be accomplished with the use of antennas having beamwidths of less than 1°. Such antennas would view the arriving signals only from within this restricted angle, thus eliminating signals arriving over wider angles.

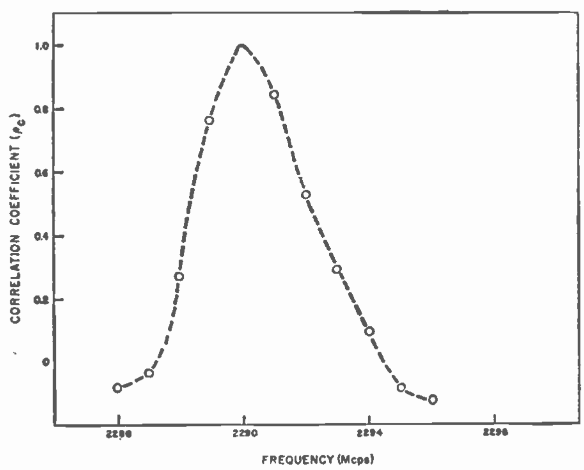

Fig. II-9—Correlation of amplitude as a function of frequency spacing.

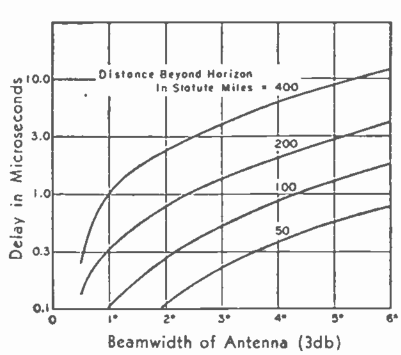

Fig. II-10—Multipath delays caused by reflection from aircraft in antenna beams.

3.8 Diversity

The deleterious effect on communications of the wide fading range characteristic of a long-range tropospheric signal can be alleviated with the use of diversity techniques. Diversity action is the selection or combination of two or more independently fading signals. The effectiveness of this diversity action is dependent upon the number of channels of contributing signals, referred to as order, and the correlation existing between the contributing signals.

There are various ways of achieving diversity action such as spaced antennas, frequency, time, and angular diversity. Early experimental results indicated that the correlation of signals received on antennas spaced 100 λ or greater was low. This low correlation predicted achievement of diversity gain for spaced antennas, and this estimate has been realized in practice.

Experimental measurements of the effectiveness of dual angular diversity indicate that the improvement of the diversity signal over that received on an individual channel approaches the improvement predicted by theory.

The frequency sweep measurements shown in Fig. II-9 indicate the possibility of a frequency diversity system. For instance, transmissions separated by 4 mc or more will fade essentially independently of each other in the 2300 mc region, thereby permitting effective frequency diversity.

3.9 Interference from Natural and Man-Made Sources

In the UHF bands, cosmic and solar noise can be of the same order as that achieved in very low-noise receivers. This is especially true in cases where the main lobe or side lobes are so oriented that they favor reception from extra terrestrial noise sources, especially during periods of high solar activity.

Observations of lightning strokes in receiving systems have also been made in the UHF region. Although the number of observed cases where such occurrences would make a circuit unusable are few, interference from this source may be serious in areas of high thunderstorm activity.

Reflections of signals from meteor trails are a source of potential multipath which have been observed. These occurrences should be rare at UHF for paths shorter than 600 miles, approximately.

Attention must be given in the selection of receiving sites to provide freedom from man-made noise. Ignition noise has been observed to be one of the chief sources of interference. Arcing in electrical machinery is another source, and even such commonplace devices as contacts on thermostats can produce interfering signals.

4. Design of Long-Range Tropospheric Communication Circuits

4.1 Introduction

Section 2 treated the general theories and Section 3 the detailed experimental observation of long-distance tropospheric propagation. A brief summary of the observed gross characteristics of tropospheric fields well beyond the average radio line-of-sight is given here; the body of this section is then devoted to a description of the methods used in exploiting these fields for radiocommunication purposes.

The average UHF-SHF loss well beyond the radio horizon is very large as compared to that encountered in line-of-sight paths. At a wavelength of one meter, the median path loss near the surface may be some 40–60 db greater than the free-space loss upon first entering the extra-diffraction region; the total path loss increases with approximately the 8th or 9th power of the distance out to a distance of 500–600 miles beyond the radio horizon. The average received power appears to increase approximately in proportion to the wavelength.

There are large variations in the hourly median signal level over diurnal and seasonal cycles, the cumulative distribution of which, at a given distance, approximates a log-normal distribution. These variations decrease rather rapidly with distance beyond the radio horizon. The fields are generally higher in climatic regions and seasons characterized by a higher average water-vapor content and/or a persistent elevated atmospheric dielectric layer structure. Usually superimposed on these longer term variations is the rapid and intense fading characteristic of the persistent extra-diffraction propagation mode. This short-time fading rate approximates 1 cps at 1000 mc and 200 miles, is nearly proportional to the frequency, and the amplitude variations are reasonably well described by a Rayleigh distribution in the absence of aircraft reflection or ducting.

Radio energy appears to arrive at the receiving antenna over a measurable solid angle, fields measured on antennas spaced normal to the propagation path begin to depart from a constant phase relationship beyond spacings of 10 λ to 20 λ, the realized gain of antennas does not continue to increase indefinitely with the area beyond a plane wave gain of some 35–40 db, and the inherent propagation medium "bandwidth" is no more than a few megacycles at 100 miles beyond the usual radio horizon as measured with wide-angle antennas.

There are many similarities in the design of long-distance and line-of-sight tropospheric circuits. Most of the criteria which apply to line-of-sight site selection, for instance, apply equally well to those of long-distance circuits, and much of the terminal equipment, audio, multiplexing, etc., may be the same. There are, of course, certain marked differences.

In site selection, particular attention must be paid to obtaining a good foreground clearance and very low horizon angles over the great circle path connecting the two terminals in order to minimize the path loss. The much larger average path losses require correspondingly larger values of effective radiated power, a large receiving antenna cross section, most sensitive receivers, and efficient modulation-detection techniques. The high order of effective radiated power creates field intensities which are potentially capable of causing severe interference, especially to line-of-sight circuits, while the sensitive receivers and large receiving antenna area mark the receiving terminal as one highly susceptible to interference. The use of very narrow radiation beamwidths, diversity techniques and careful selection of modulation-detection methods is required in order to overcome the effects of intense short-time fading and the multipath-limited relatively small effective propagation "bandwidth."

When appropriate design measures are taken, circuits using extra-diffraction fields are capable of providing radio communication of a very high degree of utility and reliability. Trunk capacity (scores or even hundreds of telephone channels) circuits can be designed for individual path lengths of as much as 150 miles beyond the radio horizon with good long- and short-time reliability (in the order of 99.9 per cent) and high quality; such circuits can be placed in tandem and expanded into an entire network of circuits. In order to maintain 99.9 per cent reliability in a long network, the requirements for individual links must be correspondingly greater. Careful circuit design can provide for the transmission of digital information rates well in excess of 1 kc/sec with very low error-rate occurrence. Good quality black and white television transmission may also be obtained at somewhat shorter distances; apparently the only limitation on picture fidelity is that set by the antenna aperture-to-wavelength ratio which must be sufficiently large as to reduce the influence of multipath components to acceptable proportions. The use of larger antennas and transmitter powers permits the achievement of lower capacity circuits having excellent reliability and good quality over path lengths of as much as 400–500 miles beyond the radio horizon. It also appears possible to design ground-to-air communication circuits of limited capacity but very high reliability for distances of some 200–300 miles beyond the average radio horizon.

4.2 Determination of Circuit Performance Requirements

The adequate and sensible design of a long-distance tropospheric circuit demands that firm and detailed specifications of the expected performances be at hand at the outset. This demand is of even greater importance in the design economics of long-distance than in line-of-sight circuits, since the initial and continued cost per circuit is higher, and the inherent vagaries of the transmission medium are greater. The circuit design engineer must know the maximum rate of information transfer to be transmitted and whether single-circuit, tandem-circuit, or trunk performance is required. He must be given detailed specifications concerning the type of data to be transmitted, both the long- and short-time reliability requirements, the minimum signal-to-noise ratio and distortion permitted in analog transmission, and the number and distribution of short-time error rates permitted on data channels.

4.3 Selection of Fundamental Modulation Techniques

Performance requirements, propagation characteristics, and economic factors all influence the choice of modulation and multiplexing techniques. Many of the arguments favoring one method of modulation over another in the case of line-of-sight UHF-SHF circuits apply with equal cogency in the design of beyond-the-horizon circuits. For instance, frequency modulation with its attendant noise reduction advantage has found favor over amplitude modulation in the design of intermediate distance (100–150 miles) large-capacity circuits, in which the information to be transmitted is either a wide-band black and white television video signal, a multiplexed bank of voice channels, or, perhaps, a combination of both. For circuits of this length, where a number of large antennas, high transmitter power and sensitive receivers are used, there need be little limitation in effective propagation medium "bandwidth," and the received signal should hardly ever fall beneath the baseband threshold. Under such circumstances, the low distortion, high signal-to-noise ratio essential to wideband trunk service can be achieved.

In the design of appreciably longer circuits, however, arguments may be advanced in favor of single sideband, especially for most efficient transmission of multiplexed phone channels. For circuit distances well beyond 200–250 miles, it becomes increasingly difficult to design circuits capable of yielding consistently adequate signal-to-noise ratios in a wide FM baseband; also, the frequency selective fading across a given baseband can be expected to worsen with the distance, inasmuch as path loss and system factors indicate the use of lower frequencies and, consequently, lower antenna aperture-to-wavelength ratios on very long-distance circuits. Minimum bandwidth single-sideband modulation, combined with an adequate number of pilot signals to minimize the effects of the frequency selective fading, appears to present an attractive alternative solution to economical circuit design.

For the transmission of digital data where short-time error rates must be held to extremely low values, forms of frequency shift keying (FSK) modulation of the individual binary channels in conjunction with adequate diversity techniques appears most attractive. The use of narrow matched information band filters and even synchronous detection can provide high reliability and efficiency in transmission. Where the incidence of aircraft-induced multipath conditions is expected to be high, extreme care must be taken in the choice of antennas, diversity order, spacing and combiner techniques, and modulation-detection methods.

4.4 Selection of Operating Frequency Band[1]

There are many factors—some of which are conflicting—in the choice of an optimum region in the frequency spectrum for the operation of a long-distance tropospheric communication circuit. These may be roughly grouped into three areas: propagation characteristics; performance and cost of various RF equipments; and circuit design specifications. All must be carefully considered, and particular attention paid to the matter of interference to, and from, other services.

An initial factor to be considered is that of the radio-wave path loss and its variation. The long-time median value of the transmission loss relative to free space and the short-time fading rate both appear to vary, on the average, as approximately the first power of the frequency in the UHF and lower SHF regions. Particularly at VHF and the lower end of the UHF region, ionospheric influences such as reflections from meteor-induced ionization trails, or an ionospheric scatter component, can appear at surface distances beyond 500–600 miles. At wavelengths shorter than 5 cm, approximately, the rainfall attenuation during occurrences of high-intensity precipitation should begin to be noticeable.

Atmospheric noise ceases to be a major factor in circuit design in most cases at frequencies above 50–60 mc, although an exception to this statement may be found in those geographic areas where electrical storms occur very often within line-of-sight of the receiving terminal. Cosmic noise, even though decreasing somewhat more rapidly than f^2, is the predominant source of noise in a well designed receiver throughout most of the VHF region, and will probably continue to place a lower bound on receiver sensitivity well into the UHF region with the advent of the parametric and maser type of preamplifier.

The antenna design is capable of exerting a major influence on circuit performance. In the lower UHF region the use of large antennas is not appreciably influenced by antenna-to-medium coupling loss until very large apertures are employed. This combination of relatively low frequency and large antenna structure has its greatest application in the design of relatively lower capacity, very long range circuits. Even though the coupling loss increases as the beamwidth is reduced appreciably below approximately 1°, and the realized increase in circuit gain no longer increases with the area of the individual antenna, the employment of very large aperture-to-wavelength antennas is indicated whenever the requirement to reduce multipath-induced distortion is of great importance. The use of large D/λ structures also permits the use of the angular diversity technique, should this appear desirable from an economic standpoint.

The available output RF power levels are smaller at the shorter wavelengths. At this writing, many tens of kilowatts are generally available near 300 mc, a few tens of kw near 2000 mc, and several kilowatts near 8000 mc; the efficiency of generating these large RF powers tends to decrease with frequency throughout the UHF-SHF region.

[1] See also Section 6.

The higher frequencies appear to be favored for shorter and intermediate length paths, especially where large bandwidths are required and multiple frequencies are needed for large system operation. Lower frequencies appear to be more efficiently used for single relatively narrow-band long-distance circuits.

4.5 Siting

The average value of the path loss depends not simply upon the surface distance alone, but primarily upon the angular distance. For this reason, sites should be selected such that the horizon angles are small and, in fact, negative if possible. The amount of height gain to be derived for antenna heights immediately above the local terrain in the image antenna interference zone is small, especially for larger antennas at the higher frequencies, but some additional height may be warranted in order to keep the foreground clear of absorbing trees or foliage.

In those situations where exact path profile information and good meteorological data are available, path losses can be predicted with accuracy adequate for most applications. However, in some cases where unusual terrain or meteorological conditions exist, path loss tests may be desirable. If such tests are undertaken it is important to obtain a sufficiently large sample of data.

Sufficient area normal to the propagation path should be provided for the required number of duplex and space diversity antennas; the diversity spacing might well be 100 λ or more.

Careful attention to siting, and some good fortune, may yield locations in which it is possible to reduce interfering field strengths at angles away from the main beam by taking advantage of high-horizon elevation angles in the direction of major side lobes or from the direction of potentially serious interfering sources.

4.6 Some Equipment Design Features

The radio engineer, in addressing himself to the design of reliable and economic long-distance tropospheric circuits, has evolved a number of interesting techniques for coping with certain of the problems peculiar to the extra-diffraction propagation mode. Some of these are mentioned here.

The fact that these long-distance UHF-SHF fields are remarkably persistent in their minimum values has placed a burden upon the equipment designer to match this characteristic with extremely good long- and short-time equipment operating performance in order that excellent over-all circuit and system performance may be attained. One method used is that of providing active spares for all of the major circuit components. With a judicious choice of multiple transmitters, receivers, and space-diversity antenna arrays, the malfunctioning of any one of these major components will result in only a relatively small degradation in circuit performance. Further, no time is lost in switching-in spare equipment.

Careful control of frequency is needed in order to permit the minimum receiver predetection bandwidths, the use of coherent diversity combination, and sophisticated detection techniques.

Very careful attention must be paid to the reduction of antenna side- and back-lobe levels, not only in the basic antenna design but also in the antenna placement at the site. This attention to antenna radiation characteristics, when matched by careful consideration of transmitter-spurious radiation, does much to minimize the interference hazards which are potentially present in the use of very large effective radiated powers. In this same vein, some consideration has also been given to the automatic control of transmitter output power on duplex circuits by monitoring the received signal-to-noise ratio and adjusting the radiated power such as to provide only that level just required for adequate receiver performance.

Very large antennas are costly and, consequently, much attention has been given to their most useful employment. Very careful RF filter design and fabrication, for instance, will permit the use of one aperture for both transmitting and receiving on a duplex circuit, even though the ratio of transmitted-to-received power might well exceed 160 db. Also, the use of dual feeds allows the transmission of two orthogonal polarizations, and, by appropriate polarization selection, quadruple space diversity can be obtained with the use of only two large apertures instead of four.

5. Fields of Application

In some cases the need for long-range tropospheric circuits is evident. An attempt to employ line-of-sight microwave techniques for the bridging of long over-water distances or over most difficult and remote terrain features either can be technically impossible or may be economically most unattractive.

Under other circumstances, however, an initial choice between line-of-sight and beyond-the-horizon techniques is not immediately obvious and can only be sensibly made on the basis of a careful analysis of several technical, operational, and economic factors.

For instance, the circuit bandwidth requirement is of prime importance, since the techniques presently available for beyond-the-horizon use cannot transmit analog signals with high fidelity over bandwidths greater than a few megacycles. Also, the inherent short-time propagation characteristics require much more careful and sophisticated circuit design for the very reliable transmission of high digital data rates than do line-of-sight circuits. On the other hand, the designer of a long-distance circuit need not concern himself with the characteristics of the hundred, or hundreds, of miles of terrain between his terminals once he has selected appropriate site locations and horizon angles. Arctic, desert, jungle, over water, over land, and mountainous terrain: all may be reliably spanned by long-range tropospheric scatter circuits.

The traffic pattern is also a fundamental consideration in deciding between a beyond-the-horizon or line-of-sight system. There is no advantage, obviously, to the employment of long-range circuits if drop-offs are required every few tens of miles.

The initial investment in a long-range tropospheric system appears to be roughly comparable to a line-of-sight system for a total path length of some 150 miles or greater in favorable regions, on the assumption that the more limited bandwidth of the tropospheric system is adequate. High-power transmitters, multiple large-antenna installations, and highly sensitive and stable receiving equipment all contribute to making the initial investment sizeable. Since the RF circuit parameters required depend upon the path loss, it should be noted that signal intensities are generally higher in tropical than in arctic regions, and a beyond-the-horizon circuit can therefore be relatively more economical to install in the former. Comparative cost figures depend upon the number and location of repeater stations needed in each instance. The requirement for land acquisition, reliable prime power and maintenance only at the terminals increases the competitive position of beyond-the-horizon circuits in regions where geographical or climatological features make access to remote sites difficult. The reverse is obviously the case for services along accessible overland routes in the temperate zones.

In many instances, an optimum solution is to be found in the judicious intermixture of both long-distance and line-of-sight circuits throughout the over-all system. Although high effective radiated power is required on the former, careful attention to frequency selection and stability, antenna design, and path orientation will often allow contiguous use of both kinds of circuits.

The future may see the application of long-distance techniques to certain long-range air-to-ground communication problems, since studies have indicated the feasibility of obtaining reliable narrow-band communications to distances of a few hundred miles beyond the radio horizon.

Current applied research and development activities in such fields as information processing, modulation-detection, and low-noise preamplifiers, for instance, promise to increase the efficiency of long-distance tropospheric circuit design.

6. Factors Influencing Frequency Allocation

6.1 Frequency Bands of Interest for Present and Future Applications

Based on the information given in the previous sections together with known military and commercial requirements, some tentative conclusions may be drawn as to the useful frequency bands and the types of systems that might be employed. Table II gives the portions of the frequency spectrum in which systems are operating or are contemplated together with the characteristics of these systems.

TABLE II

Frequency Region (mc)	Type of Service	Circuit Length	Antenna Size	Transmitter Power
300 to 500	Multi-channel voice (up to 36 channels)	Distance up to 500 miles	60–120 feet in diameter	10–50 kw
700 to 900	Multi-channel voice (up to 132 channels)	Distance up to 200 miles	30–60 feet in diameter	1–10 kw
1700 to 2400	Multi-channel voice (up to 200 channels)	Distance up to 100–150 miles	30–60 feet in diameter	1–10 kw
4000 to 5000	Multi-channel voice (up to 200 channels)	Distance up to 50–100 miles	15–30 feet in diameter	1 kw
7000 to 9000	Multi-channel voice (up to 200 channels)	Distance up to 50–100 miles	15 feet in diameter	1 kw

It is difficult at this time to make an accurate estimate of the demands for each type of system. However, it is known that extensive military systems have been built outside the United States in the 700- to 900-mc band, and some other circuits have been installed in the 2000-mc region. In addition, it is known that there is domestic commercial interest in systems covering 100- to 150-mile distances with minimum cost equipment. This would indicate possible domestic interest in the 2000- or 4000-mc bands.

6.2 Bandwidth Requirements

The bandwidth required for long-range tropospheric communication systems is similar to that required by line-of-sight systems. Typical systems are designed for the transmission of multi-channel voice signals together with a limited amount of teleprinter and digital data transmission. The voice multiplex usually employs frequency division, single-sideband techniques. In the past, frequency modulation of the radio carrier has been extensively employed for both tropospheric and line-of-sight systems. The bandwidth requirements for frequency modulation systems are of the order of 12 to 40 kc per voice channel, depending on the amount of FM advantage and the quality desired. Thus, typical tropospheric systems of 24- to 120-voice channels, with deviation ratios of the order of 3, use from 1 to 5 mc of RF bandwidth for each direction of transmission. On the other hand, some present-day single-channel mobile systems use only 10 to 15 kc, and a line-of-sight microwave relay system capable of transmitting 2200 voice channels with a frequency separation of 28 mc between adjacent radio carriers will be installed in 1960.

In addition to frequency modulation systems, recent experimental work indicates that direct single-sideband modulation at UHF is now possible and may be helpful in overcoming the bandwidth limitations of long-range tropospheric transmission. With single-sideband modulation, the bandwidth directly occupied by the RF signal is only 4 kc per voice channel. Thus, it is possible to obtain a reduction of two to ten times in the RF signal bandwidth compared to FM, the amount of the reduction depending on the deviation ratio in the FM system; *i.e.*, on the FM noise advantage obtained. However, the efficiency of spectrum usage depends not only on signal bandwidth but also on the frequency separation required between independent radio channels operating in the same area. The minimum allowable separation, or guard band, is determined primarily by the frequency space needed for practical RF filters to achieve the required adjacent channel discrimination. The necessary guard bands are roughly comparable for FM and single sideband. As a result the relative frequency economy of the two systems in practical use turns out to be considerably less than that indicated by the ratio of signal bandwidths.

When present standard television signals are transmitted beyond line of sight, the bandwidth requirement must be at least as much as the 6 mc used in television broadcasting. For better quality and ease of interconnection with line-of-sight radio relay, some 8–14 mc of the standard 20-mc band of the FM radio relay system is frequently used. Special purpose television signals, in which picture quality much lower than that of the present standard is acceptable, can be transmitted over much narrower bands.

6.3 *Frequency Allocation Techniques*

Frequency allocation techniques used for tropospheric beyond the line-of-sight communication systems should be such that the high reliability performance of these systems is not impaired by interference. This indicates that interference criteria similar to those employed for high performance line-of-sight systems should be used.

In the case of line-of-sight systems, an interference criterion is commonly established such that the noise introduced into the system from interference will be less than the noise due to natural sources, receiver front end, and intermodulation. The noise requirements of typical telephone systems differ, depending on the particular performance standards employed. Usually the standards take the form of a specification of the minimum permissible telephone channel signal-to-noise ratio for various percentages of time. For example, such a requirement might be that a voice channel peak signal-to-average noise ratio be greater than 50 db for 50 per cent of the time, greater than 45 db for 90 per cent of the time, and greater than 40 db for 99 per cent of the time.

The signal-to-interference ratio in a proposed system may be determined by a study of the system parameters together with the parameters of other systems operating in this vicinity. The parameters that must be examined consist of the transmitter powers for the various systems, the antenna patterns for the various antennas, the orientation of the various antennas, the physical location of the various terminals, the topographical features of the region, the modulation techniques employed, the receiver noise figures, the spurious outputs of the transmitters, and the spurious responses of the receivers. In addition, it is necessary to know the statistics of the propagation such as are given in an earlier section. With the above information, it is possible to calculate the probability of interference of the given level. It can be easily shown that the interference is best reduced by the use of low side-lobe, high-directivity antennas and equipment having low spurious transmissions and responses. Using these techniques to fullest advantage, it is possible to operate a number of tropospheric scatter systems in the same frequency band in close geographical proximity. It is even possible to operate low-powered line-of-sight equipment in the same frequency band near high power tropospheric scatter links; this has actually been accomplished in certain systems, but it does require careful and individual adjustments.

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The Ninth Plenary Assembly of the CCIR*

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Summary—**The International Radio Consultative Committee held its Ninth Plenary Assembly in Los Angeles, California, April 2 through 29, 1959. The CCIR studies the technical factors affecting international radio telecommunications and makes recommendations to administrations which are of importance in the international allocation and usage of the radio spectrum. These studies are made by fourteen Study Groups, each dealing with certain aspects of international telecommunication. At the Los Angeles meeting, the technical aspects of relatively new techniques for conservation of spectrum space were considered. Transmission and receiving standards, propagation characteristics, monitoring methods, modulation techniques including single sideband and stereo sound broadcasting, and the new international aspects of telecommunication with and between space vehicles were among the many topics considered by the Assembly.**

Background

THE International Radio Consultative Committee (CCIR),[1] whose initials come from the French Comité Consultatif International des Radiocommunications, began with the radio conference of Washington, D. C., in 1927. This committee was formed by a number of countries interested in developing the technical and operating characteristics of international radio on a world-wide basis. The CCIR is an official part of the International Telecommunication Union which today is a specialized agency of the United Nations. The CCIR held its Ninth Plenary Assembly in Los Angeles, April 2–29, 1959, the first Assembly to be held in the United States since its inception. The results of the Los Angeles meeting are of considerable importance to all international radio matters, since many of the decisions which are reached at the International Radio Conference held in Geneva during the fall of 1959 are based on its findings.

Radio waves do not stay within international boundaries, and because of this, it has been recognized by all international users of radio that cooperation in the use of the radio spectrum is essential. It is also important to have common systems of communication operating in the same frequency band to permit international exchange of messages. The CCIR studies the technical and operational aspects of radio which determine the extent to which a radio system will provide a useful service, and the amount of interference that it may cause to other radio services, particularly across international borders. Following these studies, it makes recommendations for the solution of international radio problems.

The CCIR accomplishes its work through the participation of the radio engineers of the various national administrations in its study groups, and periodically, usually every three years, holds a Plenary Assembly meeting, at which all of the activities of the study groups are reviewed. The results are finalized and adopted by the participating administrations. The "questions" under consideration are formulated into "study programs" in which administrations and private operating agencies participate by performing technical studies. The results of these studies are prepared in the form of "reports." When sufficient progress has been made on particular subjects, "recommendations" are made as to operational techniques to be followed, based on the technical characteristics of radio equipment and propagation of both the desired and undesired transmissions.

The Ninth Plenary Assembly

The Ninth Plenary Assembly of the CCIR was held in the United States at the invitation of this Government. The United States delegation to the Eighth Plenary Assembly at Warsaw, Poland in 1956 invited the CCIR to hold its Ninth Assembly in the United States. The Assembly was attended by 37 administrations, 9 international organizations, 16 private operating agencies, and 8 scientific and industrial organizations. The members of the United States delegation, 73 in number, were designated from the representatives of Government agencies, private companies, and scientific and industrial organizations who had contributed to CCIR studies in this country and who assisted in the United States preparatory work. In addition to the official delegation, a number of United States industry representatives also attended the Assembly, because of the special circumstance that the Assembly was held here, and because of the great interest of our communications and electronics manufacturing industry in the CCIR activities.

According to the International Telecommunication Union General Regulations, the inviting administration provides the chairman for the Assembly. The chairman of the United States delegation, A. L. Lebel, Assistant Chief, Telecommunications Division, Department of State, Washington, D. C., therefore automatically assumed the chairmanship for the Assembly. The opening Plenary sessions of the Assembly were largely concerned with organizing the work for the following period of the conference. The technical work of the Assembly is conducted by fourteen study groups which are continuously active, conducting a large part of their work by correspondence between Plenary Assemblies. Some of these study groups hold interim meetings between Plenary Assemblies in which participating administrations exchange their views on a person-to-person basis.

* Original manuscript received by the IRE, October 9, 1959.
† National Bureau of Standards, Boulder, Colo.
[1] J. S. Cross, "The International Radio Consultative Committee," PROC. IRE, vol. 45, pp. 1622–1628; December, 1957.

In addition to the technical work, however, the organizational structure is reviewed at the time of each Plenary Assembly, so that the operations may be conducted on the most efficient basis.

At the Ninth Plenary Assembly the technical study group on Ground Wave Propagation, formerly Study Group IV, was combined with the group on Tropospheric Propagation to form a new Study Group V entitled "Propagation, Including the Effects of the Earth and Troposphere."

Considerable attention was given to radio telecommunication with and between space vehicles, since this subject is regarded as one of the newest, most important, and forward-looking subjects considered by the Assembly. It was subsequent to the 1956 Warsaw Assembly that the first artificial earth satellite was launched, and forcibly brought to the attention of the world that international cooperation was necessary in the space radio telecommunication field. An entirely new study group, with the numerical designation IV, was formed with the term of reference, "To study systems of telecommunication with and between locations in space."

The positions of chairman and vice-chairman of each of the technical study groups were examined by the Assembly, and a number of new candidates were appointed to these important international positions.

A new arrangement for the publication of Plenary Assembly documents was proposed and approved by the Plenary. Five volumes will be printed, rather than three as formerly. Volume I will contain "recommendations" and those "resolutions" which are addressed to other organizations; Volume II, "questions," and "study programs," "resolutions," and instructions addressed to the continuing work of the study groups; Volume III, technical "reports" from study groups; Volume IV, "reports" of the director and special committees set up by the Plenary, lists of the participants in the Plenary, lists of documents and a report on the time and location of the succeeding Plenary; Volume V, minutes of the Plenary sessions. The Assembly also considered whether the CCIR should participate in the technical assistance program of the United Nations, which will be called to the attention of the Plenipotentiary Conference of the ITU for their advice.

CCIR Study Groups

To get a more complete picture of the current activities of the CCIR technical studies, various study groups are outlined below, with their terms of reference, their international chairmen, and chairmen of the United States committees.

Study Group I—Transmitters

Chairman—Colonel J. Lochard (France)
Vice-Chairman—Prof. S. Ryzko (Poland)

Terms of Reference: 1) To make specific studies and proposals in connection with radio transmitters, and generally to summarize and coordinate proposals for the rational and economical use of the radio spectrum. 2) To study a number of problems concerning telegraphy and telephony from the transmission point of view. 3) To study spurious radiation from medical, scientific, and industrial installations.

Study Group I held an interim meeting in Geneva during 1958. Draft documents as a result of this meeting were available at Los Angeles as Annexes to the report of the chairman of the study group. Six working groups were established during the Los Angeles meeting, as follows:

IA—Frequency Stabilization of Transmitters
IB—Spurious Emissions
IC—Bandwidth of Emissions
ID—Telegraphic Distortions and Frequency-Shift Keying
IE—Bandwidth and Interference
IF—Unwanted Industrial Radiations.

The significant results of the work of the study group resulted in a complete new table of frequency tolerances for radio transmitters in the various services. A recommendation was made to the Administrative Radio Conference concerning new tolerances for the intensity of spurious emissions. A recommendation was made concerning the measurement of the bandwidth of an emission and regarding the definition of the bandwidth of emission. Recommendations were made in the standards for frequency-shift keying, four-frequency diplex systems and the arrangement of channels for multichannel telegraph systems. Most of the questions and study programs of the study group resulted from questions referred to the CCIR by the Administrative Radio Conference, Atlantic City, 1947. Recommendations of this study group are intended to improve the performance of transmitters and systems and thus to aid in the conservation of frequency space. Many of the studies involve system problems which are principally the concern of Study Group III. Conversely, much of the work of Study Group III requires the study of transmitter and receiver characteristics. Close coordination, and frequently joint action are required by Study Groups I, II and III. Consideration is being given to some reassignment of the work so as to make it more consistent with the general terms of reference of the three groups. Much of the work also involves coordination with the special International Committee on Radio Interference (CISPR) of the International Electrotechnical Commission (IEC), and many of the documents refer to collaborative CISPR work.

Chairman for the U. S. Committee for Study Group I: J. P. Veatch, Frequency Bureau, Radio Corporation of America, 425 13th St., N.W., Washington 4, D. C.

Study Group II—Receivers

Chairman—P. David (France)
Vice-Chairman—Y. Place (France)

Terms of Reference: Measurement of characteristics of receivers and tabulation of typical values for the different classes of emission and the various services. Investigation of improvement that might be made in receivers in order to solve problems encountered in radio communication.

This study group held an interim meeting in Geneva in August, 1958, and the major portion of the work was done at that meeting. However, several new papers were received subsequently and were incorporated in the documentation. The work of the study group was divided into three main areas.

IIA—Sensitivity and Noise Factor of Receivers
IIB—Receiver Designs
IIC—Frequency Stability of Receivers.

The main work consists of compiling data on receiver performance in its various aspects. These data appear in three recommendations, the first, "Noise and Sensitivity of Receivers," tabulates noise factors and sensitivities of the various types of receivers and is a revision of a previous recommendation on this subject. The second, "Selectivity of Receivers," tabulates this item for the various types of selectivity, depending upon the type of radio service in which the equipment is used. The third, "Frequency Stability of Receivers," is a revision of a previous recommendation on this same subject. In addition to these three main subject items, the study group considered the following general topics: spurious emission from broadcast and television receivers, spurious emission from receivers of special types, spurious emission from receivers excluding sound-broadcast and television, distortion in frequency modulation receivers due to multipath propagation, recommended methods for measuring amplitude modulation suppression in frequency modulation receivers, choice of intermediate frequency for receivers, methods of measuring phase/frequency or group delay/frequency characteristics of receivers, protection against keyed interference signals, and response to impulsive interference.

Chairman for the U. S. Committee for Study Group II: A. G. Skrivseth, Federal Communications Commission, Washington 25, D. C.

Study Group III—Fixed Service Systems

Chairman—H. van Duuren (Netherlands)
Vice-Chairman—S. Namba (Japan)

Terms of Reference: 1) To study questions relating to complete systems for the fixed and allied services and terminal equipment associated therewith (excluding radio relay systems). Systems using the so-called ionospheric-scatter mode of propagation, even when working on frequencies above 30 mc, are included. 2) To study the practical application of communications theory.

The work of this group was divided as follows:

IIIA—Communication Theory
IIIB—Communication System Considerations
IIIC—Antennas and Transmission Systems.

The aim of this group is to improve the performance of fixed services and to add to the available knowledge relative to this service. Included in the work of the group was a recommendation concerning the use of radio circuits in association with start-stop telegraph apparatus which includes standard ARQ subchanneling arrangement, a technique of arranging the subchanneling in a way which provides automatic unambiguous phasing of the channels usually reported upon. Requirement is reported of signal-to-noise ratios as a function of error rate in telegraph systems on fading circuits using nondiversity and diversity reception. The advantages of intermittent communication using fading signals are explored as means of increasing the mean information rate. The concept of transmission loss in radio-system studies was introduced in the work of the CCIR; the recommendation made by the CCIR is given in full in a recent letter.[2]

Chairman of the U. S. Committee for Study Group III: R. C. Kirby, Central Radio Propagation Laboratory, National Bureau of Standards, Boulder, Colo.

Study Group IV—Space Systems

Chairman—I. Ranzi (Italy)
Vice-Chairman—W. Klein (Switzerland)

Terms of Reference: To study systems of telecommunications with and between locations in space.

The importance of space telecommunication was recognized by the CCIR at the Los Angeles Plenary Assembly, and this new study group was formed with the above terms of reference. Study Groups V and VI have already introduced questions and study programs and even proceeded with reports on the subject of space communications. However, it is anticipated that these subjects will be dealt with in considerable detail in this new study group. To date a chairman for the U. S. Committee for Space Systems has not been selected by the CCIR Executive Committee for the United States preparatory work.

Study Group V—Propagation, Including the Effects of the Earth and Troposphere

Chairman—R. L. Smith-Rose (UK)
Vice-Chairman—A. I. Kalinin (USSR)

Terms of Reference: To study the propagation of radio waves over the surface of the earth, taking into account changes in the electrical constants of the earth and irregularities of terrain, and including the effects of the troposphere.

At the Los Angeles Plenary Assembly, old Study Group IV (Ground Wave Propagation) was merged with Study Group V (Tropospheric Propagation) to form a new Study Group V with the terms of reference as above. The work of this study group has included the publication of ground-wave propagation curves for fre-

[2] K. A. Norton, "System loss in radio-wave propagation," Proc. IRE, vol. 47, pp. 1661–1662; September, 1959.

quencies below 10 mc, the publication of a ground-wave propagation atlas for frequencies between 30 mc and 300 mc, and the publication of a new Atlas for frequencies between 30 mc and 10,000 mc for antenna heights up to 20,000 meters. These curves depend upon variation of the refractive index of the atmosphere with height above the earth's surface. In the past, propagation curves have been prepared for either a vacuum atmosphere or an atmosphere which has a linear decrease of refractive index with altitude. A more realistic assumption was adopted for the way in which the refractive index varies with height above the surface, in terms of a new "basic reference atmosphere" which takes the form of an exponential decrease of refractive index with height. Modifications to previously prepared propagation curves were adopted at Los Angeles to take this into account. The director of the CCIR has published the Atlases of Curves mentioned above and these are now available for purchase from the Director of the CCIR, Palais Wilson, Geneva, Switzerland.

Significant progress was made in Los Angeles in describing the methods of measurement of the characteristics of the earth's surface (its conductivity and dielectric constant), in describing ground-wave propagation over an inhomogeneous earth, and in the study of the effects of the variability of the terrain on radio propagation. A new set of radio propagation curves, giving tropospheric field strengths for distances beyond the horizon for frequencies of 30 to 300 mc were prepared by an international working party. These new curves include extensive data which have recently been collected in geographic locations throughout the world. The new curves make provisions for the variability of field strength with geographic location by providing meteorological correction factors which may be obtained from the world-wide climatic charts of the radio refractive index prepared by an international working group. A comprehensive report was prepared on the measurement of field strength, power-flux density (field intensity) radiated power, available power from the receiving antenna and transmission loss. This report collects in one place the material which heretofore was scattered throughout the CCIR publications and brings it up to date, particularly with regard to the applicability of field strength measurements to the description of the performance of complete systems. A separate report on the measurement of field strength for VHF and UHF broadcasting services, including television, was also prepared. Other studies were made concerning the use of tropospheric scatter transmission for long distance point-to-point communications, radio relay operations, the use of orthogonal polarization in broadcast planning, propagation across mountain ridges, multipath propagation, and frequency protection for radio astronomy.

Considerable attention was given to radio telecommunication with and between space vehicles by both the tropospheric and ionospheric propagation study groups since this subject is considered to be one of the newest, most important, and forward-looking subjects considered by the Plenary. A new recommendation was made that consideration be given to setting aside a number of small frequency bands, well spaced throughout the HF and higher bands, for telecommunication with and between space vehicles. Accompanying this recommendation was a comprehensive report describing the technical radio propagation characteristics of the various frequencies for space service. The subject was also brought to the attention of both the ionospheric and tropospheric commissions of the Union Radio Scientific Internationale (URSI) through the adoption of two resolutions requesting the URSI to furnish additional scientific information on this subject.

Chairman for the U. S. Committee for Study Group V: J. W. Herbstreit, Central Radio Propagation Lab., National Bureau of Standards, Boulder, Colo.

Study Group VI—Ionospheric Propagation

Chairman—D. K. Bailey (USA)

Vice-Chairman—E. K. Smith (USA)

Terms of Reference: To study all matters relating to the propagation of radio waves through the ionosphere, insofar as they concern radio communications.

The CCIR has traditionally taken considerable interest in ionospheric propagation, and the study group concerned with this subject has the largest number of items on its agenda of any CCIR group. The work has been subdivided by subject area as follows:

- VIA—Ionospheric Forecasting, Indices, Reliability, etc.
- VIB—MUF, Meaning, Representation, and Oblique Incidence Observations, etc.
- VIC—Field Strength Absorption and Associated Problems of HF, LF, and VLF
- VID—Atmospheric Radio Noise
- VIE—Fading
- VIF—Ionospheric Propagation at VHF
- VIG—Coordinating Group (On Matters Involving Other Study Groups and Other Organizations).

A choice of a basic ionospheric index is a subject to which the CCIR has devoted considerable attention. It is a necessary ingredient of any prediction of ionospheric characteristics which is normally made three to six, or more, months in advance. The long-term characteristics of the ionosphere result from solar activity. However, there are various types of solar activity (*e.g.*, photon radiation, corpuscular radiation, solar flares), and they affect the ionosphere quite differently. An international working group studying this question presented a report on their findings at the Los Angeles meeting.

Short term variations, such as ionospheric storms, are also of great interest in ionospheric communication. The identification of precursors of these variations was the subject of a study in Los Angeles. The different types of forecasts used for ionospheric propagation were evaluated; the organizations responsible for them and the addresses from which they can be obtained were also given. Even if the perfect ionospheric index were available, there is still a problem connected with the representation of any given ionospheric characteristic used in a typical forecast. This stems from the fact that it is necessary to represent time of day, season of the year, as well as latitude and longitude (where world-wide prediction is involved). One method of presentation which has been used is to divide the world into zones and then to represent for the various zones the ionospheric characteristics involved for each month of the year in question. Another method of presentation is one which portrays the ionospheric characteristics on a world map which represents an instantaneous picture of the characteristics around the world. These maps may then be produced for each hour or each alternate hour of the month in question, thereby eliminating sharp discontinuities as one passes across zonal boundaries.

The user of ionospheric predictions for a given circuit will frequently observe deviations in fade-in and fade-out times for a given frequency from that which was predicted. These deviations may be due to propagation modes other than F-layer propagation, such as sporadic-E, ground back-scatter, or other peculiar modes, or to errors in predicting the normal layers. The CCIR has studied the problems associated with this.

One of the sources of confusion in using ionospheric prediction charts has been the different meanings which the term "MUF" (maximum usable frequency) has attached to it. A recommendation was prepared at Los Angeles which defines such terms as "Classical MUF," "Standard MUF," and "Operational MUF."

The prediction of sky-wave field strengths has been a very important problem in the CCIR. The official medium frequency ionospheric propagation curves are still those adopted in Cairo in 1938, as far as international usage is concerned. A considerable effort has been made in recent years, by many of the administrations and organizations which are members of CCIR, to collect data for a possible revision of the "Cairo curves." The European Broadcasting Union (EBU), International Broadcasting Organization (IBO) and the FCC have been the primary agencies which have been collecting the information. However, as the European studies, which are very extensive in scope, are not yet complete, no formal action was taken at Los Angeles to revise the Cairo curves. An international working party, which was organized in Warsaw in 1956, is examining the methods of estimating sky-wave field strengths for consideration by a future Plenary Assembly. All administrations have been urged to submit field strength measurements to this working party so that they may take advantage of data from all over the world.

The study of atmospheric radio noise is extremely important in practical ionospheric communication, since considerable improvement in circuit performance can be obtained by knowledge of the intensity and location of the noise sources. An international working party has prepared maps of atmospheric radio noise and has encouraged the various administrations to obtain radio noise data throughout the world. The CCIR at Los Angeles also decided to take on the study of man-made radio noise.

Fading plays an important role in practical communication. Normal predictions give a median received field strength, and what is needed in practice is data on the amplitude distribution and rapidity of field strength variation. An international working party was established at Los Angeles to study the phenomena of fading.

Considerable interest and attention was paid to ionospheric scatter and meteor-burst propagation. Because of the relative newness of these two modes of propagation, a resolution was prepared which draws attention primarily to the needs and special problems of services utilizing these modes. A report on ionospheric scatter and another one on meteor-burst propagation were prepared to give these types of propagation the status of knowledge.

Other subjects considered were the protection of frequencies used for radio astronomical measurements, frequencies useful for space telecommunications, and the whistler mode of propagation.

Chairman for the U. S. Committee for Study Group VI: D. K. Bailey, Page Communications Engineers, 710 14th St., N.W., Washington, D. C.

Study Group VII—Standard Frequencies and Time Signals

Chairman—B. Decaux (France)

Vice-Chairman—M. Boella (Italy)

Terms of Reference: Organization of world-wide service of standard frequency and time-signal transmissions. Improvement of measurement accuracy.

The provision of standard frequency transmissions and time signals in various parts of the world has long been a subject of study by the CCIR. Recommendations have been prepared concerning the frequencies to be used for standard frequency transmissions, the accuracy of these transmissions, time signals which are associated with these transmissions, and the general character of transmissions to be supplied. At Los Angeles, the CCIR recommended closer tolerances than had heretofore been used, so that the frequencies transmitted may be accurate to within one part in 10^8, and the time intervals transmitted may be accurate to within one part in 10^8 plus or minus one microsecond. Ways which may lead to partial solution of the problem of mutual interference were investigated, *e.g.*, by eventually removing all audio tone modulation from 5, 10, and 15 mc standard frequency transmissions. Staggering the frequencies within the bands already allocated is expected to reduce mutual interference, improve the use of standard frequency broadcasts for measuring radio propagation characteristics, and also leave space for radio noise measurements.

An item of particular interest is the study of the technical characteristics of a possible allocation of a new standard frequency band of 100 c in the neighborhood of 20 kc (15 to 25 kilocycles per second). It is expected that a standard frequency broadcasting transmission in this band of frequencies will give essentially world-wide coverage from one transmitting location.

Chairman for U. S. Committee for Study Group VII: W. D. George, National Bureau of Standards, Boulder, Colo.

Study Group VIII—International Monitoring

Chairman—J. D. Campbell (Australia)
Vice-Chairman—G. S. Turner (USA)

Terms of Reference: To study problems relating to the equipment, operation, and methods of measurement used by monitoring stations established for checking the characteristics of radio frequency emissions. Examples of such measurements are: frequency, field strength, bandwidth.

In order to carry out an efficient monitoring service of radio stations it is necessary for radio stations to be identified as regularly as possible during their transmissions. Identification is relatively simple for some classes of radio stations, such as those with slow-speed telegraphy. However, with some of the newer types of transmissions, identification at monitoring stations is rather complicated. The CCIR has been making studies of the types of identifying signals which should be transmitted in order that monitoring stations throughout the world may perform this function. The work of the study group has been divided under the following general headings:

VIIIA—Identification of Radio Stations
VIIIB—Measurements of Monitoring Stations
VIIIC—Automatic Monitoring and Spectrum Measurements at Monitoring Stations
VIIID—Recommendations for Radio Regulations
VIIIE—New Questions and Study Programs.

After considerable study the CCIR recommended appropriate methods for identifying radio stations to be used in the transmission of different types of radio communications. The accuracy necessary in the measurement of radio frequencies by monitoring stations was reviewed so that monitoring stations would be able to measure transmitted radio frequencies well within the tolerances recommended for transmitters by Study Group I. Methods of measuring field strength by monitoring stations were also reviewed, and automatic monitoring methods for observing the occupancy of the RF spectrum were studied. The study group recommendation concerning the identification of radio stations, the organization of an international monitoring service, and the form of report for frequency and field strength measurements made at monitoring stations was referred to the Administrative Radio Council for consideration. New subjects under consideration by the study group are: 1) the visual monitoring of the RF spectrum; 2) monitoring of radio transmissions from space vehicles; 3) identification of sources of interference to radio reception; and 4) measurement of S values at monitoring stations.

Chairman for the U. S. Committee for Study Group VIII: G. S. Turner, Chief, Field Engineering and Monitoring Bureau, Federal Communications Commission, Washington, D. C.

Study Group IX—Radio Relay Systems

Chairman—W. J. Bray (UK)
Vice-Chairman—E. Dietrich (Federal German Republic)

Terms of Reference: To study all aspects of radio relay systems and equipment operating at frequencies above about 30 mc, including systems using the so-called tropospheric-scatter mode of propagati m.

Radio relay systems using time-division multiplex vary widely. An international agreement on many of their characteristics is extremely difficult to obtain. Furthermore, there is relatively little use made of these systems across international boundaries, and under these conditions it is not absolutely necessary to have complete international agreement. However, the CCIR has studied the various types of systems and has agreed on preferred characteristics covering many of their features.

Recommendations were issued giving preferred radio-frequency channel arrangement for line-of-sight radio relay systems of several different telephone channel capacities. For one such system of 600 telephone-channel capacity per radio channel operating in the 4000-mc band, the proposed arrangement differed greatly from the one now in extensive use in the United States and certain other countries. The Los Angeles Plenary Assembly recommendation on this subject as finally adopted includes both arrangements, to be used as appropriate in the regions concerned.

A preferred arrangement of channels was also recommended for a radio relay system of up to 1800 telephone-channel capacity per radio channel, to operate in the 6000-mc band. The arrangement adopted was identical to that of the radio relay system which is expected to find widespread use in the United States.

Tropospheric-scatter systems for radio relay purposes are receiving wide acceptance for use under certain conditions, and a number of considerations concerning these systems have been studied by the CCIR. The Ninth Plenary Assembly recommended that preferred radio-frequency channel arrangements for such systems should be a matter for agreement between the administrations concerned. Another recommendation was made which suggests ways of insuring that such systems be operated with maximum spectrum economy.

As in all CCIR assemblies, there were a number of recommendations particularly concerning maintenance problems, service channels and radio relay systems of small capacity, which were of more concern to certain regions than to others. Frequently the countries involved were those having limited experience, and the CCIR meetings provided the opportunity for them to benefit from the work of others in the same line.

Chairman for the U. S. Committee for Study Group IX: E. W. Bemis, Operation and Engineering Department, American Telephone and Telegraph Company, 195 Broadway, New York, N. Y.

Study Group X—Broadcasting

Chairman—A. P. Walker (USA)

Vice-Chairman—H. Rindfleisch (Federal German Republic)

Terms of Reference: To study the technical aspects of transmission and reception in the sound broadcasting service (except for tropical broadcasting), including standards of sound recordings and sound reproduction to facilitate the international exchange of programs; to study also the technical aspects of video recording in liaison with Study Group XI.

This study group has made studies and recommendations concerning the methods of recording programs with particular emphasis on standards for the international exchange of programs. For example, it has recommended the characteristics of tape recording and reproducing systems including tape width and tape speeds. It has also recommended international standards of audio recording on disc.

Another area of interest is the desired-to-undesired signal strength for satisfactory reception of broadcasting stations operating in the medium frequency, HF, and VHF ranges. These studies are important in the allocation of broadcasting stations in the most efficient manner so that intolerable interference will not occur in the intended coverage area of the broadcasting station.

Two new areas of interest in the broadcasting field were those of single-sideband broadcasting (SSB) and stereophonic broadcasting, both AM and FM, including the stereophonic broadcasting of the sound associated with television. New questions and study programs concerning these subjects were initiated by the study group.

Chairman for U. S. Committee for Study Group X: A. P. Walker, National Association of Broadcasters, 1771 N. St., N.W., Washington 6, D. C.

Study Group XI—Television

Chairman—E. Esping (Sweden)

Vice-Chairman—G. Hansen (Belgium)

Term of Reference: Television. The most important matters dealt with at Los Angeles by this study group were those relating to the use of the UHF bands for television in Europe and the adoption of common standards for color television, especially to facilitate the exchange of programs. The study group held an interim meeting at Moscow in 1958 to discuss these problems. At the Moscow meeting, a measure of agreement was reached to adopt 8-mc channels to facilitate frequency planning in Europe and a subcarrier frequency of approximately 4.43 mc for a compatible color system. However, final agreement is awaiting further study by the various administrations. Discussions at Los Angeles indicated that countries having an existing 625-line, 7-mc, black and white service were strongly against changing the separation of vision and sound carriers from 5.5 mc to 6.5 mc. A working group has been set up to study and make recommendations on the best standards to adopt. Demonstrations will be made to show the effects on both black and white and color receivers of the different standards.

Another area of interest was the frequency tolerance of television stations where it was pointed out that with very precise frequency control a lower desired-to-undesired field-strength ratio would result in less interference between adjacent television stations. The study group is most anxious to resolve the above questions prior to the next broadcasting conference for planning VHF/UHF assignments for the European region, most likely to be held in Stockholm about November, 1960.

Chairman for the U. S. Committee for Study Group XI: E. W. Allen, Chief Engineer, Federal Communications System, Washington 25, D. C.

Study Group XII—Tropical Broadcasting

Chairman—M. B. Sarwate (India)

Vice-Chairman—Mr. Ramchandani (India)

Terms of Reference: To study standards required for good quality service in the tropical zone and for tropical broadcasting systems; interference in the shared bands; power requirement for acceptable service; design of suitable aerials for short distance tropical broadcasting; optimum conditions for the utilization of frequency bands used for broadcasting in the tropical zone; other associated questions.

At the Los Angeles Plenary Assembly, the study group confirmed most of the recommendations which had been made at the Warsaw meeting. Some supplementary material was received concerning the atmospheric-noise level in regions using tropical broadcasting methods and the required signal-to-interference ratio for satisfactory reception of a tropical broadcasting transmission.

Chairman for the U. S. Committee for Study Group XII: C. H. Pease, International Broadcasting Service, United States Information Agency, Washington, D. C.

Study Group XIII—Mobile Services

Chairman—G. H. M. Gleadle (UK)

Vice-Chairman—N. J. Soberg (Norway)

Terms of Reference: To study technical questions regarding the aeronautical, maritime, land mobile, and radio location and navigation services; and miscellaneous operating questions of concern to several services.

In some areas of the world the frequencies used by aeronautical-mobile and maritime radio telephone services are very congested. At Warsaw, it was considered that the use of single-sideband telephony equipments on board ships and aircraft would tend to reduce congestion of the radio frequency spectrum. In Los Angeles, it was recommended that single-sideband operation be introduced as far as operationally required for these services. This recommendation included the technical characteristics to be employed in single-sideband operation providing for an ultimate system of completely suppressed carrier and an interim system using -16 to -26 db of carrier suppression. It also established a means of exchange of information regarding tests and experience in the use of single sideband for the aeronautical mobile service between the CCIR and the International Aviation Organization (ICAO).

A new recommendation was adopted which calls for the earliest possible standardization of selective calling systems in the VHF maritime mobile bands. Additionally, a new study program was established so that administrations may concentrate their tests to the few systems now in operational use, looking forward to the adoption at the next Plenary Assembly of a single worldwide system.

A new recommendation was adopted which establishes standards for the maximum permissible level for spurious emissions falling within the VHF maritime mobile band.

A new recommendation was adopted which calls attention to the sources of interference in shipboard installations when two strong signals are present. The resulting interference from intermodulation may be reduced and essentially eliminated, if proper precautions are taken in design, installation and operation of maritime mobile equipments. The study group recommended that the radio-telephone distress and urgency signals of Mayday and Pan respectively, be re-established.

Chairman of the U. S. Committee for Study Group XIII: W. Mason, Frequency Bureau, Radio Corporation of America, 60 Broad St., New York 4, N. Y.

Study Group XIV—Vocabulary

Chairman, R. Villeneuve (France)

Vice-Chairman, Ferrai Toniolo (Italy)

Terms of Reference: To study in collaboration with the other study groups, and if necessary with the CCITT, the radio aspect of the following: vocabulary of terms and list of definitions, lists of letter and graphical symbols and other means of expression, systematic classification, measurement units, etc.

One of the principal concerns of the study group is the preparation of a vocabulary of radio terms useful in international radio matters in each of the three working languages, English, French, and Spanish. Extensive progress has been made over the past several years in the preparation of such a vocabulary. However, considerable difficulty is encountered in obtaining agreement on the definitions of the various terms. In the United States the IRE has been active in this work.

The Los Angeles Plenary Assembly gave attention to the nomenclature of frequencies in terms of cycles per second and Hertz. The continental European countries have a preference to standardize the Hertz as the unit of frequency rather than cps, as is done in the English speaking countries. The recommendation approved by the Assembly gives cps as the preferred term in the English text and the Hertz the preferred term in the French text.

Chairman of the U. S. Committee for Study Group XIV: L. G. Cumming, IRE headquarters.

More complete information on the technical subjects under study by the CCIR, *i.e.*, its program of work for the next three years, may be found in Volumes I, II and III of the Los Angeles Assembly, which contains the text of all recommendations, reports, questions, study programs, and resolutions adopted or reaffirmed at Los Angeles. These volumes may be obtained from the General Secretariat of the International Telecommunications Union at Geneva, Switzerland. Information in more detail on the United States activities for the Los Angeles Assembly on technical questions under study is available in the report of the United States delegation to the Assembly, and may be obtained free of charge by writing to F. C. deWolf, Chief, Telecommunications Division, Department of State, Washington 25, D. C.

The coordination of the preparatory work in the United States for the Tenth Plenary Assembly, both on the technical studies under consideration and on organizational planning for the Assembly in the United States is centralized in the Telecommunications Division of the Department of State and the over-all U. S. Preparatory Committee and the U. S. Executive Committee for the CCIR. The members of the various study groups are for the most part representatives of government agencies and private telecommunication companies or organizations having an interest in the work. The Department of State welcomes the participation of any person wishing to contribute to the studies. If further information on the United States Committee structure is required, it will be furnished in response to written requests to Mr. deWolf. On the other hand, any company or person not already participating in the work of the various study group committees and wishing to do so should get in touch with either Mr. deWolf or with the United States Chairman of the appropriate study group committee, whose name and address is shown herein.

The recommendations, resolutions and reports adopted by the Plenary Assembly were not signed and are not subject to ratification. They are drawn up and adopted for the guidance and information of countries on preferred standards or techniques applying to the operation of existing or new radio services. The results of the CCIR work are of considerable importance and interest for the deliberations of the ITU Administrative Radio Conference at Geneva, which began August, 1959, and which will review and revise as necessary the ITU radio regulations of the Atlantic City meeting in 1947.

According to the terms of the ITU Convention of Buenos Aires in 1952, each CCIR Assembly may select a site for the next assembly. Prior to the opening of the Ninth Assembly, the Indian Government notified the CCIR Secretariat of its intention to invite the Tenth Assembly to be held in that country. The Indian delegation issued the invitation at Los Angeles and it was accepted by the Assembly. (The Delegation stated informally that the best time to hold the Assembly in India, from the comfort standpoint, would be in the late fall or winter. Thus the Assembly will probably be held sometime in the period from November, 1962 to February, 1963.)

As a further item of interest, the IRE is represented in the CCIR Preparatory Committee work by L. G. Cumming, IRE Technical Secretary, who will be able to provide further detailed information about the participation of the IRE in the technical studies.

Crosstalk Due to Finite Limiting of Frequency-Multiplexed Signals*

CHARLES R. CAHN†, MEMBER, IRE

Summary—**It is a well-known fact that amplitude nonlinearity produces crosstalk between frequency-multiplexed channels, an important example being clipping in a saturating amplifier. Since the amplitude distribution of the composite signal for a large number of channels is essentially Gaussian, the crosstalk can be calculated with reasonable accuracy by assuming that the input signal is random noise, except for a narrow gap in which a sine wave is inserted to represent the signal in a selected channel. This assumption allows use of standard analytical techniques to determine the output signal-to-crosstalk ratio in the selected channel as a function of the clipping level. The resulting value for infinite clipping is about 9 db. The crosstalk decreases rapidly as the clipping level is raised, and a value of 40 db is obtained for clipping one per cent of time. An optimum clipping level, which provides the highest signal-to-total-interference ratio, may be determined when noise is present in the receiver, and allows definition of "peak factor allowance." An allowance of several db is found to be adequate for frequency-multiplexed binary data channels.**

Introduction

It is a well-known phenomenon in communication systems employing frequency-multiplexing that amplitude nonlinearity, such as amplifier saturation, causes crosstalk due to intermodulation between the signals in the various channels. As a result, when extremely high-quality transmission is desired, for example in toll-quality voice circuits, very little nonlinearity can be tolerated. However, in multiplexed teletype or other data systems, such high quality may not be necessary to insure a sufficiently low rate of errors, since the crosstalk causes no significant degradation in the transmitted data provided that its rms value lies reasonably below the signal level in each channel. Experiments[1] have been performed on multichannel radio-teletype transmitters to determine the amount of telegraph keying distortion introduced when the transmitter is driven well beyond its normal rating (for which third-order intermodulation products are negligible). The results obtained indicate that small peak telegraph distortions exist even under extreme overload conditions; they also seem to show that the crosstalk introduced by the amplifier saturation does not significantly degrade the transmission.

A model frequently employed for quantitative measurements of crosstalk between frequency-multiplexed channels utilizes random noise to represent the signals in all channels except the one reserved for the measurement. This procedure is justified by the central limit theorem, which implies that the superposition of a large number of independently varying small quantities yields a resultant with a Gaussian distribution of amplitude, regardless of the probability distributions of the individual contributions (within engineering reason). This procedure also has great theoretical advantage, since there are well-established analytical techniques for obtaining the effects of amplitude nonlinearities on Gaussian noise. The investigation presented in this paper utilizes the random noise model to study, both theoretically and experimentally, the important problem of the effects due to clipping of a frequency-multiplexed signal as a result of amplifier saturation in a

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[1] Conducted by the Army Signal Res. and Dev. Labs., under the supervision of H. F. Meyer, now at Ramo-Wooldridge.

radio transmitter. The results obtained are only for the case of a large number of channels at RF, but the conclusions derived therefrom should be applicable, at least qualitatively, to systems with a small number of channels. The method of calculation is applicable to this latter case; however, the random noise model cannot be expected to give the actual crosstalk with high precision when only a few channels exist. The calculations, therefore, have only been carried out for the case involving a large number of channels.

Description of Theoretical Model

The present investigation is concerned with a communication system in which the separate channels are frequency multiplexed in a narrow bandwidth about the carrier frequency by single-sideband (SSB) modulation. (*Note:* Single-sideband refers to the method of frequency translation from a low or audio frequency to RF. With digital data transmission, the modulation in any channel is, of course, AM, FSK, or phase-shift, all of which actually give double-sideband (DSB) signals.) The idealized amplifier saturation characteristic is shown in Fig. 1, and applies to the instantaneous total signal voltage. Thus, distortion will occur if the total instantaneous input voltage exceeds the value a. However, only the frequency components near the original frequency band are actually transmitted, since ideally the higher order harmonics are completely suppressed in a tuned amplifier. This idealized characteristic is only an approximation of actual saturation characteristics, but it is a good representation when, for example, the grid voltage swing is limited in an amplifier. The amplifier is shown with unity gain, for convenience, so that its instantaneous output (before filtering to the fundamental band) cannot exceed the value a. Thus, the peak envelope power (PEP) output from the transmitter is $a^2/2$ for no distortion with two tones simultaneously modulating the transmitter. The calculated third-order distortion terms increase rapidly as the transmitter is driven beyond this limit, as shown in Fig. 2 and discussed in Appendix I.

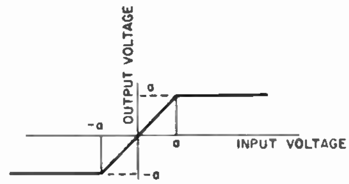

Fig. 1—Amplifier saturation characteristic.

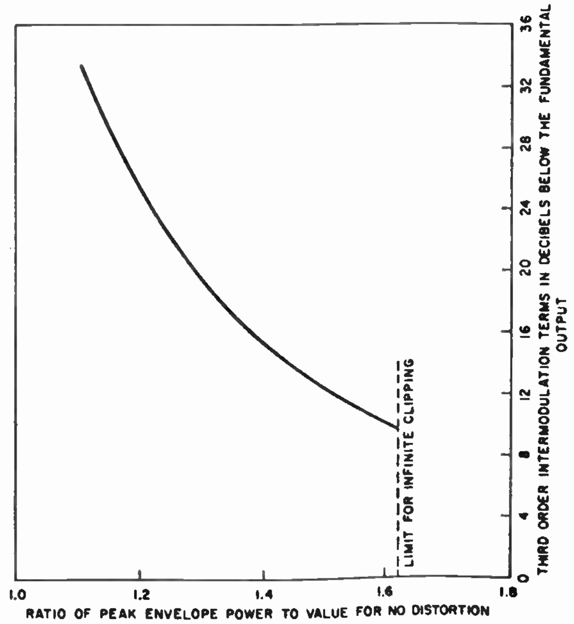

Fig. 2—Third-order intermodulation distortion for saturation amplifier.

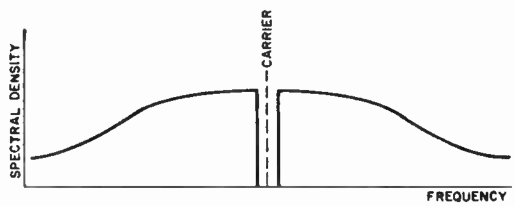

Fig. 3—Power spectrum of random input to saturating amplifier.

The crosstalk in the amplifier output will be calculated for the center channel, and for simplicity the composite input signal is assumed to have a symmetrical power spectrum about the center frequency. The center channel is selected because the computations for crosstalk are most easily carried out for it and, as will be shown later, the crosstalk can be expected to be most severe in this channel. Since random noise is assumed in all channels except the center channel in which the crosstalk is to be determined, the spectrum of the total noise-like input to the saturating amplifier is essentially like that shown in Fig. 3. The gap in the spectrum results from the elimination of noise in the center channel, and any fine spectrum structure resulting from guard bands between channels, etc., is ignored.

The signal in the center channel is represented by a steady sine wave, the power of which is just sufficient to "fill" the gap in noise spectrum. After distortion by the amplifier, the output sine-wave power will be changed, and crosstalk will appear uniformly distributed over the bandwidth of the gap. The output signal-to-crosstalk ratio is obtained by dividing the output sine-wave power by the total crosstalk power contained within the gap bandwidth.

Output Autocorrelation Function and Power Spectrum

As mentioned in the introduction, to insure validity of the representation of actual data signals by random noise, only the case of a large number of channels is treated. In this case, the signal sine-wave power in the center channel is very weak relative to the total power in the remaining channels, and the crosstalk power in the gap is essentially independent of the signal power. Thus, the signal power can be assumed to be zero for the calculation of the power spectrum of the crosstalk. For the assumed "memoryless" transfer device, the transform method of Rice[2] can be used to derive the correlation function and power spectrum of the amplifier output when the input power spectrum is given, as in Fig. 3. The general solution for the output correlation function, resulting with a nonlinear transfer characteristic having odd-function symmetry, is found to be given by the infinite series:

$$R_0(\tau) = \sum_{\substack{k=1 \\ k \text{ odd}}}^{\infty} \frac{h_{0k}{}^2}{k!} R_i{}^k(\tau), \tag{1}$$

where $R_0(\tau)$ is the output correlation function, $R_i(\tau)$ is the input correlation function, and the coefficients h_{0k} depend on the nonlinear transfer characteristic. For the limiter of Fig. 1, the coefficients have been derived previously[3] and are presented in Appendix II.

For the assumed situation where the spectrum exists in a narrow band about the center frequency f_0, and a tuned amplifier is employed, only the output spectral terms in the vicinity of the original center frequency will be amplified. Following Price,[4] a low-frequency normalized correlation function σ_τ is defined by:

$$R_i(\tau) = \sigma^2\sigma_\tau \cos 2\pi f_0\tau, \tag{2}$$

where σ is the rms value of the input and a symmetrical spectrum is assumed. For each value of k in (1), only the fundamental component of $\cos^k 2\pi f_0\tau$ is retained so that (1) becomes:

$$R_0(\tau) = \cos 2\pi f_0\tau \sum_{\substack{k=1 \\ k \text{ odd}}}^{\infty} \frac{h_{0k}{}^2\sigma^{2k}}{2^{k-1}\left(\frac{k-1}{2}\right)!\left(\frac{k+1}{2}\right)!} \sigma_\tau{}^k. \tag{3}$$

Eq. (3) may be used to obtain two important results: 1) the total output power by setting $\tau = 0$, and 2) the output spectral density by taking the Fourier transform. For typical σ_τ, the maximum value of the output spectrum occurs at $f = f_0$; hence, the crosstalk is most severe in the center channel.

Output Signal-to-Crosstalk Ratio

If a sine-wave signal of peak amplitude P exists in the gap in the input spectrum and $P \ll \sigma$, the output sine-wave power is found[5] to be:

$$S_0 = 2h_{10}{}^2 \cong \frac{P^2}{2} h_{01}{}^2. \tag{4}$$

As mentioned above, the amplitude P is chosen so that the input sine-wave power, $P^2/2$, fills the gap in the input spectrum. The output signal-to-crosstalk ratio is obtained by dividing the output sine-wave power, (4), by the output noise falling within the center channel bandwidth. The derivation is carried out in detail in Appendix II. It is found, not surprisingly, that the output signal-to-crosstalk ratio is a function only of the parameter a/σ, which expresses the ratio of clipping level to rms input. The exact form of the function depends on the shape of the input power spectrum.

To determine a typical form of the function, a Gaussian power spectrum is assumed for the input noise signal. It may be argued that a rectangular spectrum would be a better representation of a frequency-multiplexed signal. However, the computations for a rectangular spectrum are more difficult to carry out and should not yield significantly different results. In addition, the Gaussian spectrum approximates the finite slope of the skirts of typical band-pass filters or IF amplifiers more realistically. The computations for average power output and signal-to-crosstalk ratio are carried out in Appendix III and involve summation of relatively slowly converging series.[4] The computations were therefore programmed for the IBM 704 digital computer so that a sufficient number of terms in the series could be included. The results are presented as the curves in Figs. 4 and 5. To corroborate the theoretical results, a physical model was set up in accordance with the theoretical model described above. A number of points showing the experimentally determined relation between signal-to-crosstalk ratio and clipping level are presented in Fig. 5 along with the theoretical curve. The experimental points for a/σ large are approximate, since the maximum amount of notch filtering in the physical model was 23 db.

Another indication of the amount of clipping which occurs in the saturating amplifier is the fraction of time p during which envelope clipping occurs. This fraction of time is related to the parameter a/σ by means of the Rayleigh distribution for the envelope of Gaussian noise, according to:

$$p = e^{-a^2/2\sigma^2}. \tag{5}$$

The theoretical signal-to-crosstalk ratio is shown as a function of p in Fig. 6. This curve should provide a reasonably accurate specification of the allowable fractional clipping time in frequency-multiplexed systems when the tolerable signal-to-interference ratio for each channel is specified. For example, if 15 db signal-to-noise ratio is desired in any channel, Fig. 6 shows that clipping can be allowed for a maximum of 52 per cent of the time, if no other noise exists in the system.

Optimum Clipping Level

Since an increase in fractional clipping time increases the crosstalk while a decrease reduces the average transmitter power output, it may be questioned whether an optimum clipping level exists in the transmitter for a

[2] S. O. Rice, "Mathematical analysis of random noise," in "Noise and Stochastic Processes," N. Wax, Ed., Dover Publications, Inc., New York, N. Y., pp. 133–294; 1954.

[3] J. H. Laning and R. H. Battin, "Random Processes in Automatic Control," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 167–170; 1956.

[4] R. Price, "A note on the envelope and phase-modulated components of narrow-band Gaussian noise," IRE Trans. on Information Theory, vol. IT-1, pp. 9–13; September, 1955.

[5] W. B. Davenport, "Signal-to-noise ratios in band-pass limiters," *J. Appl. Phys.*, vol. 24, pp. 720–727; June, 1953.

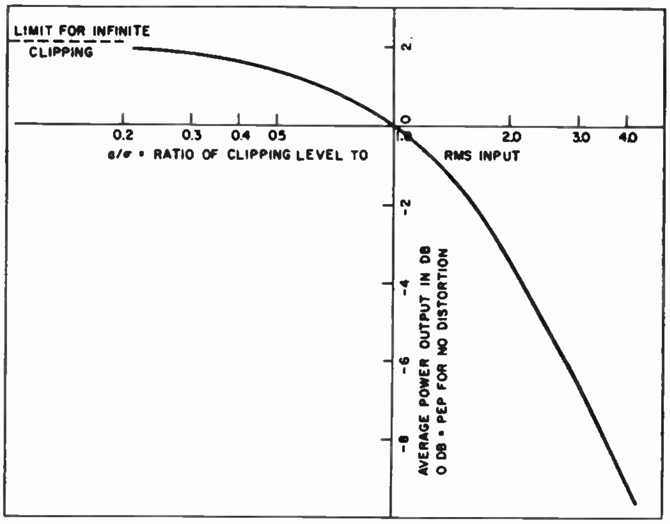

Fig. 4—Average power output as function of clipping level.

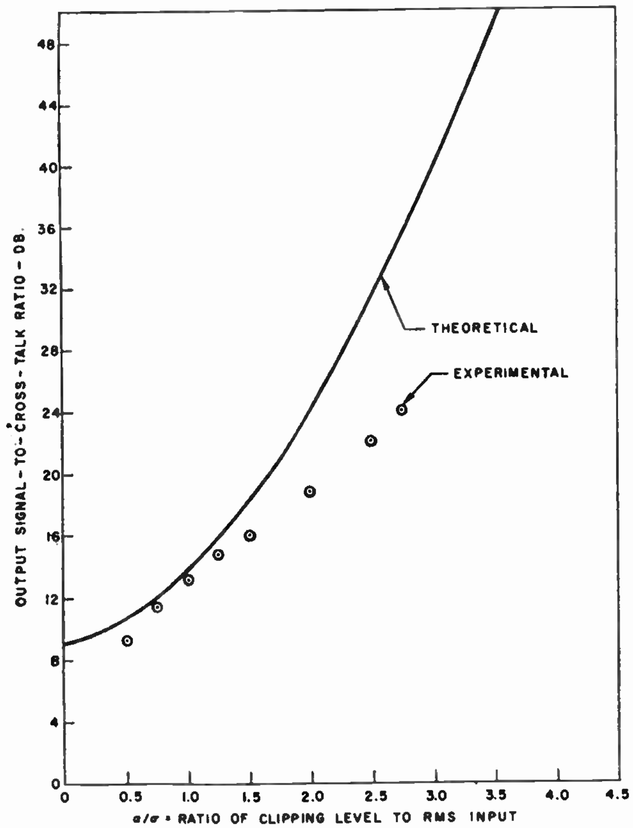

Fig. 5—Output signal-to-crosstalk ratio as function of clipping level.

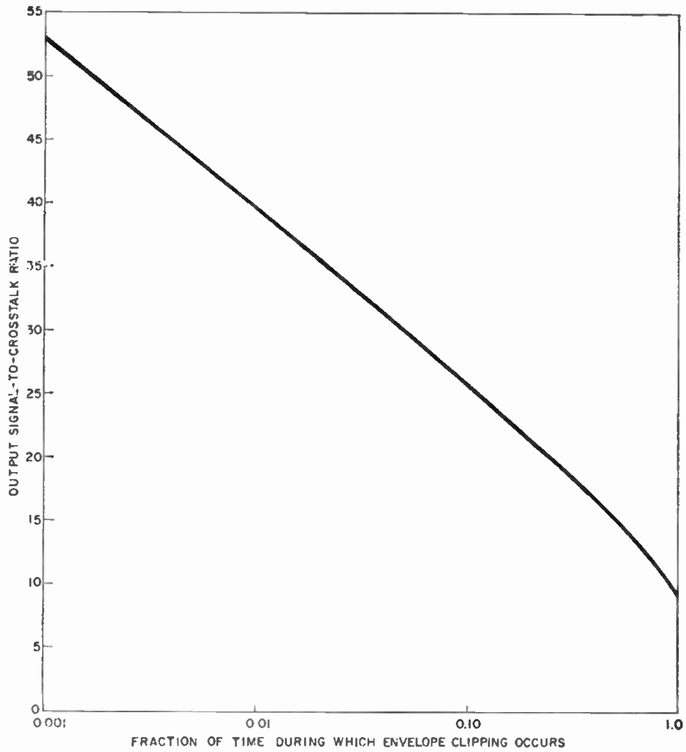

Fig. 6—Output signal-to-crosstalk ratio as function of fractional clipping time.

given communication system. Obviously, if no background noise existed in the receiver, no clipping should be used at the transmitter. Since some background noise is always present, it is desirable to increase the amplifier drive until the total interference given by the combination of crosstalk due to clipping and background noise is minimized in any channel relative to the signal intensity in that channel. That such a minimum exists will be demonstrated by using the results obtained above for the center channel.

Suppose, for simplicity, that the background noise has the same power spectrum as the (undistorted) signal and (referred to the transmitter) has the rms value σ_1. Thus, the quantity $a^2/2\sigma_1^2$ gives the ratio of peak envelope power (PEP) for no distortion divided by the background noise power. If there were no clipping in the power amplifier, and the average transmitter power output were made equal to $a^2/2$, the quantity $a^2/2\sigma_1^2$ would give the output signal-to-noise ratio for each channel in the receiver. When clipping occurs, the output signal-to-total-interference ratio may be expected to be less than $a^2/2\sigma_1^2$. For the center channel, the signal-to-interference ratio in the receiver is

$$\left(\frac{S}{N}\right)_0 = \frac{S_0}{N_C + N_R} = \frac{1}{N_C/S_0 + N_R/S_0}, \qquad (6)$$

where S_0 is the output sine-wave power, N_C is the crosstalk power, and N_R is the specified background noise power in the channel. Since S_0/N_C and S_0 have already been obtained as functions of the parameter a/σ, the variation of $(S/N)_0$ can be obtained for any value of $a^2/2\sigma_1^2$. The average power output from the transmitter is also a function of a/σ, so that $(S/N)_0$ can be plotted as a function of average power output. Further details are given in Appendix III. The resulting curves are shown in Fig. 7.

As expected, an optimum adjustment can be found for any specified PEP to background noise ratio so that the output signal-to-total-interference ratio is maximized. The locus of the maximum points is indicated in the figure. The curves also show that the adjustment is not critical; *i.e.*, a relatively broad range exists over which near optimum performance is obtained.

Additional useful information is obtained from the variation of the maximum output signal-to-total-interference ratio as a function of $a^2/2\sigma_1^2$, as given in Fig. 8. Since $a^2/2\sigma_1^2$ may be viewed as the "ideal" signal-to-noise ratio in each channel, the unavoidable degradation from this value indicates the additional transmitter power which must be provided to allow for signal peaking. This "peak factor allowance" is also plotted in Fig. 8. For a PEP-to-background-noise ratio in the range 10 to 20 db, the allowance ranges from 2 to 4 db, a result which appears to corroborate a published statement[6] that an allowance of a few db, for example 3 or 4 db with 10 or 12 channels, is sufficient to accommodate signal peaking.

Conclusion

The results obtained in this study indicate that the amount of crosstalk which results in the individual channels due to peak clipping of a frequency-multiplexed signal is surprisingly low even when clipping occurs for a large fraction of time. Even with infinite clipping, the crosstalk is 9 db down. These theoretical results are reasonably well corroborated by actual measurements, as indicated in Fig. 5, and appear to show that a relatively large amount of clipping or nonlinearity can be tolerated in a frequency-multiplexed communication system when moderate output signal-to-interference ratios are sufficient. The results are also in close agreement with those of a previous approximate analysis.[7]

[6] P. Mertz and D. Mitchell, "Transmission aspects of data transmission service using private line voice telephone channels," *Bell Sys. Tech. J.*, vol. 36, pp. 1451–1486; November, 1957.

[7] Private communication from W. Morrow of M.I.T. Lincoln Laboratory, Lexington, Mass. The approach used by Morrow was to take the power contained in the clipped-off portion of the wave and spread it over the bandwidth of the signal.

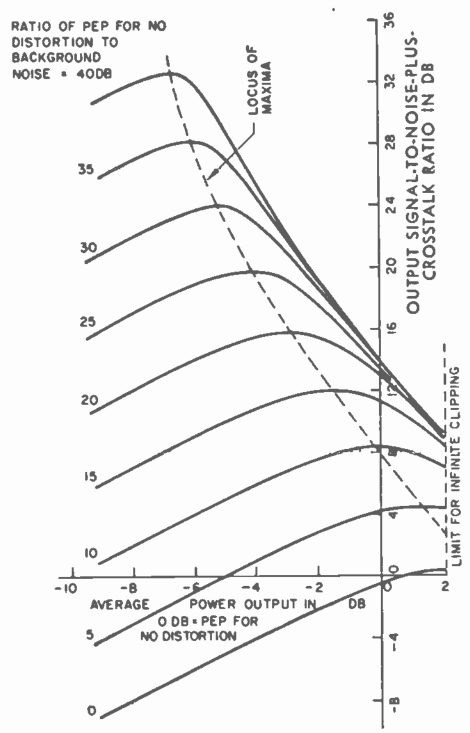

Fig. 7—Output signal-to-total-interference ratio as function of average power output.

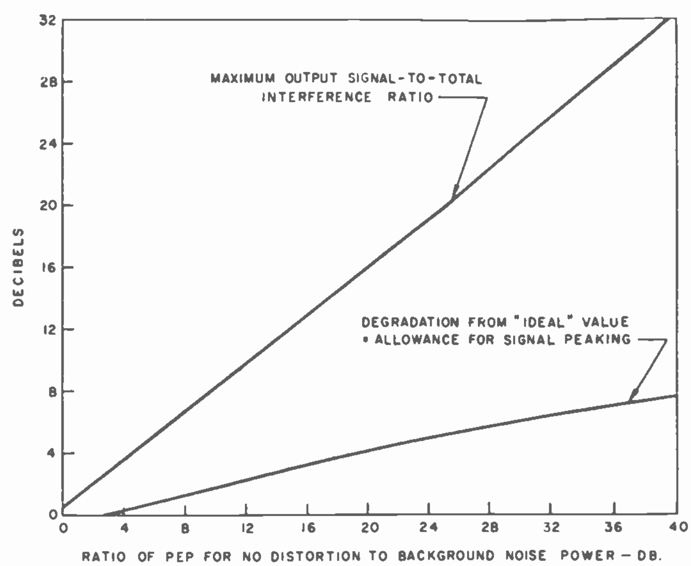

Fig. 8—Maximum output signal-to-total-interference ratio.

The results presented in Fig. 8 indicate the allowance in average signal power which must be made for signal peaking in a frequency multiplexed system. Over the normal range used for data transmission relatively little allowance is actually necessary; a few db are ample despite the existence of a large number of channels.

The study also indicates the possibility of transmitting frequency-multiplexed information over binary channels by infinite clipping. It is recognized, of course, that signal integration techniques are necessary to provide a usable signal-to-interference ratio in the channels.

A corollary of the present study is further insight into the reasons for the high intelligibility of constant-amplitude[8] speech. It is seen that if the speech waveform is considered as a frequency-multiplexed signal, the individual channels specify the time variations of the power in various portions of the audio spectrum, and thereby contain the information usually considered necessary for intelligibility. If the speech signal is translated to RF by SSB modulation and then limited, "crosstalk" will be developed between the various portions of the spectrum but on the basis of speech articulation tests, not in sufficient degree to seriously degrade intelligibility, provided that the spectrum was equalized initially by pre-emphasis of the higher audio frequencies. This is in qualitative accord with the results of the present study, which indicates that for infinite clipping the crosstalk is about 9 db below the signal level. Quantitative agreement is less satisfactory, since a 9-db average signal-to-noise ratio in a speech signal would yield lower intelligibility than is found for clipped speech. When clipping is performed at audio,[9] the results of the present investigation are not directly applicable, but qualitatively lead to the same conclusions.

The present investigation has been concerned only with independent channels multiplexed at RF. It would be desirable to extend the results to cover the case of clipping when the signal spectrum extends from a very low or zero frequency. The extended results would then be applicable to FM or AM multiplex systems in which a definite peak limitation is imposed on the composite modulating function, and they would indicate the tolerable amount of clipping in such systems.

Appendix I

Third Harmonic Distortion with Limiter

A measure of the linearity of a tuned power amplifier may be obtained by the two-tone test, in which the third-order intermodulation products between two equal tones are measured. These distortion terms are the lowest order that fall in the vicinity of the amplifier passband. The two equal tones may be expressed as:

$$A \sin \omega_1 t + A \sin \omega_2 t = 2A \cos \frac{\omega_1 - \omega_2}{2} t \cdot \sin \frac{\omega_1 + \omega_2}{2} t, \quad (7)$$

indicating that the envelope is sinusoidal, as is typical for a standing wave. If the envelope is clipped at unit amplitude, the resulting distorted envelope wave can be expressed in a Fourier series, the third harmonic of which yields the third-order intermodulation terms.

The results of the analysis yield an output signal of the form:

$$\frac{A_1}{2}[\sin \omega_1 t + \sin \omega_2 t] + \frac{A_3}{2}[\sin(2\omega_1 - \omega_2)t + \sin(2\omega_2 - \omega_1)t], \quad (8)$$

the coefficients of which are easily shown to be given by

$$|A_1| = \frac{2}{\pi}\sqrt{1 - (1/2A)^2} + \frac{4A}{\pi}\sin^{-1}(1/2A),$$

$$|A_3| = \frac{4}{3\pi}[1 - (1/2A)^2]^{3/2}, \quad (9)$$

if $2A > 1$. The PEP due to the fundamental is $A_1^2/2$, compared with the PEP for no distortion of 1/2. The power in the third-order terms, relative to the output power in the fundamental terms, is given by A_3^2/A_1^2. This quantity, expressed in db, is plotted as a function of the PEP in Fig. 2. The PEP for no distortion is shown as unity on the abscissa of the figure.

Appendix II

Derivation of Output Signal-to-Crosstalk Ratio

If a sine wave plus Gaussian noise is applied to a "memoryless" nonlinear transfer device, the output correlation function can be obtained from that of the input by an infinite series involving coefficients defined by a contour integral as follows:

$$h_{mk} = \frac{1}{2\pi j}\int_C f(z) e^{\sigma^2 z^2/2} z^k I_m(zP) dz. \quad (10)$$

in which P is the (constant) peak amplitude of the input sine wave, σ is the rms value of the noise, I_m is the modified Bessel function of order m, and $f(z)$ is the bilateral Laplace transform[10] of the nonlinear transfer function. The contour C is essentially along the imaginary axis of the complex z plane. For the present analysis where $P \ll \sigma$, (10) may be simplified by using only the first term in the power series for the modified Bessel function. The only significant coefficients for the output noise terms when $P \ll \sigma$ are h_{0k}, and for these $I_0(zP) \cong 1$. The simplified integrals yield the results, for k odd and $\neq 1$:

$$k_{0k} = -\sqrt{\frac{2}{\pi}}\frac{1}{\sigma^{k-1}}\left[\frac{d^{k-2}}{dx^{k-2}} e^{-x^2/2}\right]_{x=a/\sigma}$$

$$= -\sqrt{\frac{2}{\pi}}\frac{e^{-a^2/2\sigma^2}}{\sigma^{k-1}} H_{k-2}(a/\sigma). \quad (11)$$

The coefficients are zero for k even, due to the assumption of a symmetrical limiter characteristic. For $k=1$, the formula is

$$h_{01} = \frac{2}{\sqrt{\pi}}\int_0^{a/\sqrt{2}\sigma} e^{-x^2} dx = \operatorname{erf}(a/\sqrt{2}\sigma). \quad (12)$$

[8] P. Marcow and J. Dagnet, "New methods of speech transmission," in "Proceedings of the London Symposium on Information Theory," E. C. Cherry, Ed., Butterworths Scientific Publications, London, Eng., pp. 231–244; 1956.

[9] J. M. C. Dukes, "The effect of severe amplitude limitation on certain types of random signal: a clue to the intelligibility of 'infinitely-clipped' speech," *J. IEE*, vol. 102, pt. C, pp. 88–97; 1955.

[10] The question of convergence is deliberately avoided here. See Davenport, *op. cit.*, for more complete discussion.

In (11), $H_n(x)$ is the Hermite polynomial of order n. $H_n(x)$ satisfies the recursion relation:

$$H_{n+1}(x) = xH_n(x) - nH_{n-1}(x). \tag{13}$$

As indicated in (4), the output sine-wave power is determined from h_{10}. Since $I_1(zP) \cong zP/2$, the coefficients h_{10} and h_{01} are related by:

$$h_{10} \cong \frac{P}{2} h_{01} = \frac{P}{2} \operatorname{erf}(a/\sqrt{2}\sigma), \tag{14}$$

as used in obtaining the last term of (4).

If (11) and (12) are substituted for the coefficients in (3) and τ is set equal to zero, the average output power P_0 is found to be given by the series:

$$\frac{P_0}{a^2/2} = \left[\frac{\operatorname{erf}(a/\sqrt{2}\sigma}{a/\sqrt{2}\sigma}\right]^2 + \frac{2}{\pi}\frac{e^{-a^2/\sigma^2}}{a^2/\sigma^2}\sum_{n=1}^{\infty}\frac{H_{2n-1}^2(a/\sigma)}{2^{2n-1}n!(n+1)!}. \tag{15}$$

Eq. (15) is plotted as Fig. 4.

To find the crosstalk power N_C in the center channel, the output spectral density $w_0(f)$ at $f=f_0$ is multiplied by the gap bandwidth B. This output spectral density is found by:

$$w_0(f_0) = 4\int_0^\infty R_0(\tau)\cos 2\pi f_0 d\tau$$
$$= \frac{2}{\pi}\sigma^2 e^{-a^2/\sigma^2}\sum_{n=1}^{\infty}\frac{H_{2n-1}^2(a/\sigma)}{2^{2n-1}n!(n+1)!}\int_0^\infty \sigma_\tau^{2n+1}d\tau. \tag{16}$$

Note that the fundamental output term is not included in (16), since it has no power within the center channel. The integrals in (16) may be found by a process of iterated convolutions[2] of the input spectral density $w_i(f)$. The existence of a very small gap at $f=f_0$ has a negligible effect on the results of the convolutions, and, hence, σ_τ in (16) may be regarded as the "low frequency" correlation function of an input spectrum similar to that in Fig. 3, but without the gap. This greatly simplifies the necessary computations and is followed henceforth.

The sine-wave amplitude P in (14) is found by equating $P^2/2$ to the power within the bandwidth B of the new "gapless" input spectrum $w_i(f)$ The spectral density at f_0 of this new spectrum is given by:

$$w_i(f_0) = 2\sigma^2\int_0^\infty \sigma_\tau d\tau, \tag{17}$$

as may be seen by substituting (2) in place of $R_0(\tau)$ in (16). The input sine-wave power required to "fill" the gap is thus found to be:

$$P^2/2 = Bw_i(f_0) = 2B\sigma^2\int_0^\infty \sigma_\tau d\tau. \tag{18}$$

Therefore, from (14), (16), and (18), the output signal-to-crosstalk ratio becomes:

$$\frac{S_0}{N_C} = \frac{\frac{P^2}{2}[\operatorname{erf}(a/\sqrt{2}\sigma)]^2}{Bw_0(f_0)}$$
$$= \frac{\pi e^{a^2/\sigma^2}[\operatorname{erf}(a/\sqrt{2}\sigma)]^2\int_0^\infty \sigma_\tau d\tau}{\sum_{n=1}^{\infty}\frac{H_{2n-1}^2(a/\sigma)}{2^{2n-1}n!(n+1)!}\int_0^\infty \sigma_\tau^{2n+1}d\tau}, \tag{19}$$

which is a function of the ratio a/σ, the functional variation depending on the spectrum of the input signal.

Appendix III

Computation for Gaussian Power Spectrum

The integrals in (19) may be evaluated for a specific form of σ_τ. As discussed previously, a Gaussian power spectrum is assumed for the input signal to the saturating amplifier. As may easily be shown, the correlation function for noise with a Gaussian spectrum is also Gaussian, so that one may write:

$$\sigma_\tau = e^{-\alpha\tau^2}, \tag{20}$$

where α is a constant related to the spectrum bandwidth. The integrals in (19) then become:

$$\int_0^\infty \sigma_\tau^{2n+1}d\tau = \frac{1}{\sqrt{2n+1}}\int_0^\infty \sigma_\tau d\tau. \tag{21}$$

Consequently, the output signal-to-crosstalk ratio becomes:

$$\frac{S_0}{N_C} = \frac{\pi e^{a^2/\sigma^2}[\operatorname{erf}(a/\sqrt{2}\sigma)]^2}{\sum_{n=1}^{\infty}\frac{H_{2n-1}^2(a/\sigma)}{2^{2n-1}n!(n+1)!\sqrt{2n+1}}}. \tag{22}$$

The results of computations performed on the IBM 704 are presented in Fig. 5.

The relation needed in (6) between background noise power N_R in the center channel and output sine-wave power S_0 may be found easily. N_R is given by (18) with σ^2 replaced by σ_1^2, where σ_1 is the rms value of the background noise. S_0 is found from (18) and (19) to be:

$$S_0 = \frac{P^2}{2}[\operatorname{erf}(a/\sqrt{2}\sigma)]^2 = \frac{\sigma^2}{\sigma_1^2}N_R[\operatorname{erf}(a/\sqrt{2}\sigma)]^2. \tag{23}$$

Hence, the desired ratio may be expressed as

$$N_R/S_0 = \frac{1}{a^2/2\sigma_1^2}\left[\frac{a/\sqrt{2}\sigma}{\operatorname{erf}(a/\sqrt{2}\sigma)}\right]^2. \tag{24}$$

This function, in conjunction with (22), may be used to obtain the curves in Figs. 7 and 8.

Acknowledgment

The author wishes to thank H. F. Meyer, Ramo-Wooldridge, for valuable discussions on the material in the paper and for many suggestions concerning the method of presentation. The experimental data in Fig. 5 were obtained by W. Ryti of Ramo-Wooldridge.

IRE Standards on Methods of Measuring Noise in Linear Twoports, 1959*

59 IRE 20. S1

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* Approved by the IRE Standards Committee, June 11, 1959. Reprints of this Standard, 59 IRE 20. S1, together with the tutorial paper, "Representation of Noise in Linear Twoports," which immediately follows the Standard in this issue, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York, N. Y., at $0.75 per copy. A 20 per cent discount will be allowed for 100 or more copies mailed to one address.

1. Introduction

SPURIOUS undesired signals are always present in signaling systems and their components. These spurious signals are usually called noise. Since noise reduces the amount of information that can be transmitted with a specific signal power, quantitative measures of noise are often indispensable to engineering evaluations of signaling systems.

Test measurements for noise might logically be expected to begin with some generally useful quantitative measure of information-handling capacity in the presence of noise, or of the amount of annoyance which it creates. Such tests can be made in specific instances, but numerous difficulties are encountered. Often they are subjective and hard to evaluate accurately. Furthermore, the effects of noise vary enormously, depending on the system in question, and on the levels and characteristics of both signal and noise. It is useful, therefore, to measure noise in the engineering terms commonly used to describe signals.

Measures in terms of power are particularly comprehensive for stationary Gaussian noise, which is completely characterized by its power spectral density. *Thermal noise*[1] and *shot noise* under constant operating conditions are examples of stationary Gaussian noise. Any noise that is steady or stationary in character, and which originates from the linear superposition of a large number of small independent events, is almost certainly Gaussian. Other types of noise exist that may not be Gaussian—for example, "impulse" noise from ignition systems, atmospheric noise, powerline hum, and crosstalk. Complete characterization of these types requires, in addition to spectral density, other information such as the waveform of a typical pulse or the phase of the interference. However, for practical purposes, treatment by the methods below is often adequate; if not, special methods, not described here, will be necessary to deal with these types of non-Gaussian noise.

For simplicity, linearity of system response is assumed in the test methods to be described. Since noise usually enters the system at a point where the signal level is low, linearity will generally exist in the sense that the principle of superposition applies; that is, signals and noise produce currents and voltages that are simply additive, without the complicated intermodulation effects that occur in nonlinear systems. When the noise arises from independent uncorrelated sources, the spectral densities of these independent noises are also additive in a linear system. Heterodyne systems, incorporating frequency shifters, are linear in this sense when the signals are small.

2. Noise Factor

The system whose noise performance is under consideration here is essentially the linear two-port transducer. Signals enter at the input *port*, are processed internally, and leave the transducer at its output port. Any noise input accompanying the signal input is processed in an identical manner. Noise sources internal to the transducer contribute an additional noise at the output port that is independent of the signal or noise input. The noise performance of the transducer is commonly rated by comparing the noise-power outputs of the actual transducer and of its noise-free equivalent.[2] One such measure of performance is the *noise factor* (*noise figure*). The noise factor, at a specified input frequency, is defined as the ratio of 1) the total noise power per unit bandwidth at a corresponding output frequency available at the output port when the noise temperature of the input termination is *standard* (290°K) to 2) that portion of 1) engendered at the input frequency by the input termination. The standard noise temperature 290°K approximates the actual noise temperature of most input terminations. An alternative but related measure of performance useful for very low-noise transducers designed to operate from input terminations with noise temperatures substantially below 290°K is the "effective input noise temperature." (See definition in Section 5.)

2.1 Variation of Noise Factor with Source Admittance

As defined, the noise factor depends upon the internal structure of the transducer and upon its input termination, but not upon its output termination. Thus, the noise performance of a transducer is meaningfully characterized by its noise factor only if the input termination is specified. The noise factor F of any linear transducer, at a given transducer operating point and input frequency, varies with the admittance Y_s of its input termination (called "source admittance" hereafter) in the following manner:[3]

$$F = F_0 + \frac{R_n}{G_s} \mid Y_s - Y_0 \mid^2 \qquad (1)$$

where G_s is the real part of Y_s, and the parameters F_0, Y_0, and R_n characterize the noise properties of the transducer and are independent of its input termination. Thus, the noise performance of a transducer can be meaningfully characterized for all input terminations through specification of the parameters F_0, $Y_0 = G_0 + jB_0$ and R_n.

The "optimum noise factor" F_0, at the given transducer operating point and frequency, is the lowest noise factor that can be obtained through adjustment of the source admittance Y_s. The "optimum source admittance" $Y_0 = G_0 + jB_0$ is that particular value of source admittance Y_s for which this optimum noise factor F_0 is realized. These interpretations of F_0 and Y_0 are evident from (1). The parameter R_n is positive and has the dimensions of a resistance. This parameter appears as the

[1] IRE standard definitions of italicized terms may be found in "IRE Standards on Electron Tubes: Definitions of Terms, 1957," 57 IRE 7. S2, Proc. IRE, vol. 45, pp. 983–1010; July, 1957.

[2] The noise-free equivalent of a transducer is a fictitious transducer with identical port-to-port transfer properties but with no internal noise sources.

[3] See tutorial paper, Haus, *et al.*, "Representation of noise in linear twoports," this issue, p. 69.

coefficient of the $|Y_s - Y_0|^2$ term in the general expression for F and, therefore, characterizes the rapidity with which F increases above F_0 as Y_s departs from Y_0.

The necessity of using more than one parameter to specify the noise properties of linear transducers was not at first widely recognized. In the low-frequency triode, for example, a single noise generator characterized by an equivalent noise resistance describes its noise properties with sufficient precision over the range of circuit parameters in which the tube is generally employed. The necessity for additional parameters did not become acute until transistors and high-frequency electron tubes were developed.

The parameters F_0, Y_0 and R_n can be calculated if the noise theory of the transducer is known or, alternatively, can be determined empirically from noise measurements. The empirical method is discussed in Section 4. Until this section is reached, it will be assumed 1) that the input termination of the transducer is specified and 2) that the noise factor under discussion is the noise factor appropriate to this particular input termination. Used in this context, the noise factor meaningfully characterizes the noise performance of the transducer.

2.2 Average Noise Factor

In any communication system the signal is distributed over some finite bandwidth over which both signal and noise time averages may vary with frequency. Theoretically, in treating the interfering effect of the noise, it would be necessary to consider the frequency distributions (spectra) of both noise and signal; but in practice many cases are sufficiently well approximated by considering only the total powers of signal and noise. When only the total noise power in the band need be considered, the noise performance of the transducer is again rated by comparing its actual noise output for some standard noise input to that noise output which would have been obtained had the transducer been noiseless. One such measure of performance is the average noise factor. The average noise factor is defined as the ratio of 1) the total noise power delivered into the output termination by the transducer when the noise temperature of the input termination is standard (290°K) at all frequencies to 2) that portion of 1) engendered by the input termination. For heterodyne systems, 2) includes only that portion of the noise from the input termination which appears in the output via the principal-frequency transformation of the system, and does not include spurious contributions such as those from an image-frequency transformation.

The quantitative relation between the average noise factor $\overline{F}$ and the noise factor $F(f)$ is

$$\overline{F} = \frac{\int F(f)G(f)df}{\int G(f)df} \tag{2}$$

where f is the input frequency and $G(f)$ is the *transducer gain*. The average noise factor is the weighted average of the noise factor over the band in question, the weighting factor being the transducer gain. To emphasize that the noise factor, as opposed to the average noise factor, is a point function of frequency, the term *spot noise factor* may be used. Either average or spot noise factor is a numeric, designating a power ratio, which may be expressed in decibels by multiplying its common logarithm by 10.

The average noise factor also depends upon the internal structure of the transducer and upon the admittance of its input termination, but not upon its output termination, except insofar as the output power mismatch varies with frequency, and thus modifies the frequency dependence of transducer gain.

The noise-factor concept is useful over the entire frequency range of engineering interest. Since it compares the actual noise with the fundamental limit set by thermal agitation, the noise factor gives a broad and direct evaluation of the degree to which a system approaches the noise-free ideal.

3. Measurement of Average Noise Factor

In order to measure the average noise factor, it is necessary to obtain a measure of the noise power that is actually delivered to the output termination. This measure is divided by a similar measure of the output noise that would have been obtained if the transducer were noiseless and merely transmitted the *thermal noise* of the input termination at standard temperature. The following general methods of measurement are suitable for performing the evaluation. In each method a measure of the output noise is taken directly, but the methods differ in the ways of determining the reference noise output of the ideal noiseless transducer.

a. CW-Signal-Generator Method (*See Section 3.1*): In this method a power meter at the system output and a calibrated signal generator at the system input are used to determine the frequency dependence of transducer gain. From this dependence a "noise bandwidth" is determined. By use of this noise bandwidth, that portion of the output noise resulting from the input-termination noise can be determined. Dividing total output noise by this reference noise gives the average noise factor.

b. Dispersed-Signal Source Method (*See Section 3.2*): In this method a signal generator having its available power dispersed uniformly over the pass band of the system, and calibrated in terms of available power per unit bandwidth,[4] is used to determine that portion of the

[4] The National Bureau of Standards intends to initiate a calibration service for X-band microwave noise sources on January 1, 1960, and to extend this to other common microwave bands as rapidly as possible. Calibration will be in terms of the noise power delivered by the terminals of the customer's source to a matched waveguide of standard dimensions for the appropriate microwave frequency band. This power will be expressed in db above kT_0B, with an accuracy of 0.1 db. Service will be available at three frequencies within the band. Costs are not yet determined, but past experience indicates that cost of calibration is comparable with the cost of the instrument to be calibrated.

output-noise power which results from the input-termination noise. Suitable dispersed-signal generators are thermionic noise diodes, gas discharge tubes, resistors of known temperature, or an oscillator whose frequency is swept through the band at a uniform rate.

c. Comparison Method (See Section 3.3): This method consists of direct comparison between the network being tested and a secondary standard in the form of a network of the same type for which the average noise factor has been determined.

Of these three methods the first, involving the direct measurement of noise bandwidth, is complicated. The noise-source dispersed-signal method is simpler, and therefore preferred in both laboratory and production situations. The gas-discharge-generator dispersed-signal method is especially useful for high frequencies where the noise output of a diode may be affected by transit-time or lead effects. The swept-oscillator dispersed-signal method may have some application as a supplementary method. Comparison methods are primarily of use in production testing.

The general arrangement of the apparatus for any of these methods is shown in Fig. 1. Certain requirements must be met by each of the three components of Fig. 1.

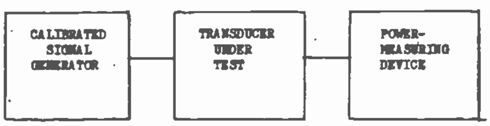


Fig. 1—*Average-noise-factor* test arrangement.

1) Since the calibrated signal generator is the input termination of the transducer, the behavior of the generator output admittance with frequency, over the pass band of the system, must duplicate that of the input termination with which the measured average noise factor is intended to be associated. The signal generator may deliver either power at a single frequency or power distributed over a frequency spectrum. In the latter case, its available power must be uniform over the pass band of the system. In either case, its available power must be accurately known. The calibration of the generator should take account of any elements that have been added to the basic generator to realize desired output admittance characteristics.

2) The transducer under test must be linear in the sense that its available-output-power change for a given available-input-power change is independent of the initial power level for all values used in testing. It may include linear elements or linear frequency shifters, but, particularly, simple envelope detectors must be excluded unless a noise source with a uniform power spectrum is being used as a signal generator. For example, in testing a conventional heterodyne receiver the power-measuring device must be connected ahead of the second detector unless the latter is itself a suitable power meter meeting the requirements described later.

3) The power-measuring device must indicate quantitatively the relative values of a), the noise power at the output of the system with no signal input and b), the total power at the output when an input signal is applied. The measuring device may be either a true power-measuring device, such as a bolometer or thermocouple, or some other type of instrument that has been calibrated to read power for the particular wave forms used in testing.

3.1 CW-Signal-Generator Method

The measurement of *average noise factor* with a CW sinewave signal generator will now be described. The *average noise factor* to be determined may be written

$$\overline{F} = \frac{N_0}{kT_0 \int G(f)df} \tag{3}$$

where

$G(f)$ = transducer gain at frequency f,
N_0 = noise-power output in watts,
k = Boltzmann's constant, 1.38×10^{-23} joules per degree K, and
T_0 = standard noise temperature, 290°K.

It is convenient to introduce quantities G_0 and B such that

$$G_0 B \equiv \int G(f)df \tag{4}$$

where

G_0 = transducer gain at some convenient reference frequency f_0.
B = noise bandwidth of the system in cycles per second (cps), and
$G(f)$ = transducer gain at frequency f.

In general, B is a function of f_0.

With the signal generator connected to the transducer, but with the generator output at zero except for its thermal noise at standard temperature, let the transducer-power output be P_1. With the available signal power of the generator set at P_s at the reference frequency f_0, let the transducer-power output be P_2. Then the transducer gain at f_0 is $G_0 = (P_2 - P_1)/P_s$. Also the noise-power output will correspond to the original value P_1, so that (3) becomes

$$\overline{F} = \frac{1}{\dfrac{P_2}{P_1} - 1} \frac{P_s}{kT_0 B}. \tag{5}$$

To measure $\overline{F}$ accurately it is desirable to choose P_s so that P_2 is several times greater than P_1. However, for convenience and to avoid saturation, it is common to choose P_s so that $P_2 = 2P_1$. Eq. (5) then becomes

$\bar{F}=P_s/kT_0B$, and P_s may be considered to be a power equivalent to the noise output as referred to the input circuit. It will be noted that, since only the ratio P_2/P_1 enters $\bar{F}$, an absolute calibration of the power-output measuring meter is not necessary, although power ratios must be determined accurately. Alternatively, a calibrated attenuator may be employed between the output of the transducer under test and the power output indicator.

3.1.1 Determination of Noise Bandwidth: The noise bandwidth may be determined from a plot, on linear scales, of the curve of relative transducer gain versus frequency, as shown in Fig. 2. The noise bandwidth is

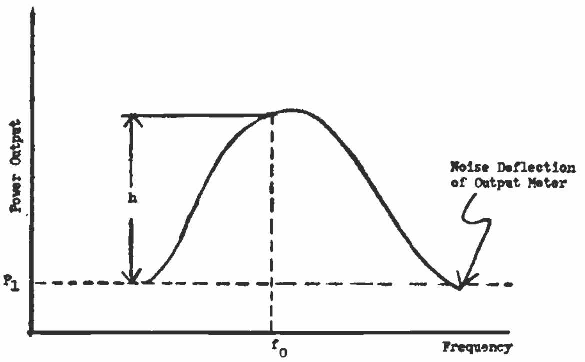


Fig. 2—Transducer relative power output as a function of the frequency of a sine-wave input from a signal generator with fixed *available power* output.

the width, along the frequency axis, of a rectangular response curve with an area M and with a height h at f_0 which are the same as those of the actual response curve. The frequency f_0 is usually, but not always, the frequency of maximum response. The area M and height h are those above the noise contribution to the power output, since only the signal output is of interest here. If the area is M, if the height at f_0 is h, and if the frequency scale is D cps per unit of length, the noise bandwidth B referred to f_0, is MD/h. Since the average noise factor refers only to the principal response of the system, spurious or image responses should be excluded from the response curve.

If the characteristics of the network are such that signals large compared with the noise can be handled without saturation, the effect of noise can be ignored in determining B. That is, the P_1 level in Fig. 2 may become negligibly low. It may be possible to achieve this same effect by reducing the gain of the network to considerably less than its normal value, but it should be ascertained that this does not cause undesirable saturation effects and that bandwidth is not altered by reducing the gain, as may happen if regenerative effects exist.

3.2 Dispersed-Signal-Source Method

Determinations of *average noise* factor may conveniently be made with a signal generator whose available power is distributed uniformly over the response band of the system. This method may eliminate the need for a direct determination of the noise bandwidth, as will be shown below. It is necessary that the available power of the generator in watts per cps of bandwidth be accurately calibrated. If the system is connected as in Fig. 1 with the generator output zero except for the thermal noise at standard temperature, a reading P_1 will be obtained on the power-output meter. (If desired, an equivalent passive termination may be substituted for the generator to obtain P_1.) If the generator is now made to have an available power density of p watts per cps in addition to the initial thermal noise, the power output will be increased by pBG_0 and the new output-meter reading will be P_2 or, in terms of the same symbols as used previously, $G_0=(P_2-P_1)/pB$. Also, as before, $N_0=P_1$, or by substitution in (3),

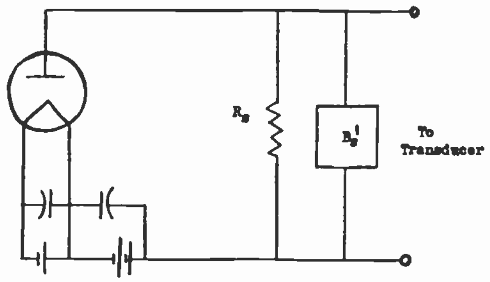
$$\bar{F} = \frac{1}{\dfrac{P_2}{P_1} - 1} \cdot \frac{p}{kT_0} \cdot \tag{6}$$

If the temperature T of the internal impedance of the generator (with the generator output zero except for the thermal noise) is not equal to the standard temperature T_0, (6) does not give the correct noise figure expression and a correction has to be introduced:

$$\bar{F} = \left(1 - \frac{T}{T_0}\right) + \frac{1}{\dfrac{P_2}{P_1} - 1} \cdot \frac{p}{kT_0} \cdot \tag{6a}$$

It should be noted that this measurement may involve spurious or image responses, which will ordinarily give an average noise factor; that is deceptively good, *i.e.*, too low, unless appropriate correction is made. In other words, if the dispersed signal covers pass-band responses not ordinarily used, such as images, which have appreciable gains compared to that of the pass band ordinarily used, the test signal will produce an effect in the output circuit greater than would be produced if the dispersed-signal source were limited so as to include only the response ordinarily used. Hence, if important spurious responses exist, a bandwidth measurement must be made and the *average noise factor* initially determined must be increased in the ratio of 1) the noise bandwidth including all spurious responses, to 2) the noise bandwidth when only the desired response is considered.

3.2.1 Noise-Diode Generators: A temperature-limited diode can be used as a noise generator when connected in a circuit such as that shown in Fig. 3(a). This circuit is typical, but many variations are possible. It is assumed here that the frequency-independent resistive



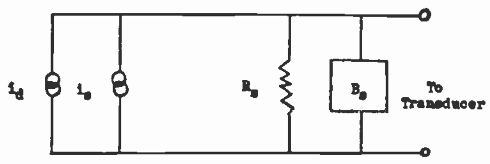


Fig. 3—(a) Typical diode noise-generator circuit; (b) equivalent circuit.

load R_s simulates the real part of the desired transducer source admittance Y_s at every frequency in the pass band of the system. The susceptance B_s' plus any stray susceptance associated with the diode and wiring must simulate the susceptive part B_s of Y_s. The equivalent circuit of this array is shown in Fig. 3(b). The *noise-current generator* i_s represents *thermal noise* in Y_s. For each small element of bandwidth Δf, the mean-square value of i_s for R_s at *standard noise temperature* is

$$\overline{i_s^2} = 4kT_0\Delta f/R_s. \tag{7}$$

The *noise current generator* i_d represents full shot noise in the diode current I_d. If the cathode temperature is adjusted to give a temperature-limited direct plate current of I_d amperes, then, for each small element of bandwidth Δf, the mean-square value of i_d is

$$\overline{i_d^2} = 2eI_d\Delta f, \tag{8}$$

where e is the electronic charge, 1.60×10^{-19} coulomb.

This noise-diode generator circuit has, over and above its thermal noise, a constant available noise power per unit bandwidth of p watts per cps, where

$$p = \frac{eI_dR_s}{2}. \tag{9}$$

Inserting (9) into (6), one obtains for the average noise factor the expression

$$\overline{F} = \frac{1}{\left(\frac{P_2}{P_1} - 1\right)} \frac{eI_dR_s}{2kT_0}. \tag{10}$$

Since the numerical value of $kT_0/e = 0.0250$ volt,

$$\overline{F} = 20I_dR_s \frac{1}{\left(\frac{P_2}{P_1} - 1\right)} \tag{11}$$

where I_d is in amperes, R_s is in ohms, and P_2/P_1 is the ratio of noise-power outputs.

As before, it is desirable that P_2 be considerably larger than P_1 (preferably several times larger) so that the difference between P_2/P_1 and unity can be determined accurately. For convenience, P_2 is often made equal to $2P_1$. Sometimes a smaller value of P_2 must be used because of limitations imposed on the diode plate current I_d. In such cases additional care must be taken to insure that P_1 and P_2 are stable, repeatable readings to assure reasonable accuracy in the result.

In the foregoing discussions no account has been taken of electron transit time in the diode. If the transit time is an appreciable fraction of a cycle, the noise output of the diode will be lowered.[5] At low frequencies the noise output of diodes may increase above the value (8) because of flicker noise.

3.3 Comparison Methods of Noise Measurement

The methods just described may not be convenient or necessary for production testing. In such cases, noise factors can be checked approximately by carefully chosen comparisons with the performance of a master standard unit of known noise factor. A measurement of signal-to-noise ratio after detection, with a fixed modulated-signal input, may be used for checking individual units, provided that the bandwidths of the networks both preceding and following the detector, and also the detector characteristics, are maintained within close limits.

3.4 Precautions

Care should be taken to insure that the apparatus attached to the network to be measured does not materially affect the bandwidth of the system. For example, if regeneration is introduced that greatly reduces the bandwidth, the noise factor may be markedly altered. In any event, undesired feedback usually makes the measurement much more difficult.

Careful shielding and filtering of input and output elements is essential, particularly when the difference in power level between output and input is large. If the measuring-signal generator has a temperature different from standard temperature, an appropriate correction must be applied.

[5] For transit-time correction, see D. B. Fraser, "Noise spectrum of temperature-limited diodes," *Wireless Engr.*, vol. 26, pp. 129–131; April, 1949.

4. Measurement of Spot-Noise Parameters

The noise factor discussed in Section 3 is the weighted average noise factor over the entire pass band of the network. The *spot noise factor*, which is the noise factor at a particular frequency, can be determined by including a very narrow band filter of the desired frequency between the network and the power-measuring device. When the spot frequency is near the center of the band, the factor so obtained may not be greatly different from the average.

4.1 Noise Factor of Transducers in Cascade

Frequently, several networks are connected in cascade, and it is desirable to know how the noise factor of each affects the noise factor of the over-all system. This is necessary both in evaluating the effect of improvement in any part of the system and in measuring the noise factor of a single unit in a system.

For a number of networks in cascade, as shown in Fig. 4, the system spot noise factor is given in terms of

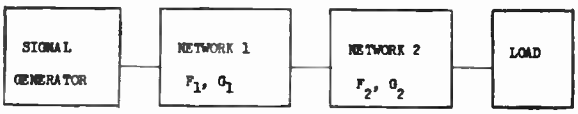


Fig. 4—Networks in cascade.

the component spot noise factors by

$$F = F_1 + (F_2 - 1)/G_1 + (F_3 - 1)/G_1G_2 + \cdots \quad (12)$$

where G_1, G_2, and so forth, are the available power gains of the component networks.[6]

The spot noise factor F_i and the available gain G_i of the ith network are those obtained with a source impedance equal to the impedance presented by the output of the preceding part of the cascade. An analogous expression for average noise factor can be derived, but is more involved. When the frequency for spot noise factor is chosen near the center of the noise-transmission characteristic, (12) for spot noise factor is often a satisfactory approximation for average noise factor.

These formulas also apply to networks that attenuate rather than amplify, in which case the corresponding gains are less than unity. It should be mentioned that the spot noise factor of a passive attenuating network at standard temperature is the factor by which the available power is attenuated in passing through it. It should also be noted that, when each of the various networks has substantial gain and the noise factor of the later stages is not excessive, then the over-all noise factor is largely determined by the noise factor of the first stage as is obvious from (12).

[6] If a negative output resistance exists somewhere in the cascade, the noise factor of the over-all system can be calculated as described by H. A. Haus and R. B. Adler, "An extension of the noise figure definitions," Proc. IRE, vol. 45, pp. 690–691; May, 1957.

4.2 The Noise Parameters F_0, G_0, B_0, and R_n

As stated in Section 2, the *noise factor* F of any linear transducer at a given input frequency varies with the admittance $Y_s = G_s + jB_s$ of its input termination in the manner shown in (1), which can be expanded as follows

$$F = F_0 + \frac{R_n}{G_s}\left[(G_s - G_0)^2 + (B_s - B_0)^2\right] \quad (13)$$

where G_0 and B_0 are the conductive and susceptive parts respectively of the optimum source admittance Y_0 cited in Section 2.

The four parameters F_0, G_0, B_0, and R_n characterize the noise properties of the transducer and are independent of the source admittance Y_s. These noise parameters can be determined empirically by fitting this four-parameter expression to observed values of F as a function of Y_s.

A suitable program for determining F_0, G_0, B_0, and R_n from noise-factor measurements is the following:

1) With the source conductance G_s maintained constant, measure several values of the *noise factor* F for different values of the source susceptance B_s. Plot the curve F vs B_s and determine the optimum source susceptance B_0.

2) With the source susceptance B_s maintained at its optimum value B_0, measure several values of the *noise factor* F for different values of the source conductance G_s. Plot the curve F vs G_s and determine the optimum source conductance G_0.

3) Using the data already obtained, plot F vs x, where x is the quantity $|Y_s - Y_0|^2/G_s$. These data should lie on a straight line of the form $F = F_0 + R_n x$. The slope of this line is the resistance parameter R_n and its F intercept at $x=0$ is the optimum noise factor F_0.

If a direct-reading noise-factor meter is available, the noise-factor minima in steps 1) and 2) can be observed directly on this meter, yielding values for B_0 and G_0. Additional data of F vs x for step 3) can then be obtained by using other convenient values of Y_s.

If the points in step 3) do not line on a straight line, then

a) the estimates of B_0 and G_0 in steps 1 and 2 may be inaccurate,[7]

b) the individual noise-factor measurements may be in error, or

c) the transducer under test may be nonlinear.

4.2.1 Equipment: The measuring equipment for determining the noise factor is discussed in Section 3. For the measurement of the noise parameters, it is necessary

[7] This can occur if the minima in the F-vs-B_s and -G_s curves are very shallow. In this case, slightly adjusted values of B_0 and G_0 may improve the linearity of the F-vs-x curve.

to provide, in addition, a means for adjusting the source admittance. The type of equipment needed depends on the frequency of the measurement.

An arrangement suitable for use at frequencies below one hundred or two hundred megacycles is shown in Fig. 5. The susceptance is adjustable by means of a calibrated capacitance or inductance and the conductance is varied by using different resistances or a variable resistance. At the higher frequencies of this range, care must be taken that the susceptance B_s is maintained at its optimum value B_0 when the conductance G_s is varied in the measurements specified in step 2). A temperature-limited diode or a resistor at known temperature can be used as a noise source. Gas discharge tubes combined with calibrated attenuators have also been used in the upper part of this frequency range.

Noise generators suitable for use at higher frequencies are, by nature of their construction, devices with fixed output conductance. A calibrated source with variable internal admittance can be made by combining such a generator with a network that acts like a variable-ratio transformer. An arrangement used for measurements on amplifiers at frequencies above 500 mc is shown in Fig. 6.

Fig. 5—Test arrangement for determination of noise parameters.

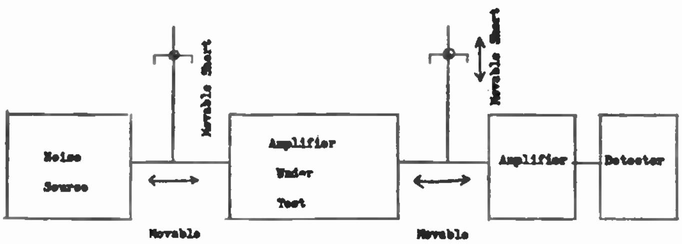


Fig. 6—Test arrangement used for measurements on amplifiers at frequencies above 500 mc.

The transformer consists of a section of transmission line fitted with an adjustable stub that can be moved along the length of the line. Two identical units are used. One unit couples the amplifier to the noise generator or signal generator. The second unit couples the output of the amplifier to the succeeding stage and is used to facilitate the measurement by maximizing the output. The correction due to the noise of the second stage is correspondingly reduced.

4.2.2 Sample Determination of the Noise Parameters at Low Frequencies: The determination of the noise parameters of a germanium alloy junction *p-n-p* transistor, operated in the common-emitter connection at a frequency of 900 cps, is presented in this section.

Fig. 7 shows observed F-vs-B_s data for a constant source conductance of 1.00 millimho. The optimum source susceptance B_0 is zero. Fig. 8 shows observed F-vs-G_s data for $B_s = B_0 = 0$. The optimum source conductance G_0 is 0.5 millimho. Fig. 9 shows the F-vs-x plot obtained from the data used previously in Figs. 7 and 8. The intercept and slope of the line plotted by use of these data yield $F_0 = 1.55$ and $R_n = 540$ ohms. The curves superposed on the experimental data in Figs. 7

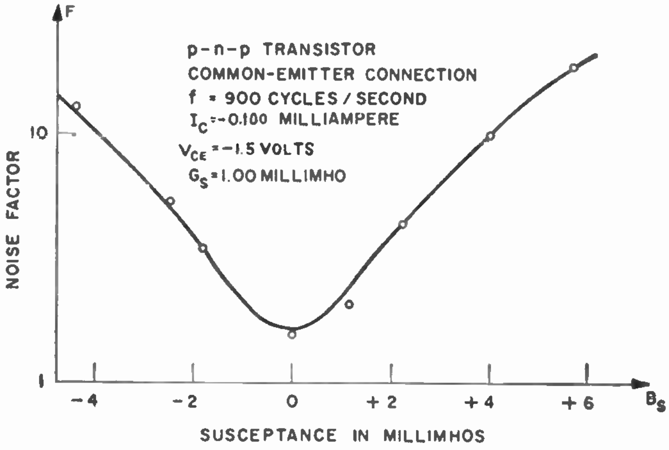


Fig. 7—Noise factor vs source susceptance.

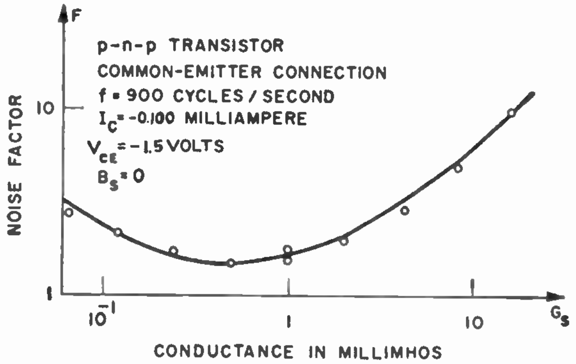


Fig. 8—Noise factor vs source conductance.

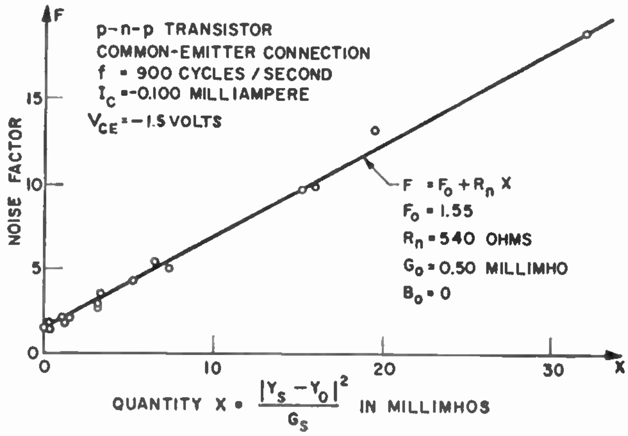


Fig. 9—Noise factor vs quantity x.

and 8 were computed from (13) for the values of the noise parameters F_0, G_0, B_0, and R_n determined above.

4.2.3 Sample Determination of the Noise Parameters at High Frequencies: The determination of the noise parameter for a disk-seal triode, made at a frequency of 870 mc, is described in this section. The arrangement of equipment shown in Fig. 6 was used. A series of readings with input-stub position and length chosen to give $G_s = 38$ millimhos for various values of B_0 provided the data shown in Fig. 10. The optimum source susceptance B_0 is found to be -54.0 millimhos. Fig. 11 shows observed F-vs-G_s data for $B_s = B_0 = -54.0$ millimhos. The optimum source conductance is 32.0 millimhos. Fig. 12 shows the F-vs-x plot obtained from the data used in Figs. 10 and 11. The intercept and slope of the line drawn by use of these data yield $F_0 = 9.9$ and $R_n = 117$ ohms. The curves superposed on the experimental data in Figs. 10 and 11 were computed from (13) by using the values of the noise parameters F_0, G_0, B_0, and R_n determined above.

5. Definition

Effective Input Noise Temperature (of a Two-Port Transducer). The input-termination *noise temperature* which, when the input termination is connected to a noise-free equivalent of the transducer, would result in the same output noise power as that of the actual transducer connected to a noise-free input termination.

Note 1: For heterodyne systems there is, in principle, more than one output frequency corresponding to a single input frequency, and vice versa; an *Effective Input Noise Temperature* is defined for each pair of corresponding frequencies.

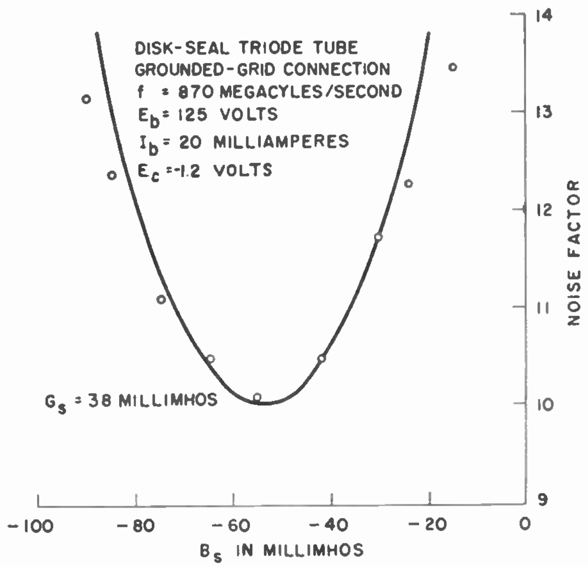


Fig. 10—Noise factor vs source susceptance.

Note 2: The *Effective Input Noise Temperature* depends upon the impedance of the input termination.

Note 3: The *Effective Input Noise Temperature*, T_e, in degrees Kelvin is related to the *Noise Factor* F by the equation

$$T_e = 290(F - 1).$$

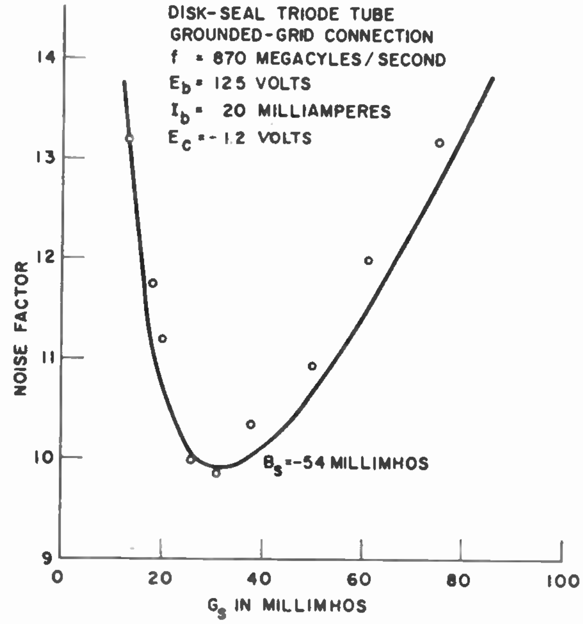


Fig. 11—Noise factor vs source conductance.

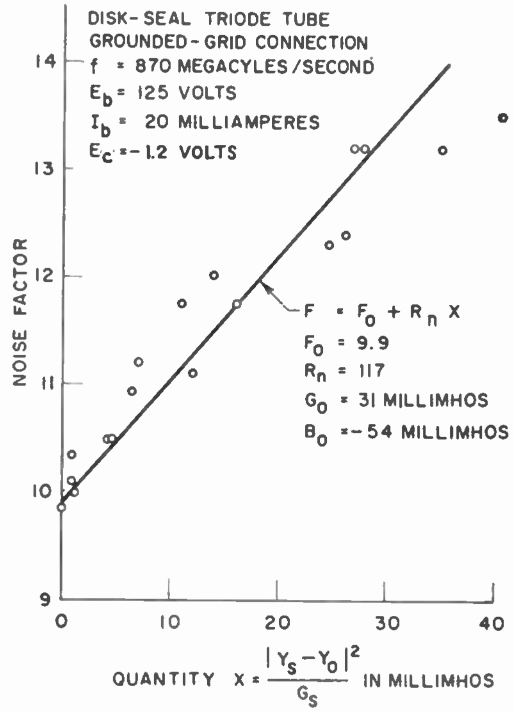


Fig. 12—Noise factor vs quantity x.

Representation of Noise in Linear Twoports*

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The compilation of standard methods of test often requires theoretical concepts that are not widely known nor readily available in the literature. While theoretical expositions are not properly part of a Standard they are necessary for its understanding and it seems desirable to make them easily accessible by simultaneous publication with the Standard. In such cases a technical committee report provides a convenient means of fulfilling this function.

The Standards Committee has decided to publish this technical committee report immediately following "Standards on Methods of Measuring Noise in Linear Twoports." A copy of this paper will be attached to all reprints of the Standards.

Summary—**This is a tutorial paper, written by the Subcommittee on Noise, IRE 7.9, to provide the theoretical background for some of the IRE Standards on Methods of Measuring Noise in Electron Tubes. The general-circuit-parameter representation of a linear twoport with internal sources and the Fourier representations of stationary noise sources are reviewed. The relationship between spectral densities and mean-square fluctuations is given and the noise factor of the linear twoport is expressed in terms of the mean-square fluctuations of the source current and the internal noise sources. The noise current is then split into two components, one perfectly correlated and one uncorrelated with the noise voltage. Expressed in terms of the noise voltage and these components of the noise current, the noise factor is then shown to be a function of four parameters which are independent of the circuit external to the twoport.**

I. Introduction

One of the basic problems in communication engineering is the distortion of weak signals by the ever-present thermal noise and by the noise of the devices used to process such signals. The transducers performing signal processing such as amplification, frequency mixing, frequency shifting, etc., may usually be classified as twoports. Since weak signals have amplitudes small compared with, say, the grid bias voltage of a vacuum tube, or emitter bias current of a transistor, etc., the amplitudes of the excitations at the ports can be linearly related. Consequently, the description and measurement of noise that is presented in this paper, and which forms the basis for the "Methods of Test" described in this same issue,[1] can be restricted to linear noisy twoports. Even if inherently nonlinear characteristics are involved, as in mixing, etc., linear relations still exist among the signal input and output amplitudes, although these may not be associated with the same frequency. Image-frequency components may be eliminated by proper filtering, but a generalization of our results to cases in which image frequencies are present is not difficult.

The effect of the noise originating in a twoport when the signal passes through the twoport is characterized at any particular frequency by the (*spot*) *noise factor* (*noise figure*). Since this has meaning only if the source impedance used in obtaining the noise factor is also specified, the noise contribution of a twoport is often indicated by a minimum noise factor and the source impedance with which this minimum noise factor is achieved. However, the extent to which the noise factor depends upon the input source impedance, upon the amount of feedback, etc., can be indicated only by a more detailed representation of the linear twoport.[2,3]

As a basis for understanding the representation of networks containing (statistical) noise sources, we shall consider first the analysis of networks containing Fourier-transformable signal sources. We shall then describe noise in two ways: a) as a limit of Fourier-integral transforms, and b) as a limit of Fourier-series transforms. The reasons for the wider use of the latter description will be presented. With this background we shall be ready to show that the four noise parameters used in the standard "Methods of Test" completely characterize a noisy linear twoport.

II. Representations of Linear Twoports

The excitation at either port of a linear twoport can be completely described by a time-dependent voltage $v(t)$ and the time-dependent current $i(t)$. (For waveguide twoports, where no voltage or current can be identified uniquely, an equivalent voltage and current can always

* Original manuscript received by the IRE, September 15, 1958; revised manuscript received March 6, 1959.

[1] "IRE Standards on Methods of Measuring Noise in Linear Twoports, 1959," this issue, p. 60.

[2] H. Rothe and W. Dahlke, "Theory of noisy fourpoles," Proc. IRE, vol. 44, pp. 811–818; June, 1956. Also, "Theorie rauschender Vierpole," *Arch. elekt. Übertragung*, vol. 9, pp. 117–121; March, 1955.

[3] A. G. T. Becking, H. Groendijk, and K. S. Knol, "The noise factor of four-terminal networks," *Philips Res. Repts.*, vol. 10, pp. 349–357; October, 1955.

be used.) Let it be assumed that the voltage and current functions can be transformed from the time domain to the frequency domain, and that V and I stand for the Fourier transforms when the function is aperiodic and for the Fourier amplitudes when the function is periodic. The linearity of the twoport *without internal sources* then allows an impedance representation:

$$\begin{aligned} V_1 &= Z_{11}I_1 + Z_{12}I_2 \\ V_2 &= Z_{21}I_1 + Z_{22}I_2. \end{aligned} \tag{1}$$

The subscripts 1 and 2 refer to the input and output ports, respectively, and the coefficients Z_{jk} are, in general functions of frequency. The currents are defined to be positive if the flow is into the network as shown in Fig. 1.

If the twoport contains internal sources, then (1) and the equivalent circuit must be modified. By a generalization of Thévenin's theorem, the twoport may be separated into a source-free network and two voltage generators, one in series with the input port and one in series with the output port. If the time-dependent functions $e_1(t)$ and $e_2(t)$ describing these equivalent generators can be transformed to functions of frequency E_1 and E_2, respectively, the impedance representation of the linear twoport *with internal sources* becomes

$$\begin{aligned} V_1 &= Z_{11}I_1 + Z_{12}I_2 + E_1 \\ V_2 &= Z_{21}I_1 + Z_{22}I_2 + E_2 \end{aligned} \tag{2}$$

and the equivalent network is that shown in Fig. 2.

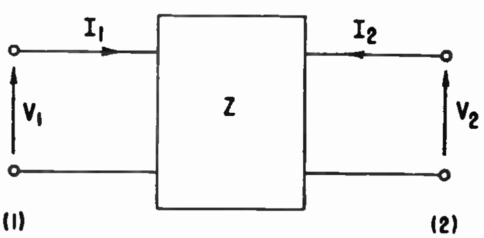

Fig. 1—Sign convention for impedance representation of linear twoport.

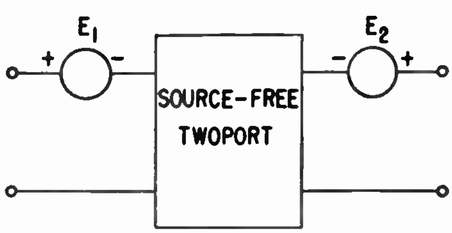

Fig. 2—Separation of twoport with internal sources into a source-free twoport and external voltage generators.

For the purpose of the analysis to follow, it should be emphasized again that such equations in general characterize the behavior of the twoport as a function of frequency. For practical purposes, it should be pointed out that in most cases the impedance parameters of a linear twoport can be measured at a particular frequency by applying sinusoidal voltages that produce outputs large compared with those caused by the internal sources.

The impedance representation of a linear twoport with internal sources has been reviewed because of its familiarity. However, it is well known that other representations, each leading to a different separation of the internal sources from the twoport, are possible. A particularly convenient one for the study of noise is the general-circuit-parameter representation[4]

$$\begin{aligned} V_1 &= AV_2 + BI_2 + E \\ I_1 &= CV_2 + DI_2 + I \end{aligned} \tag{3}$$

where E and I are again functions of frequency which are the Fourier transforms of the time-dependent functions $e(t)$ and $i(t)$ describing the internal sources.

As shown in Fig. 3, the internal sources are now represented by a source of voltage acting in series with the input voltage and a source of current flowing in parallel with the input current. It will be seen that this particular representation of the internal sources leads to four noise parameters that can easily be derived from single-frequency measurements of the twoport noise factor as a function of input mismatch. It has the further advantage that such properties of the twoport as gain and input conductance do not enter into the noise factor expression in terms of these four parameters.

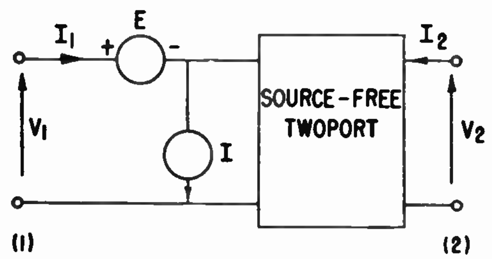

Fig. 3—Separation of twoport with internal sources into a source-free twoport and external input current and voltage sources.

III. Representations of Stationary Noise Sources

When a linear twoport contains stationary internal noise sources, the frequency functions E and I in (3) cannot be found by conventional Fourier methods from the time functions $e(t)$ and $i(t)$, since these are now random functions extending over all time and have infinite energy content. Two alternatives are possible. One may consider the substitute functions

$$\begin{aligned} e(t, T) &= e(t) \quad \text{for } -\frac{T}{2} < t < \frac{T}{2} \\ &= 0 \quad \text{for } \frac{T}{2} < |t| \\ i(t, T) &= i(t) \quad \text{for } -\frac{T}{2} < t < \frac{T}{2} \\ &= 0 \quad \text{for } \frac{T}{2} < |t| \end{aligned} \tag{4}$$

[4] E. Guillemin, "Communication Networks," John Wiley and Sons, New York, N. Y., vol. 2, p. 138; 1935.

where T is some long but finite time interval, and use the Fourier transforms

$$E(\omega, T) = \frac{1}{2\pi}\int_{-T/2}^{T/2} e(t, T)e^{-j\omega t}dt$$
$$I(\omega, T) = \frac{1}{2\pi}\int_{-T/2}^{T/2} i(t, T)e^{-j\omega t}dt. \tag{5}$$

Alternatively, one may construct periodic functions,

$$e(t, T) = e(t) \quad \text{for } -\frac{T}{2} < t < \frac{T}{2}$$
$$e(t + nT, T) = e(t, T) \quad \text{with } n \text{ an integer,}$$
$$i(t, T) = i(t) \quad \text{for } -\frac{T}{2} < t < \frac{T}{2}$$
$$i(t + nT, T) = i(t, T) \quad \text{with } n \text{ an integer,} \tag{6}$$

and expand these functions into Fourier series with amplitudes

$$E_m(\omega, T) = \frac{1}{T}\int_{-T/2}^{T/2} e(t, T)e^{-j\omega t}dt$$
$$I_m(\omega, T) = \frac{1}{T}\int_{-T/2}^{T/2} i(t, T)e^{-j\omega t}dt \tag{7}$$

where $\omega = m2\pi/T$ with m an integer, and T is again finite. In either case, the substitute functions can be made to approach the actual functions as closely as desired by making the interval T larger and larger.

In either the Fourier-integral or the Fourier-series approach, we consider a set of substitute functions obtained in principle from a series of measurements a) on an ensemble of systems with identical statistical properties or b) on one and the same system at successive time intervals sufficiently separated so that no statistical correlation exists.[5]

Since noise is a statistical process, we are in general interested in statistical averages[6] rather than the exact details of any particular noise function. These averages are important since they relate to physically measurable stationary quantities. For example, in the Fourier-integral approach the spectral density of the noise-voltage excitation is defined by

$$W_e(\omega) = \lim_{T\to\infty} \frac{\overline{|E(\omega, T)|^2}}{2\pi T} \tag{8}$$

where the bar indicates an arithmetic average of the Fourier transforms of an ensemble of functions described by (4). This spectral density is proportional to the noise power (in a narrow frequency band[7] around a given frequency) in a resistor across which the fluctuating voltage $e(t)$ appears.

Unfortunately, the notation developed in the literature for dealing with spectral densities is unwieldy, since the quantity with which the density is associated is relegated to a subscript. This is one of the reasons why researchers on noise have tended to use the historically older Fourier-series approach. The use of spectral densities is preferred in rigorous mathematical treatments of noise and in questions involving definitions of noise processes, since this assures that all points on the frequency axis within a narrow range are equivalent. For practical purposes, however, the noise amplitudes that are associated with discrete frequencies in the Fourier-series method can be as closely distributed as desired by choosing the time interval T sufficiently large.

IV. Relationship of Spectral Densities and Fourier Amplitudes to Mean-Square Fluctuations

If there is an open-circuit noise voltage $v(t)$ across a terminal pair, the mean-square value $\overline{v^2(t)}$ is related to the spectral density $W_v(\omega)$ and to the Fourier amplitudes $V_m(\omega)$ as follows:

$$\overline{v^2(t)} = \lim_{T\to\infty} \frac{1}{T}\int_{-T/2}^{T/2} v^2(t)dt$$
$$= \int_{-\infty}^{+\infty} W_v(\omega)d\omega = \sum_{m=-\infty}^{\infty} \overline{|V_m(\omega)|^2} \tag{9}$$

where the amplitudes V_m are now those obtained as $T\to\infty$.

Suppose that the stationary time function $v(t)$ is passed through an ideal band-pass filter with the narrow bandwidth $\Delta f = \Delta\omega/2\pi$ centered at a frequency $f^0 = \omega^0/2\pi$. The mean-square value of the voltage appearing at the filter output, usually denoted by the symbol $\overline{v^2}$ and called the mean-square fluctuation of $v(t)$ within the frequency interval Δf" is

$$\overline{v^2} = 4\pi\Delta f W_v(\omega_0) = 2\overline{|V_m|^2}. \tag{10}$$

This equation relates the mean-square fluctuations, the spectral density, and the Fourier amplitude. If the filter is narrow-band but not ideal, the quantity Δf is the noise bandwidth.[1] The factor 2 in (10) arises because $W_v(\omega)$ and V_m have been defined on the negative as well as the positive frequency axis.

Eq. (10) shows that spectral densities and mean-square fluctuations are equivalent. Although mathematical limits are involved in the definitions, these

[5] The equivalence of the ensemble and time average is known as the ergodic hypothesis. An interesting discussion of the implications of this hypothesis can be found in Born, "Natural Philosophy of Cause and Chance," Oxford University Press, New York, N. Y., p. 62; 1948. See also W. B. Davenport, Jr. and W. L. Root, "An Introduction to the Theory of Random Signals and Noise," McGraw-Hill Book Company, Inc., New York, N. Y., p. 66 ff.; 1958.

[6] W. R. Bennett, "Methods of solving noise problems," Proc. IRE, vol. 44, pp. 609–637; May, 1956.

[7] A frequency interval Δf is called narrow if the physical quantities under consideration are independent of frequency throughout the interval, and if $\Delta f \ll f_0$, where f_0 is the center frequency.

quantities can be measured to any desired degree of accuracy. For example, if one measures the power flowing into a termination connected to the terminals with which $v(t)$ is associated, one measures essentially $W_v(\omega_0)$, provided that the following requirements are met:

1) The termination is known and has a high impedance so that the voltage across the terminals remains essentially the open-circuit voltage $v(t)$, or the internal impedance associated with the two terminals is also known so that any change in voltage can be computed. The noise voltage contributed by the termination must either be negligible, or its statistical properties must be known so that its effect can be taken into account.

2) The termination is fed through a filter with a pass band sufficiently narrow so that the spectral density and internal impedance are essentially constant throughout the band; or the termination is itself a resonant circuit with a high Q corresponding to a sufficiently narrow bandwidth.

3) The power measurement is made over a period of time that is long compared to the reciprocal bandwidth, $1/\Delta f$, of the filter. In this way the power-measuring device takes an average over many long time intervals that is equivalent to an ensemble average.

In the description of noise transformations by linear twoports that follows, the mean-square fluctuations will be used. Mean-square current fluctuations can be related to the Fourier amplitudes $I_m(\omega)$ by a procedure similar to the one just described. Since fluctuations of the cross products of statistical functions will also be used it should be noted that

$$\lim_{T\to\infty} \frac{1}{T}\int_{-T/2}^{T/2} v(t, T)i(t, T)dt = \sum_{m=-\infty}^{\infty} \overline{V_m I_m^*} \qquad (11)$$

where $i(t)$ is a noise current and V_m and I_m are again the amplitudes obtained as $T\to\infty$.

We may interpret the real part of the complex quantity $2\overline{V_m I_m^*}$ as the contribution to the average of vi from the frequency increment $\Delta f = 1/T$ at the angular frequency $\omega = m\Delta\omega$. However, the imaginary part of $V_m I_m^*$ also contains phase information that has to be used in noise computations. We shall, therefore, use the expression

$$\overline{vi^*} = 2\overline{V_m I_m^*} \qquad \text{for } m > 0, \omega > 0 \qquad (12)$$

as the complex cross-product fluctuations of $v(t)$ and $i(t)$ in the frequency increment Δf. They are related to the cross-spectral density by

$$\overline{vi^*} = 4\pi\Delta f W_{iv}(\omega) \qquad (13)$$

where

$$W_{iv}(\omega) = \lim_{T\to\infty} \frac{\overline{I^*(\omega, T)V(\omega, T)}}{2\pi T}. \qquad (14)$$

There are several ways of specifying the fluctuations (or the spectral densities) that characterize the internal noise sources. A mean-square voltage fluctuation can be given directly in units of volt² second. It is often convenient, however, to express this quantity in resistance units by using the Nyquist formula, which gives the mean-square fluctuation of the open-circuit noise voltage of a resistor R at temperature T as

$$\overline{e^2} = 4kTR\Delta f$$

where k is Boltzmann's constant. For any mean-square voltage fluctuation $\overline{e^2}$ within the frequency interval Δf, one defines the equivalent noise resistance R_n as

$$R_n = \frac{\overline{e^2}}{4kT_0\Delta f} \qquad (15)$$

where T_0 is the standard temperature, 290°K. The use of R_n has the advantage that a direct comparison can be made between the noise due to internal sources and the noise of resistances generally present in the circuit. Note that R_n is not the resistance of a physical resistor in the network in which e is a physical noise voltage and therefore does not appear as a resistance in the equivalent circuit of the network.

In a similar manner, a mean-square current fluctuation can be represented in terms of an equivalent noise conductance G_n which is defined by

$$G_n = \frac{\overline{i^2}}{4kT_0\Delta f}. \qquad (16)$$

V. Noise Transformations by Linear Twoports

The statistical averages discussed in Sections III and IV will now be used to describe the internal noise sources of a linear twoport. We may consider functions of the type given by (6) and represent the noise sources E and I in (3) by the Fourier amplitudes $E_m(\omega, T)$ and $I_m(\omega, T)$. Thus, if the circuit shown in Fig. 3 represents a separation of noise sources from a linear twoport, the noise-free circuit is preceded by a noise network. Since a noise-free network connected to a terminal pair does not change the signal-to-noise ratio (noise factor evaluated at that terminal pair) the noise factor of the over-all network is equal to that of the noise network.

To derive the noise factor, let us connect the noise network to a statistical source comprising an internal admittance Y_s and a current source again represented by a Fourier amplitude $I_s(\omega, T)$. The network to be used for the noise-factor computation is then as shown in Fig. 4. By definition,[8] *the spot noise factor (figure)* of a network at a specified frequency is given by the ratio

[8] "IRE Standards on Electron Tubes: Definitions of Terms, 1957," Proc. IRE, vol. 45, pp. 983–1010; July, 1957.

of 1) the total output noise power per unit bandwidth available at the output port to 2) that portion of 1) engendered by the input termination at the standard temperature T_0.

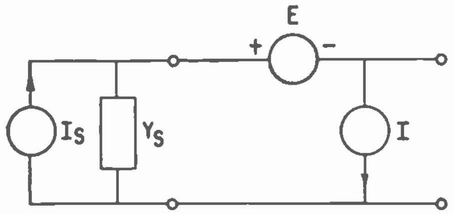

Fig. 4—Truncated network for noise-factor computation.

Now the total short-circuit noise current at the output of the network shown in Fig. 4 can be represented in terms of Fourier amplitudes by

$$I_s(\omega, T) + I_m(\omega, T) + Y_s E_m(\omega, T).$$

Let us assume that the internal noise of the twoport and the noise from the source are uncorrelated. If we then square the total short-circuit noise-current Fourier amplitudes, take ensemble averages and use equations similar to (10) and (12) to introduce mean-square fluctuations, we obtain a mean-square current fluctuation

$$\overline{i_s^2} + \overline{|i + Y_s e|^2} = \overline{i_s^2} + \overline{i^2} + |Y_s|^2 \overline{e^2} + Y_s^* \overline{ie^*} + Y_s \overline{i^* e} \tag{17}$$

to which the total output noise power is proportional. It should be noted here that when e and i are complex, the symbols $\overline{e^2}$ and $\overline{i^2}$ denote, by convention, $\overline{|e|^2}$ and $\overline{|i|^2}$. Since the noise power due to the source alone is proportional to $\overline{i_s^2}$, the noise factor becomes

$$F = 1 + \frac{\overline{|i + Y_s e|^2}}{\overline{i_s^2}}. \tag{18}$$

In the denominator, the mean-square source noise current is related to the source conductance G_s by the Nyquist formula

$$\overline{i_s^2} = 4kT_0 G_s \Delta f. \tag{19}$$

In the numerator of (18), the four real variables involved in $\overline{i^2}$, $\overline{e^2}$ and $\overline{ie^*}$, where e^* is the complex conjugate of e, describe the internal noise sources. If the Fourier transforms, (5), are used to characterize the noise, the self-spectral densities of noise current and voltage and the cross-spectral density of these quantities would have to be specified.

Before proceeding, let us review in general terms the methods employed and the conclusions reached. The representation of the internal noise sources of a noisy linear twoport by a voltage generator and a current generator lumps the effect of all internal noise sources into the two generators. A complete specification of these generators is thus equivalent to a complete description of the internal sources, as far as their contribution to terminal voltages and currents is concerned. Since the sources under consideration are noise sources, their description is confined to the methods applicable to noise. The extent and detail of the description depends on the amount of detail envisioned in the analysis. Since, in the case of the noise factor, only the mean-square fluctuations of output currents are sought, the specification of self and cross-product fluctuations of the generator voltages and currents is adequate.

The expression for the noise factor given in (18) can be simplified if the noise current is split into two components, one perfectly correlated and one uncorrelated with the noise voltage. The uncorrelated noise current, designated by i_u, is defined at each frequency by the relations

$$\overline{ei_u^*} = 0 \tag{20}$$

$$\overline{(i - i_u)i_u^*} = 0. \tag{21}$$

The correlated noise current, $i - i_u$, can be written as $Y_\gamma e$, where the complex constant $Y_\gamma = G_\gamma + jB_\gamma$ has the dimensions of an admittance and is called the correlation admittance. The cross-product fluctuation $\overline{ei^*}$ may then be written

$$\overline{ei^*} = \overline{e(i - i_u)^*} = Y_\gamma^* \overline{e^2}. \tag{22}$$

The noise-voltage fluctuation can be expressed in terms of an equivalent noise resistance R_n as

$$\overline{e^2} = 4kT_0 R_n \Delta f \tag{23}$$

and the uncorrelated noise-current fluctuation in terms of an equivalent noise conductance G_u as

$$\overline{i_u^2} = 4kT_0 G_u \Delta f. \tag{24}$$

The fluctuations of the total noise current are then

$$\begin{aligned} \overline{i^2} &= \overline{|i - i_u|^2} + \overline{i_u^2} \\ &= 4kT_0[|Y_\gamma|^2 R_n + G_u]\Delta f. \end{aligned} \tag{25}$$

From (18)–(24), the formula for the noise factor becomes

$$F = 1 + \frac{1}{4kT_0 G_s \Delta f}\left[\overline{i_u^2} + |Y_s + Y_\gamma|^2 \overline{e^2}\right] \tag{26}$$

$$= 1 + \frac{G_u}{G_s} + \frac{R_n}{G_s}\left[(G_s + G_\gamma)^2 + (B_s + B_\gamma)^2\right]. \tag{27}$$

Thus, the noise factor is a function of the four parameters G_u, R_n, G_γ and B_γ. These depend, in general, upon the operating point and operating frequency of the twoport, but not upon the external circuitry. In a vacuum tube triode the correlation susceptance B_γ is negligibly small at frequencies such that transit times are small compared to the period. Also, G_u and G_γ are vanishingly small when there is little grid loading. Thus, tube noise at low frequencies is adequately characterized by the single nonzero constant R_n. Tube noise at high frequencies and transistor noise at all frequencies have no such simple representation.

Since the noise factor is an explicit function of the source conductance and susceptance, it can readily be shown that the noise factor has an optimum (minimum) value at some optimum source admittance $Y_o = G_o + jB_o$ where

$$G_o = \left[\frac{G_u + R_n G_\gamma^2}{R_n}\right]^{1/2} \tag{28}$$

$$B_o = -B_\gamma \tag{29}$$

and the value of this minimum noise factor is

$$F_o = 1 + 2R_n(G_\gamma + G_o). \tag{30}$$

In terms of G_o, B_o and F_o, the noise factor for any arbitrary source impedance then becomes

$$F = F_o + \frac{R_n}{G_s}\left[(G_s - G_o)^2 + (B_s - B_o)^2\right]. \tag{31}$$

Eq. (31) shows that the four real parameters F_o, G_o, B_o and R_n give the noise factor of a twoport for every input termination of the twoport.[9] As shown in the methods of test that are published in this issue, a measurement of the minimum noise factor and of the source admittance Y_o with which F_o is achieved gives the first three parameters. The parameter R_n can be computed from an additional measurement of the noise factor for a source admittance Y_s other than Y_o.

From the given values, F_o, G_o, B_o and R_n, one can compute, if desired, the noise fluctuations $\overline{i^2}$, $\overline{e^2}$, and $\overline{ei^*}$ (or the corresponding spectral densities). To do this, one uses (28) to (30) to find Y_γ and G_u. From these one evaluates $\overline{e^2}$, $\overline{i^2}$, and $\overline{ei^*}$ from (23), (25), and (22). The fluctuations of any terminal voltage or current of the twoport produced with a given source and load can then be evaluated from the known coefficients, (A, B, C, D) of (3), and the known noise fluctuations. Thus, the noise in a twoport is completely characterized (with regard to the fluctuations or the spectral densities at the terminals) by the noise fluctuations $\overline{e^2}$, $\overline{i^2}$ and $\overline{ei^*}$, or alternately by the four noise parameters F_o, G_o, B_o and R_n.

[9] Reasoning similar to that leading to (31) can be carried out in a dual representation, where impedances and admittances are interchanged. An equation similar to (31) results, again involving four noise parameters, some of which are different from the ones used here.

VI. Conclusion

The preceding discussion showed that, with limited objectives, the noise in a linear twoport can be characterized adequately by a limited number of parameters. Thus, if one seeks only information concerning the mean-square fluctuations or the spectral densities of currents or voltages into or across the ports of a linear twoport at a particular frequency and for arbitrary circuit connections of the twoport, it is sufficient to specify two mean-square fluctuations and one product fluctuation or two spectral densities and one cross-spectral density. This involves the specification of four real numbers. If a band of frequencies is being considered, these quantities have to be given as functions of frequency unless they are approximately constant over the band.

Different separations of internal noise sources will lead, in general, to different frequency dependences of the resulting fluctuations or spectral densities. Thus, a particular separation of the noise sources may be preferable to another separation if it is found that fluctuations (spectral densities) are less frequency-dependent in the band. In particular, if available information about the physics of the noise in a particular device suggests introducing, *inside the twoport*, appropriate noise generators of fluctuations (spectral densities) with no frequency dependence, or with a simple frequency dependence, it may be more advantageous to specify the device in terms of these physically suggestive generators.

However, usually one resorts to the characterizations of noise presented in this paper, since they have the advantage that they do not require any knowledge of the physics of the internal noise. Furthermore, noise-factor measurements performed on the twoport yield (more or less) directly the noise parameters of the general circuit representation, namely the fluctuations (spectral densities) of the voltage and current generators attached to the input of the noise-free equivalent of the twoport. This fact recommends the use of the noise parameters natural to the general-circuit-parameter representation.

Multipole Representation for an Equivalent Cardiac Generator*

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Summary—**One facet of electrocardiography has traditionally dealt with the relationship between body surface potentials and an equivalent generator that could have produced them. A key problem facing investigators today is whether a single fixed-location dipole is a suitable equivalent generator. The equivalent cardiac generator developed here provides a means for resolving this question. It consists of a collection of multipoles which, if placed at a point in a homogeneous isotropic conductor shaped like the human body, will give rise to the same potential distribution on the surface as that observed on the body itself. The multipole components can be evaluated by performing appropriate integrations of the potential over the body surface.**

Introduction

An electrocardiogram is a display, with respect to time, of the potential difference between two points on the body or between a body point and a suitable reference terminal. The heart is the dominant source of these potentials. A region of the excitable muscle tissues that comprise the heart, called the pacemaker, acts as a free running pulse generator which produces a current impulse for each heartbeat. Currents are carried through the walls of the heart along an electrical conduction system and cause the surrounding muscle to contract and relax giving rise to the pumping action.

Since body tissues conduct electricity, the currents which trigger the heartbeat continue through the body and give rise to potential differences on the surface which can be measured and recorded by special voltmeters called electrocardiographs. When portions of the heart muscle are damaged during a "heart attack," the normal flow of electricity is changed, resulting in changes in the recorded electrocardiograms. Diagnoses are made on the basis of comparison with well-catalogued normal and abnormal waveforms.

There are an infinite number of electrocardiograms which may be recorded on an individual. To get the complete story we would have to use all of them, *i.e.*, obtain a potential distribution over the entire body surface as a function of time—an impractical procedure. Physicians do use as many as twelve waveforms for diagnosis. The hope is that considerable redundancy is present and the number of waveforms to be studied can be reduced substantially.

This reduction could be accomplished if the potential distribution were identified with a simple (fictitious) electrical source in the body. The information available could then be expressed in terms of the parameters of this equivalent generator for the heart. Furthermore, working with the equivalent generator instead of surface potentials could eliminate variations in observed waveforms caused by differences among individuals in body build, heart position, and other factors not of diagnostic significance. Hence the equivalent cardiac generator is a useful concept because, once determined, it can indicate how much information is available, how to extract it, and how to present it.

Most work in electrocardiography, dating back to the original contributions of Einthoven,[1] has centered around the fixed-location dipole of variable orientation and magnitude as the candidate for the equivalent generator. If the dipole is a satisfactory equivalent source, then only three quantities—the components of a vector—are sufficient to describe it at each instant of time. This fact provides the basis for vectorcardiography, in which the locus of the terminal points of this vector is displayed.

There is evidence to indicate that at least in certain cases the dipole may be a very good approximation to the equivalent generator.[2–4] However, the evidence is far from conclusive, and many feel that the dipole source is inadequate, especially in the case of abnormals, *i.e.*, people with heart disease. This question of the suitability of the dipole equivalent generator is a paramount one in electrocardiographic research today. A tool for resolving it is provided here through the development of an exact equivalent cardiac generator.

Equivalent Generator

The human body can be considered, for the present purposes, to be an inhomogeneous conducting medium surrounded by an insulating medium (air). Potential

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[1] W. Einthoven, G. Fahr, and A. de Waart, "Ueber die Richtung und die manifeste Groesse der Potentialschwankungen in menschlichen Herzen und ueber den Einfluss der Herzlage auf die Form des Elektrokardiogramms," (in English), *Am. Heart J.*, vol. 40, p. 163; August, 1950.

[2] O. H. Schmitt, R. B. Levine, and J. Dahl, "Electrocardiographic mirror pattern studies, Parts I, II, III," *Am. Heart J.*, vol. 45, pp. 416, 500, 655; March, April, May, 1953.

[3] E. Frank, "Spread of currents in volume conductors of finite extent," *Ann. N. Y. Acad. Sciences*, vol. 65, pp. 980–1002; August, 1957.

[4] H. C. Burger and J. B. van Milaan, "Hearts vector and leads; Parts I, II, III," *Brit. Heart J.*, vol. 8, pp. 157–161, September, 1946; vol. 9, pp. 154–160, September, 1947; vol. 10, pp. 229–233, October, 1948.

differences exist on the surface of the body. An equivalent generator is a set of sources which, if located in a conductor with the same shape as the body, would give rise to the identical potential distribution on the surface.

It is important to note that the equivalent generator depends on the type of conductor chosen. The simplest to analyze is the homogeneous isotropic conductor, so this one will be used. Now the problem is to find a set of sources which, if located in a homogeneous isotropic conductor with the same shape as the human body, will give rise to potential differences on the surface identical to those observed on the body itself. These sources constitute the equivalent cardiac generator.

There is no unique solution to the problem; indeed there are an infinite number of solutions. Since the dipole appears to be a first approximation, it would seem fruitful to relate the equivalent generator in some way to a fixed-location dipole. There is one type of source which immediately permits separation and comparison of dipole and nondipole terms. This source consists of current singularities, or multipoles, at a point. Since the fixed-location multipole source is most convenient for testing the validity of the dipole approximation, it is the one which will be developed below.

Gabor and Nelson, in a classical paper,[5] also develop an exact equivalent cardiac generator. They consider a multiple dipole source consisting of a single ("resultant") dipole as a first approximation, two dipoles as a second approximation, etc. The n dipole *locations*, as well as their magnitudes and orientations, will in general vary with time, whereas in the multipole generator the location of all components is fixed at a single point. Therefore, the multiple dipole generator provides a less convenient tool for evaluating the single fixed-location dipole approximation. Gabor and Nelson's resultant dipole is identical to the dipole term in the multipole expansion. Yeh and Martinek[6] have treated a multipole source at the center of a sphere and have developed formulas for approximating the multipole components from measurements of a given number of potential differences. The present analysis permits an exact evaluation of all multipole components at an arbitrary point inside a conductor of arbitrary shape.

Mathematical Development

Consider an isolated homogeneous conductor with a given potential distribution on its surface. Let

V = potential inside and on the surface of the conductor,

g = conductivity of the conductor,

J = current density inside the conductor.

Then

$$J = -g\nabla V. \tag{1}$$

The divergence of J is everywhere 0 except where there is a current source. (The term "source" is used in this paper in a general sense to include the concept of a sink. A sink occurs wherever the source distribution is negative.) Let us define the current source distribution by J_v, where

$$J_v = \nabla \cdot J. \tag{2}$$

Then

$$\nabla^2 V = -\frac{1}{g} J_v. \tag{3}$$

If $dS = n\,dS$ = vector element of area of the boundary of the conductor, the boundary condition becomes

$$\frac{\partial V}{\partial n} = 0. \tag{4}$$

Our problem then is, given V on the surface, to find a J_v such that (3) and (4) are satisfied.

Let τ be the volume of the conductor, S its surface, and ρ the distance from an element of volume or area at (r, θ, ϕ) to an arbitrary fixed point $P\,(r', \theta', \phi')$ lying outside the conductor. Then from Green's theorem[7,8]

$$\int_\tau \left[\frac{1}{\rho}\nabla^2 V - V\nabla^2\left(\frac{1}{\rho}\right)\right] d\tau = \int_S \left[\frac{1}{\rho}\frac{\partial V}{\partial n} - V\frac{\partial}{\partial n}\left(\frac{1}{\rho}\right)\right] dS. \tag{5}$$

Furthermore, from the solution of Poisson's equation,

$$V'(P) = \frac{1}{4\pi g}\int_\tau \frac{J_v}{\rho}\, d\tau, \tag{6}$$

where $V'(P)$ is the potential that would exist at P if the source distribution J_v were in an infinite homogeneous medium of conductivity g. Substitute (3), (4), and (6) into (5) and observe that $\nabla^2(1/\rho)$ is zero everywhere inside the volume of integration. The remarkable result is

$$V'(P) = \frac{1}{4\pi g}\int_S gV\,\nabla\left(\frac{1}{\rho}\right)\cdot dS, \tag{7}$$

which states that the potential that would exist outside the conductor if the sources were in an unbounded medium is the same as the potential in an unbounded medium, caused by a double layer of moment gV spread on the boundary of the original conductor.

The inverse distance from the point (r, θ, ϕ) to (r', θ', ϕ') can be expanded as follows for $r' > r$:

[5] D. Gabor and C. V. Nelson, "Determination of the resultant dipole of the heart from measurements on the body surface," *J. Appl. Phys.*, vol. 25, pp. 413–416; April, 1954.

[6] G. C. K. Yeh and J. Martinek, "Comparison of surface potentials due to several singularity representations of the human heart," *Bull. Math. Biophys.*, vol. 19, pp. 293–308; December, 1957.

[7] W. P. Smythe, "Static and Dynamic Electricity," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 48–58 and 129–138; 1950.

[8] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 160–193; 1941.

$$\frac{1}{\rho} = \frac{1}{r'} \sum_{n=0}^{\infty} \sum_{m=0}^{n} \left(\frac{r}{r'}\right)^n (2 - \delta_m^0) \frac{(n-m)!}{(n+m)!} \cdot P_n^m(\cos\theta) P_n^m(\cos\theta') \cos m(\phi - \phi'), \tag{8}$$

where $\delta_m^0 = 1$ for $m=0$, 0 for $m \neq 0$, and P_n^m are associated Legendre functions.[7,9] If (8) is substituted in (6) we obtain:

$$V'(r', \theta', \phi') = \frac{1}{4\pi g r'} \sum_{n=0}^{\infty} \sum_{m=0}^{n} \left(\frac{1}{r'}\right)^n \cdot P_n^m(\cos\theta')(a_{nm}\cos m\phi' + b_{nm}\sin m\phi'), \tag{9}$$

where

$$\left\langle \frac{a_{nm}}{b_{nm}} \right\rangle = (2 - \delta_m^0) \frac{(n-m)!}{(n+m)!} \int r^n P_n^m(\cos\theta) \cdot \left\langle \begin{matrix} \cos m\phi \\ \sin m\phi \end{matrix} \right\rangle J_v(r, \theta, \phi) d\tau. \tag{10}$$

Thus it is seen that the infinite medium potential does not uniquely determine the source distribution J_v. As a matter of fact, the significant parameters of J_v are the coefficients a_{nm} and b_{nm}. The expression for V' in (9) will converge everywhere outside the sphere containing all J_v, provided J_v is bounded.

Let us now substitute (8) in (7) and compare the result with (9):

$$V' = \frac{1}{4\pi r'} \sum_{n=0}^{\infty} \sum_{m=0}^{n} \left(\frac{1}{r'}\right)^n (2 - \delta_m^0) \frac{(n-m)!}{(n+m)!} P_n^m(\cos\theta') \int_S V \left(dS_r \frac{\partial}{\partial r} + \frac{1}{r} dS_\theta \frac{\partial}{\partial \theta} + \frac{1}{r\sin\theta} dS_\phi \frac{\partial}{\partial \phi} \right) \cdot r^n P_n^m(\cos\theta) \cos m(\phi - \phi'). \tag{11}$$

$$\left\langle \begin{matrix} a_{nm} \\ b_{nm} \end{matrix} \right\rangle = (2 - \delta_m^0) \frac{(n-m)!}{(n+m)!} g \int_S V(r, \theta, \phi) r^{n-1} \left\{ n P_n^m(\cos\theta) \left\langle \begin{matrix} \cos m\phi \\ \sin m\phi \end{matrix} \right\rangle dS_r + \left[\frac{m\cos\theta}{\sin\theta} P_n^m(\cos\theta) - P_n^{m+1}(\cos\theta) \right] \left\langle \begin{matrix} \cos m\phi \\ \sin m\phi \end{matrix} \right\rangle dS_\theta + \frac{m}{\sin\theta} P_n^m(\cos\theta) \left\langle \begin{matrix} -\sin m\phi \\ \cos m\phi \end{matrix} \right\rangle dS_\phi \right\}. \tag{12}$$

From the relations given in the Appendix an equivalent formulation in rectangular coordinates can be derived:

$$\left\langle \begin{matrix} a_{nm} \\ b_{nm} \end{matrix} \right\rangle = (2 - \delta_m^0) \frac{(n-m)!}{(n+m)!} g \int_S V r^{n-1} \left\{ \left[(n+m) P_{n-1}^m(\cos\theta) \left\langle \begin{matrix} \cos m\phi \\ \sin m\phi \end{matrix} \right\rangle \right] dS_z + \left[\frac{1}{2}(n+m)(n+m-1) P_{n-1}^{m-1}(\cos\theta) \left\langle \begin{matrix} \cos (m-1)\phi \\ \sin (m-1)\phi \end{matrix} \right\rangle - \frac{1}{2} P_{n-1}^{m+1}(\cos\theta) \left\langle \begin{matrix} \cos (m+1)\phi \\ \sin (m+1)\phi \end{matrix} \right\rangle \right] dS_x + \left[\frac{1}{2}(n+m)(n+m-1) P_{n-1}^{m-1}(\cos\theta) \left\langle \begin{matrix} -\sin (m-1)\phi \\ \cos (m-1)\phi \end{matrix} \right\rangle - \frac{1}{2} P_{n-1}^{m+1}(\cos\theta) \left\langle \begin{matrix} \sin (m+1)\phi \\ -\cos (m+1)\phi \end{matrix} \right\rangle \right] dS_y, \tag{13}$$

where (x, y, z) are the coordinates corresponding to (r, θ, ϕ).

MULTIPOLES

The expression for the potential given in (9) is an expansion in terms of multipoles.[8] The index n gives the order of the multipole. Thus $n=1$ is the dipole, $n=2$ the quadrupole, $n=3$ the octupole, etc. The index m gives the components of the multipole. Since from (10) b_{n0} is always 0, there are $(2n+1)$ components for the nth order multipole. Also,

$$a_{00} = \int J_v d\tau = 0, \tag{14}$$

since the net current source must be zero.

If the values for the associated Legendre functions given in the Appendix are substituted in (9) we obtain:

$$V' = \frac{1}{4\pi g(r')^2}(a_{10}\cos\theta + a_{11}\sin\theta\cos\phi + b_{11}\sin\theta\sin\phi) + \cdots.$$

Hence the coefficients a_{10}, a_{11}, and b_{11} are the familiar z, x, and y components of a dipole, respectively. Furthermore, from (13) and the Appendix,

[9] E. W. Hobson, "Spherical and Ellipsoidal Harmonics," Cambridge University Press, Cambridge, Eng.; 1931.

$$a_{10} = g \int V dS_z,$$

$$a_{11} = g \int V dS_x,$$

$$b_{11} = g \int V dS_y,$$

$$a_{20} = g \int V(z dS_z - x dS_x - y dS_y),$$

$$a_{21} = g \int V(z dS_x + x dS_z),$$

$$b_{21} = g \int V(z dS_y + y dS_z),$$

$$a_{22} = \frac{1}{2} g \int V(x dS_x - y dS_y),$$

$$b_{22} = \frac{1}{2} g \int V(y dS_x + x dS_y). \qquad (15)$$

Notice that the components of the dipole are independent of the choice of origin. This fact is explained when one considers that at a very large distance from the source, the dipole term will be the only one of significance. Furthermore, at a very large distance we cannot distinguish where in the volume we put the origin. If the origin is shifted so that

$$z' = z - z_0, \quad x' = x - x_0, \quad y' = y - y_0, \qquad (16)$$

the new quadrupole terms become:

$$a'_{20} = a_{20} - 2z_0 a_{10} + x_0 a_{11} + y_0 b_{11},$$

$$a'_{21} = a_{21} - z_0 a_{11} - x_0 a_{10},$$

$$b'_{21} = b_{21} - z_0 b_{11} - y_0 a_{10},$$

$$a'_{22} = a_{20} - \tfrac{1}{2} x_0 a_{11} + \tfrac{1}{2} y_0 b_{11},$$

$$b'_{22} = b_{22} - \tfrac{1}{2} y_0 a_{11} - \tfrac{1}{2} x_0 b_{11}. \qquad (17)$$

This procedure could be extended to higher-order multipoles. The result is that once the components for a particular origin are known, they can be calculated for any other origin. In practice we would like to find the origin where the relative contribution of the dipole term is a maximum.

Finally, it is necessary to investigate the convergence of the multipole expansion for the potential. If the sources are confined within a sphere whose center is at the origin, the multipole expansion will converge in the region outside this sphere. Consider another set of sources confined within a sphere of smaller radius. If the potentials everywhere outside the larger sphere are identical in the two cases, then the multipole expansions must be identical. But the expansion will now converge outside the smaller sphere. We thus conclude that the multipole expansion will converge outside the sphere containing the most compact bounded-source distribution which could give rise to the desired potential.

Let r max and r min be, respectively, the largest and smallest distances from the origin to the surface of the conductor. The expansions (9) and (11) will converge in the region $r' > r$ max. In order for the multipoles to constitute an equivalent generator, the expansion for the potential must also converge outside $r' < r$ min. If this condition is not met for any choice of origin, it means that a bounded distribution J_v, which will yield the desired V on the surface and which can be contained within a sphere lying entirely inside the conductor, does not exist. Conversely, if it is known that all the sources and surfaces of discontinuity in the original conductor can be contained within such a sphere, then it follows that the resulting multipole expansion will converge everywhere on the surface.

Discussion

We have found a collection of multipoles which, when located at a point inside an insulated homogeneous conductor, will give an arbitrary potential distribution on the surface. The multipole components depend only on the surface potential, the shape of the conductor, its conductivity, and the origin. If the conductor takes on the shape of the human body, and the potential distribution is equal to that found on the body surface, then the multipoles consitute an equivalent cardiac generator which can be evaluated from information available at the body surface.

While the dipole terms are independent of the location of the origin, the higher order poles are not. It is then possible to locate the origin to minimize the contribution of the higher order terms. The relative contribution to the potential of the terms for $n > 2$ is a direct measure of the "non-dipolarity" of the equivalent generator.

Strictly speaking, the multipoles constitute an equivalent generator only if the multipole expansion converges everywhere on the surface of the conductor. Even if it does not converge, the multipole expansion is of value in electrocardiography because it then will indicate immediately that the single fixed-location dipole hypothesis is a poor one. However, there is good reason to expect convergence. In the first place, the sources and significant inhomogeneities in the torso are localized in the region of the heart. Furthermore, there is the limited experimental evidence that the dipole is at least a first approximation.

The multipole expansion for the equivalent cardiac generator was developed in the course of studying electrocardiographic phenomena from a basic electrical standpoint. Considerable effort, both experimental and computational, is necessary to evaluate the terms in the expansion. Consequently, this technique is not intended to be applied clinically. Rather, it is a basic tool for further research in electrocardiography.

It is hoped, for example, that the question of the equivalent dipole generator can be resolved hereby. If the dipole is inadequate, then perhaps a dipole plus a simple non-dipole term will be useful. On the other hand, if many terms in the expansion are necessary, then attempts to work with a simple equivalent generator may have to be abandoned unless empirical diagnostic studies indicate otherwise.

Example

It is difficult to construct an example of the method of multipole expansion without either being trivial or getting involved in excessive computation. One can show without too much difficulty, for example, that the method gives the correct results for the sphere.[10] The following simple illustration is an even more special case.

If the volume under consideration is a sphere of radius a, then $dS_\theta = 0$, $dS_\phi = 0$, and

$$dS_r = a^2 \sin\theta \, d\phi \, d\theta. \tag{18}$$

Hence,

$$\begin{pmatrix} a_{nm} \\ b_{nm} \end{pmatrix} = (2 - \delta_m{}^0)\frac{(n-m)!}{(n+m)!} n g a^{n+1} \int_0^{2\pi}\int_0^{\pi} V(a,\theta,\phi) P_n{}^m(\cos\theta)\begin{pmatrix} \cos m\phi \\ \sin m\phi \end{pmatrix} \sin\theta \, d\theta \, d\phi. \tag{19}$$

Consider that the sphere is homogeneous and contains a centric dipole of moment p oriented in the z direction. Then

$$V(a,\theta,\phi) = \frac{3p\cos\theta}{4\pi g a^2} = \frac{3p}{4\pi g a^2} P_1{}^0(\cos\theta). \tag{20}$$

Let $\mu = \cos\theta$ and substitute (20) into (19):

$$\begin{pmatrix} a_{nm} \\ b_{nm} \end{pmatrix} = \frac{3p}{2} n a^{n-1} \begin{pmatrix} \delta_m{}^0 \\ 0 \end{pmatrix} \int_{-1}^{1} P_1{}^0(\mu) P_n{}^0(\mu) d\mu. \tag{21}$$

From the orthogonality relations for the Legendre polynomials, (21) reduces to

$$\begin{pmatrix} a_{nm} \\ b_{nm} \end{pmatrix} = p \begin{pmatrix} \delta_m{}^0 \\ 0 \end{pmatrix} \delta_n{}^1. \tag{22}$$

Therefore $a_{10} = p$ and all other terms are 0, which checks.

Appendix

To obtain (13), use must be made of the relations existing among the components of a vector in spherical and rectangular coordinates. These are:

$$\left.\begin{aligned} dS_r &= dS_z \cos\theta + dS_x \cos\phi \sin\theta + dS_y \sin\phi \sin\theta \\ dS_\theta &= -\,dS_z \sin\theta + dS_x \cos\phi\cos\theta + dS_y \sin\phi\cos\theta \\ dS_\phi &= -\,dS_x \sin\phi + dS_y \cos\phi \end{aligned}\right\}. \tag{23}$$

In addition, the following recursion relations are needed:

$$\begin{aligned} &P_{n-1}^{m+1}(\cos\theta) \\ &\quad = (m-n)\sin\theta P_n^m(\cos\theta) + \cos\theta P_n^{m+1}(\cos\theta). \end{aligned} \tag{24}$$

$$\begin{aligned} &\frac{2m}{\sin\theta} P_n^m(\cos\theta) \\ &\quad = P_{n-1}^{m+1}(\cos\theta) + (m+n-1)(m+n)P_{n-1}^{m-1}(\cos\theta). \end{aligned} \tag{25}$$

$$\begin{aligned} 2mP_{n+1}^m(\cos\theta) &= \sin\theta P_n^{m+1}(\cos\theta) \\ &\quad + (m+n)(m+n+1)\sin\theta P_n^{m-1}(\cos\theta). \end{aligned} \tag{26}$$

$$\begin{aligned} &(m+n)\sin\theta P_n^{m-1}(\cos\theta) \\ &\quad = P_{n+1}^m(\cos\theta) - \cos\theta P_n^m(\cos\theta). \end{aligned} \tag{27}$$

$$\begin{aligned} &(m-n-1)P_{n+1}^m(\cos\theta) + (2n+1)\cos\theta P_n^m(\cos\theta) \\ &\quad = (m+n)P_{n-1}^m(\cos\theta). \end{aligned} \tag{28}$$

Note also:

$$n(n-1)P_{n-1}^{-1}(\cos\theta) = -P_{n-1}^1(\cos\theta). \tag{29}$$

$$P_n^m(\cos\theta) = 0 \text{ for } m > n. \tag{30}$$

$$\frac{d}{d\theta} P_n^m(\cos\theta) = m\frac{\cos\theta}{\sin\theta} P_n^m(\cos\theta) - P_n^{m+1}(\cos\theta). \tag{31}$$

Finally, the following values for the first few associated Legendre functions are listed here for convenience.

$$\begin{aligned} P_0^0(\cos\theta) &= 1. \\ P_1^0(\cos\theta) &= \cos\theta. \\ P_1^1(\cos\theta) &= \sin\theta. \\ P_2^0(\cos\theta) &= \tfrac{1}{2}(3\cos^2\theta - 1) = \cos^2\theta - \tfrac{1}{2}\sin^2\theta. \\ P_2^1(\cos\theta) &= 3\cos\theta\sin\theta. \\ P_2^2(\cos\theta) &= 3\sin^2\theta. \end{aligned}$$

Acknowledgment

The author gratefully acknowledges the advice and encouragement of Dr. R. H. Okada, and also wishes to thank Dr. H. Schwan and Dr. B. Epstein for their criticism of the manuscript.

[10] D. B. Geselowitz, "Application of Potential Theory to the Study of an Equivalent Cardiac Generator," Ph.D. dissertation, University of Pennsylvania, Philadelphia, Pa.; 1958.

The Effect of a Cathode Impedance on the Frequency Stability of Linear Oscillators*

C. T. KOHN†

Summary—**High stability oscillators usually employ tubes having rather high values of transconductance. Excellent frequency stability as to valve capacitance and supply voltage changes results, but a change of the cathode impedance produces large frequency changes. Cathode impedance changes in a valve may reach 100 ohms during its lifetime because of the growth of an interface layer between the oxide coating and the cathode sleeve. In order to reduce the effect of the cathode impedance, the g_m value must be reduced considerably, the optimum being near $1/X_g$, where X_g is the grid-cathode reactance. The ratio of the anode/grid RF voltages should be made equal to the ratio of grid/anode capacitances, or equal to unity. Full advantage cannot always be taken of the recommended g_m value, as this makes the oscillator more susceptible to random capacitance variations and to changes of supply voltage, but a considerable reduction below customary values is possible. In this way, the long-term stability can be improved at least ten times.**

I. List of Symbols

A = circuit parameter (9)
B = circuit parameter (10)
C = capacitance in the grid-anode branch of tuned circuit in Clapp oscillator
C_a = effective anode-cathode valve capacitance (Section III)
C_g = effective grid-cathode valve capacitance (Section III)
C_{ga} = grid-anode valve capacitance
D = circuit parameter (11)
E = circuit parameter (12)
K = slope of the $\gamma^e = f(g)$ characteristic (29)
L = total inductance of tuned circuit
N_R = circuit parameter (16)
N_x = circuit parameter (17)
N_1 = circuit parameter (13)
N_3 = circuit parameter (14)
N_4 = circuit parameter (15)
Q = quality factor of the circuit external to valve, except Z_k (4)
R_k = series resistive component of Z_k
X = reactance of the grid-anode branch of the tuned circuit (3)
X_a = anode reactance $= -1/C_a\omega$
X_g = grid reactance $= -1/C_g\omega$
X_k = series reactive component of Z_k
X_1 = reactance of the grid-cathode branch of the tuned circuit ($L_1\omega$ or $-1/C_1\omega$)
X_2 = reactance of the anode-cathode branch of the tuned circuit ($L_2\omega$ or $-1/C_2\omega$)
Z = impedance of the grid-anode branch of the tuned circuit (3)
Z_k = cathode impedance (1)
Z_1 = impedance of the grid-cathode branch of the tuned circuit
Z_2 = impedance of the anode-cathode branch of the tuned circuit
f = frequency of oscillation
g = mutual conductance of valve at operating point
g_0 = circuit parameter (18)
ppm = parts in one million
r = resistance connected in series with the inductance L, representing all the circuit losses except R_k
Δ = increment of the quantity following Δ
Δf = change of oscillation frequency
α = circuit parameter (19)
β = circuit parameter (20)
γ^e = contribution to the grid-cathode capacitance due to the valve current (Section VI)
γ_k = static capacitance between the cathode surface and the grid, within the space occupied by the electron stream (Section VI)
γ_s = static capacitance between the cathode and grid, outside the electron stream (Section VI)
δ = increment of the quantity following δ
μ = valve amplification factor
$\omega = 2\pi f$.

II. Introduction

In a valve oscillator, the frequency is mainly determined by the frequency sensitive element (crystal, LC circuit). In addition, however, it is affected to some extent by the valve. Changes of the valve parameters which take place during its lifetime, and those brought about by variations of supply voltages, produce in turn frequency changes. These are mainly due to variations of the valve input capacitance and of the amplification factor, and are minimized by a suitable oscillator design. In the simplest LC oscillators, care is taken that the external impedances facing the valve are much smaller than the corresponding valve impedances.[1]

In the study of the sources of frequency instability situated in the valve, one important factor seems to have been neglected; this is the interface layer impedance of oxide coated cathodes. To the knowledge of the

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[1] J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability," Proc. IRE, vol. 36, pp. 356–358; March, 1948.

author, the first note on this subject appeared in Kohn[2] and a detailed study was also published.[3] The interface layer impedance[4] has a reactive component which produces a phase shift of the valve current, and this in turn alters the frequency of oscillation. In valves commonly used in oscillators, the interface layer resistance can easily reach 100 ohms during their serviceable lives, and such a change can cause frequency drifts of some hundreds of parts per million (ppm). This is at least ten times more than is expected from the other sources of instability. To illustrate the magnitude of the effects under consideration, some experimental data on an oscillator operating at 3 mc and designed for good stability in the conventional way are given in Table I.

TABLE I

PERFORMANCE DATA OF AN OSCILLATOR WITH AN INTERFACE LAYER FREE VALVE, AND ONE WITH A PRONOUNCED INTERFACE LAYER

Effects of	Frequency change in ppm in oscillator with new valve	old valve
10 per cent change of anode voltage	+1 to 3	+1 to 3
10 per cent change of filament voltage	−2 to −3	− 20
Change from low to high amplitude operation	−15	+100
Frequency drift in life due to interface layer	—	+300

The most direct way of dealing with this problem is, obviously, by producing valves without interface layers. In fact, valves have been built either without interface layers or at least with low impedance layers,[5,6] and some types of such special-purpose valves, often referred to as computer valves, are now available.

Although valves without any pronounced tendency to developing high resistance interface layers are likely to become commonly used eventually, an analysis of this problem is still warranted, as residual cathode changes will hardly be avoidable. An apparently insignificant change in the interface layer resistance of 1 ohm can produce a frequency change of 6 ppm in an oscillator with a circuit Q of 200 and operating with a mutual conductance of 5ma/v. This is of the same order as frequency changes produced by changes of valve capacitances or of oscillation amplitude.

In this paper, the subject will be treated in a general manner, and not specifically limited to the interface layer. In this way, the results can also be applied to oscillators which use a cathode resistor for negative feedback, for assessing unintentional feedback due to cathode lead inductances, or for determining the effect of insufficiently by-passed cathode resistors. The analysis also takes into account valve capacitances, as they affect the results considerably. Linear operation only is considered.

III. The Oscillator Circuit

In order to make the performance analysis of an oscillator with a cathode impedance, the circuit shown in Fig. 1 and its equivalent circuit diagram in Fig. 2 have been chosen. The oscillator is of the Gouriet-Clapp type,[7] of which several descriptions have been published.[1,8–11] The calculation will be carried out in general terms so that it will also be applicable to the simple Colpitts and Hartley circuit without mutual inductance between the coil sections, and to crystal oscillators.

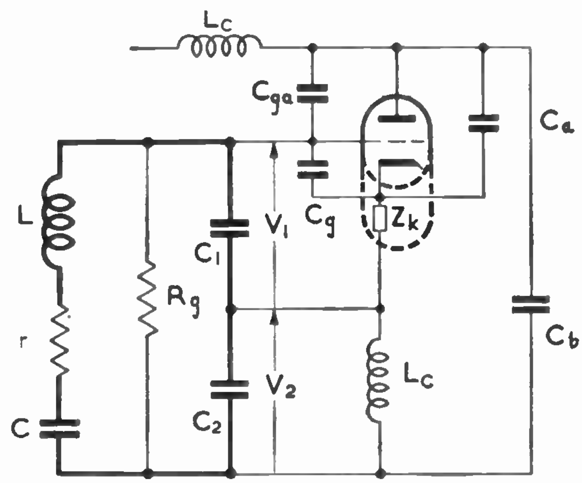

Fig. 1—Circuit diagram of the analyzed oscillator.

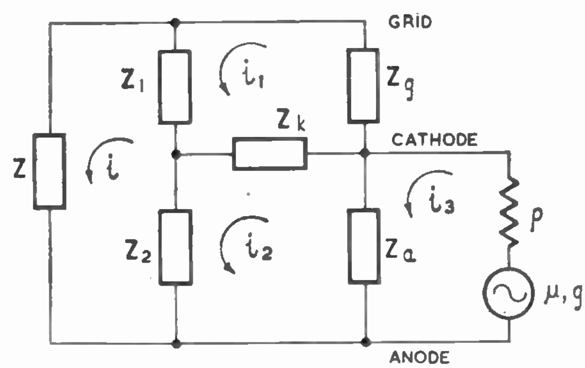

Fig. 2—Equivalent circuit diagram including the anode-cathode valve capacitance.

[2] C. T. Kohn, "The effect of the cathode interface layer on the frequency stability of oscillators," *ATE J.*, vol. 12, p. 106; 1956.
[3] C. T. Kohn, "The Effect of a Cathode Impedance on the Frequency Stability of Oscillators," British Telecommun. Res. Tech. Rept. 164/3/TR; June, 1958.
[4] A. Eisenstein, "The leaky-condenser oxide cathode interface," *J. Appl. Phys.*, vol. 22, pp. 138–148; February, 1951.
[5] G. H. Metson, "The platinum cored oxide cathode repeater valve," *Post Office Elec. Engrs. J.*, vol. 47, pp. 208–211; January, 1955.
[6] A. M. Bounds, T. H. Briggs, and C. D. Richard, "Development of new cathode nickels with improved performance," *Le Vide*, vol. 9, pp. 18–21; May, 1954.

[7] G. G. Gouriet, "High stability oscillator," *Wireless Engr.*, vol. 27, pp. 105–112; April, 1950.
[8] W. A. Edson, "Vacuum Tube Oscillators," John Wiley and Sons, New York, N. Y., pp. 169–172; 1953.
[9] J. K. Clapp, "Frequency stable LC oscillators," Proc. IRE, vol. 42, pp. 1295–1300, August, 1954; vol. 43, pp. 875–876, July, 1955.
[10] J. Vackář, "LC Oscillators and Their Stability," Tesla Tech. Rept. (Czechoslovakia), no. 12, pp. 1–9; December, 1949.
[11] O. Landini, "Oscillatore ad alta stabilità," *Radio Rivista*, pp. 15–17; November, 1948.

The cathode impedance,

$$Z_k = R_k + jX_k, \tag{1}$$

will be represented here by series reactive and resistive components to facilitate analysis, although the interface impedance more closely approaches a parallel RC combination. Physically, the cathode impedance may reside inside the valve (interface layer, lead inductance) or outside (external cathode connections and feedback networks).

The grid-cathode capacitance C_g is here defined as that part of the total grid-cathode capacitance which is in series with the cathode impedance. If the latter is external to the valve, the whole of the conventional grid-cathode capacitance is to be taken as C_g. In the case of an interface impedance, however, C_g is that part of the total grid-cathode capacitance which is confined to the space between the grid and the emitting surface of the cathode. The remaining part of the total grid-cathode capacitance extends to the cathode structure, thus bypassing the interface layer, and is connected directly across C_1; this part will be included in the reactance of C_1.

The grid impedance Z_g is assumed to have no resistive component. This limits the validity of the analysis to cases where transit time effects can be neglected. Also, the dc grid current must be zero; this condition is fulfilled as low amplitude operation only is considered.

With regard to the anode-cathode capacitance C_a, a similar distinction has to be made as with C_g; only that fraction of the total capacitance is to be taken which carries the current passing through the cathode impedance. The remainder falls across C_2 and is to be included in its reactance.

$$Z_1 = jX_1, \qquad Z_2 = jX_2, \tag{2}$$

which may contain part of C_g and C_a, are assumed to be pure reactances, both having the same, though arbitrary, sign. Their actual losses are included in the branch

$$Z = r + jL\omega + 1/jC\omega = r + jX, \tag{3}$$

containing a series connected inductor and capacitor. Any other circuit losses, except R_k, are included in r. The over-all Q factor of the circuit is thus

$$Q = L\omega/r. \tag{4}$$

$$Z_g = jX_g \quad \text{and} \quad Z_a = jX_a \tag{5}$$

are loss free capacitances. L_c and C_b are assumed not to affect the performance.

IV. The General Solution

The network of Fig. 2 can be solved from the following equations:

$$\begin{aligned} i(Z + Z_1 + Z_2) - i_1Z_1 - i_2Z_2 &= 0 \\ i_1(Z_1 + Z_k + Z_g) - iZ_1 - i_2Z_k &= 0 \\ i_2(Z_2 + Z_k + Z_a) - iZ_2 - i_1Z_k - i_3Z_a &= 0 \\ E - i_3\rho - i_3Z_a + i_2Z_a &= 0. \end{aligned} \tag{6}$$

The calculations are straightforward though rather lengthy,[3] and give finally

$$\begin{aligned} 2Q\frac{\Delta f}{f} = &-\frac{\alpha}{N_x} - \frac{1}{A^2 g_0 X_g}\frac{X_k}{X_g}\left(1 + \frac{N_4}{A}\frac{X_k}{X_g}\right)\left[\frac{X_1}{X_2}\left(1 - \frac{X_2}{X_1}\frac{X_g}{X_a}\right)^2 + \left(\frac{X_1 + X_2}{X_a}\right)^2\right] \\ &+ N_R\frac{N_4}{A}\frac{R_k}{X_g}\left[1 - \frac{BN_3}{DN_4} - \frac{B}{D}g_0R_k\right] + \frac{R_k}{AX_g}\left\{\frac{AE}{D\mu}(1 + 2\alpha^2) - \frac{BN_3}{A^2Dg_0X_g}\frac{R_k}{X_g}\frac{X_1}{X_2}\left(1 - \frac{X_2}{X_1}\frac{X_g}{X_a}\right)^2\right. \\ &\left.+ \frac{BE}{D}\frac{X_k}{X_a} - N_4\frac{B}{D}g_0R_k\left[\alpha^2 - 2\beta\left(2 - \frac{B}{D}\frac{N_3}{N_4} - \frac{AEX_g}{N_3N_4X_a} + \frac{AEX_g}{2N_3X_a}\right)\right]\right\} \end{aligned} \tag{7}$$

where

$$\begin{aligned} \frac{g_0}{g} = \frac{D}{AN_1}&\left\{1 - \frac{B}{D}g_0R_k\left[1 - \frac{2N_4}{A}\frac{X_k}{X_g}(1 + \alpha^2) + \beta(1 + 2\alpha^2) - \beta^2 - \frac{E}{D}\frac{X_k}{\mu X_g} - \left(\frac{N_4}{A}\frac{R_k}{X_g}\right)^2\left(1 - \frac{B}{D}\frac{N_3}{N_4}\right)\right.\right. \\ &\left.+ \frac{BN_3N_4}{A^2D}\left(\frac{X_k}{X_g}\right)^2\left(1 - \frac{AEX_g}{N_3N_4X_a}\right) - \frac{X_2}{A\mu X_g}\left(\frac{2N_4}{B} - \frac{N_3}{D}\right)\right] + (\beta - \alpha^2)(1 + \alpha^2) \\ &+ \frac{N_3}{AD}\frac{X_2}{\mu X_g}\left(1 + \frac{X_1}{X_g} + \beta\right) - 2\alpha\frac{g_0X_2}{D\mu}\left(1 + \frac{X_1}{X_g}\right) + \frac{BN_3N_4}{A^2D}\left(\frac{R_k}{X_g}\right)^2 \\ &\left.- \left(\frac{B}{D}g_0R_k\right)^2\frac{2N_4}{A}\left(\frac{X_k}{X_g}(1 - 2\beta + 2\alpha^2) + \frac{X_2}{B\mu X_g}\right)\right\} \end{aligned} \tag{8}$$

$$A = 1 + X_1/X_g + X_2/X_a + X_1X_2/X_aX_g + N_4X_k/X_g \quad (9)$$

$$B = 1 + 1/\mu + (X_1 + X_2)/\mu X_g \quad (10)$$

$$D = 1 - X_2/\mu X_1 - (X_1 + X_2)/\mu X_g - EX_k/\mu X_g \quad (11)$$

$$E = 2 + X_1/X_2 + X_2/X_1 \quad (12)$$

$$N_1 = 1 + \left(\frac{N_4}{A}\frac{R_k}{X_g}\right)^2 + \frac{1}{A^2 g_0 X_g}\frac{R_k}{X_g} \cdot\left[\frac{X_1}{X_2}\left(1 - \frac{X_2}{X_1}\frac{X_g}{X_a}\right)^2 + \left(\frac{X_1 + X_2}{X_a}\right)^2\right] \quad (13)$$

$$N_3 = \frac{X_1}{X_2} + \frac{X_2}{X_1}\frac{X_g}{X_a} + \frac{X_1 + X_2 + EX_k}{X_a} \quad (14)$$

$$N_4 = 1 + \frac{X_g}{X_a} + \frac{X_1 + X_2}{X_a} \quad (15)$$

$$N_R = 1 + 2\alpha^2(1 - 2\beta + 2\alpha^2) - \beta \quad (16)$$

$$N_x = 1 + (\beta - \alpha^2)(1 + \alpha^2) + \frac{2N_3}{AD}\frac{X_2}{\mu X_g} - 3\alpha\frac{g_0X_2}{D\mu}. \quad (17)$$

A, B, D, N_1, N_R and N_x are factors which do not differ much from unity. E has a minimum value of 4 at $X_1/X_2 = 1$ and increases when this ratio deviates from unity. N_3 and N_4 depend mainly on X_1/X_2 and X_g/X_a.

$$g_0 = r/X_1X_2. \quad (18)$$

This is a pure circuit parameter and, as such, does not depend on the valve, although it represents the mutual conductance at which the oscillator would work with a valve having $\mu = \infty$, $C_g = C_a = 0$, and with $Z_k = 0$; g is the mutual conductance at which the valve actually works. The parameters α and β are

$$\alpha = \frac{B}{D} g_0 X_k \quad (19)$$

$$\beta = \frac{N_3}{A}\frac{B}{D}\frac{X_k}{X_g}. \quad (20)$$

Eq. (7) gives the difference Δf between the frequency of the actual oscillator and one in which $Z_k = 0$. Δf is given to 1 ppm if the following conditions are fulfilled:

$$Q > 200;\ \mu > 50;\ X_1/X_2 < 1.5;\ f < 10 \text{ mc};$$
$$X_2X_g/X_1X_a < 2;\ X_1/X_g \text{ and } X_2/X_g < 0.1;\ R_k/X_g \text{ and}$$
$$X_k/X_g < 0.02;\ g_0X_g > 1;\ g_0R_k \text{ and } g_0X_k < 0.25. \quad (21)$$

The ratio X_1/X_2 is limited to less than 1.5 because values higher than this lead to poor frequency stability.[3] The valves employed in oscillators have usually $\mu > 50$ and a g_{max} of about 10 ma/v, and a suitable maximum value of g_0 is 5 ma/v. With R_k and X_k limited to 50 ohms, g_0X_k is about 0.25. It is true that interface layer resistances as high as 1000 ohms have been reported,[12] but such valves should not be used in high stability oscillators in which even 10 ohms may be objectionable. Usually the shunt resistance of the interface layer does not exceed 100 ohms, so that a series component of 50 ohms will cover most cases.

As transit time effects are not considered here, a frequency of 10 mc may be regarded as the upper limit for which the analysis still gives a complete picture of the oscillator performance. With a grid capacitance of 5 pF, X_g is 3000 ohms at 10 mc, and this accounts for the limits given for R_k/X_g and X_k/X_g.

If the conditions in (21) are exceeded, (7) will be less accurate than 1 ppm. For a practical performance assessment, (7) can be simplified, and if bare essentials only are retained it can be written

$$2Q\frac{\Delta f}{f} \simeq -g_0X_k - \frac{1}{g_0X_g}\frac{X_1}{X_2}\frac{X_k}{X_g}\left(1 - \frac{X_2}{X_1}\frac{X_g}{X_a}\right)^2 + \frac{R_x}{X_g}[N_4(1 - g_0R_k) - N_3]. \quad (22)$$

Eq. (8) gives the mutual conductance in operation to an accuracy of 0.1 per cent. With only the most essential terms retained, this can be simplified to

$$g_0/g = 1 - g_0R_k. \quad (23)$$

The increase of g over g_0 is caused by negative feedback on the cathode resistor; the contribution from X_k is only of second order.

V. Case $C_a \ll C_g$

If an interface layer is the source of the cathode impedance, (22) can be written

$$2Q\frac{\Delta f}{f} = -g_0X_k - \frac{1}{g_0X_g}\frac{X_1}{X_2}\frac{X_k}{X_g} + \frac{R_k}{X_g}\left(1 - g_0R_k - \frac{X_1}{X_2}\right). \quad (24)$$

The total frequency change consists of three contributions. The first two terms depend on X_k only, while the third is determined by R_k.

The origin of these terms can be understood by considering the physical reasons for frequency changes in an oscillator to which a cathode impedance has been added. Without Z_k the valve current is approximately V_1g_0. This current flowing through jX_k produces a voltage $jV_1g_0X_k$ in quadrature with the original grid voltage V_1 so that the new anode current is shifted against the original current by an angle $\phi \simeq V_1g_0X_k/V_1 = g_0X_k$. The negative phase shift divided by $2Q$ gives the resulting frequency change[13] $= -(1/2Q)g_0X_k$ which is represented by the first term in (24).

The same valve current, flowing through R_k, produces a voltage $V_1g_0R_k$ which, being in phase with the valve current, does not introduce any direct change of

[12] L. S. Nergaard, "Studies of the oxide cathode," *RCA Rev.*, vol. 13, p. 468; December, 1952.

[13] A. C. Lynch and J. R. Tillman, "The principles and design of valve oscillators," *Electronic Engrg.*, vol. 17, p. 414; March, 1945.

frequency. There is, however, an indirect effect due to the resulting reduction of the (capacitive) grid current V_1/jX_g. The voltage drop $V_1 g_0 R_k$ reduces the voltage across the valve input to $V_1 - V_1 g_0 R_k$ so that the grid current is now $V_1(1-g_0R_k)/jX_g$, which is equivalent to a change of the input impedance from X_g to $X_g' = X_g/(1-g_0R_k)$. Thus, this is the known phenomenon of increasing the input reactance of a valve by means of negative feedback on a cathode resistor. A change of X_g to X_g' in an oscillator produces a frequency change

$$\frac{\Delta f}{f} = -\frac{1}{2Q}\frac{X_1}{X_2}\frac{R_k}{X_g}.$$

This forms part of the third term in (24).

Further frequency changes are caused by the input current $V_1(1-g_0R_k)/jX_g$ flowing through R_k and X_k. The voltage drop across R_k, $V_1R_k(1-g_0R_k)/jX_g$, is in quadrature with the initial grid voltage V_1, thus giving a phase change of $\phi = -(R_k/X_g)(1-g_0R_k)$ and a frequency change of

$$\frac{\Delta f}{f} = \frac{1}{2Q}\frac{R_k}{X_g}(1 - g_0R_k).$$

This expression can be found in the third term of (24).

The voltage drop across jX_k, V_1X_k/X_g is in phase with V_1; hence, there is no change of phase. However, a frequency change is produced, as now the input reactance jX_g connected across jX_1 is changed to $jX_g + jX_k$. Such a change gives

$$\frac{\Delta f}{f} \simeq -\frac{X_1}{2L\omega}\frac{X_1}{X_g}\frac{X_k}{X_g}\frac{1}{A\left(1+\frac{X_1}{X_g}\right)}.$$

which is the second term in (24).

Hence, there are four major ways in which the frequency of an oscillator with cathode impedance is affected. Two act by injecting into the grid circuit quadrature voltages giving rise to phase changes, the other two by directly modifying the input capacitance of the valve. The $(1/2Q)(\alpha/N_x)$ term is practically independent of the input capacitance and remains important over the entire frequency band. The other three are the result of the valve input reactance not being infinite, and, therefore, they disappear at low frequencies. In general, however, their contributions can be of the same order as that of the first term.

It is now obvious that a simple consideration of the cathode impedance alone, without the grid capacitance, is insufficient for a full appraisal of the effect on the frequency of oscillation. Large contributions come from the interaction with the valve input capacitance which thus becomes such an important parameter in the design of an oscillator.

The optimum performance conditions are determined separately for the X_k and R_k terms. Differentiation of the first two terms in (24) against g_0 and equation to zero gives

$$g_0 |X_g|_{\text{opt}} = \sqrt{X_1/X_2}. \tag{25}$$

The R_k/X_g term becomes zero if

$$X_1/X_2 = 1 - g_0R_k. \tag{26}$$

Both equations combined lead to the following optimum performance conditions:

$$X_1/X_2 \simeq 1 \quad \text{and} \quad g_0|X_g| \simeq 1. \tag{27}$$

A comparison of these conditions with recommendations given in the literature on oscillator stability[7,8] with the data provided by (27), shows that they are at variance. The familiar design methods aim at alleviating the harmful effects of valve capacitance variations, swamping them by means of low impedances X_1 and X_2 connected across the valve. Low values of X_1 and X_2 lead automatically to high values of g_0 and these create very unfavorable conditions if Z_k is not constant. In practice, values of a few ma/v are customary. This is in complete contrast with (27) which calls for values one order lower, or less. The X_1/X_2 ratio is sometimes chosen on the basis that the total frequency change caused by a simultaneous change of C_g and C_a by ΔC_g and ΔC_a, respectively, should be a minimum. This leads to $X_1/X_2 = \sqrt{\Delta C_a/\Delta C_g}$ which is usually less than one.[8] In contrast with this, (27) calls for $X_1/X_2 = 1$. The latter figure agrees with recommendations given in Clapp and Gouriet,[1,7] though for other reasons.

The present analysis is based on the recognition that the most changeable electrode in a valve is its cathode, and that changes in valve characteristics and parameters are to a great extent the result of feedback voltages across cathode impedances. The design aims, therefore, at establishing conditions in which cathode impedances have little or no effect on the frequency. These are given by (27).

VI. Variable Input Capacitance

So far, the input capacitance was assumed to be constant. It now becomes necessary to examine to what extent the present results will be modified when grid capacitance changes, which are the basis of standard design methods, are taken into account.

For the purpose of this section, the input capacitance can be divided into two types. The first one consists of the structural capacitances γ_s outside the electron stream (see Fig. 3) and of the static capacitance γ_k between the grid and the emissive cathode surface. If a valve of good mechanical design, for example a ruggedized version, is used, no significant variations of these capacitances should occur, if carefully handled, during its lifetime. The static valve capacitances can also be altered by barium, evaporated from the cathode and deposited on insulating parts close to the valve electrodes. If the grid capacitance C_g is changed in this way by

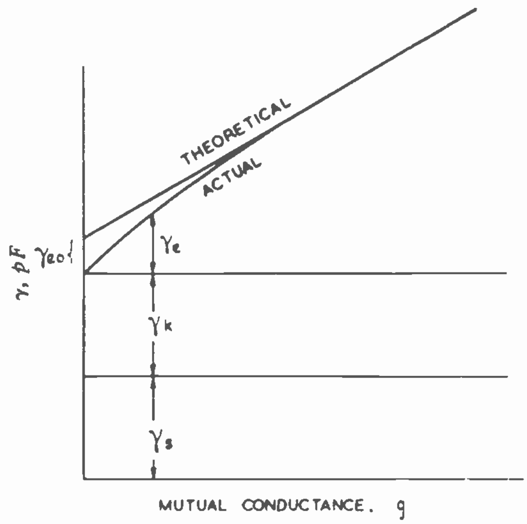

Fig. 3—Dependence of the grid-cathode capacitance γ on the mutual conductance g.

ΔC_g, the resulting frequency change is

$$\frac{\Delta f}{f} = \frac{1}{2Q} \frac{X_1}{X_2} \frac{1}{g_0 X_g} \frac{\Delta C_g}{C_g}, \tag{28}$$

which is small if X_1/X_2 is small and g_0 high. If, quite arbitrarily, it is assumed that this instability, in optimum working conditions, is not to exceed $\frac{1}{10}$ of the other major terms (24), the permissible random grid capacitance changes must not be greater than 0.01 pF. With mechanically stable valves and cathodes operating at reduced temperatures where barium evaporation is negligible, this may just be achievable.

The second source of grid capacitance changes is found in the variations of the capacitance γ_e which is the result of the space charge and the current in the valve.[14,15] Such a capacitance change may be assumed, within narrow limits, proportional to the mutual conductance

$$\gamma_e = Kg. \tag{29}$$

In the oscillator, g changes are introduced by the cathode resistance R_k (23) and produce a frequency change of[3]

$$\frac{\Delta f}{f} \simeq -\frac{1}{2Q} \frac{BX_1}{DX_2} K\omega g_0 R_k = +\frac{1}{2Q} \; BX_1 \; R_k \cdot \tag{30}$$

where $X_{sc} = -1/K\omega g_0$.

This shows clearly that frequency changes caused by the space charge capacitance are lowest when g_0 is small. If in an oscillator $Q=200$, $X_1/X_2=1$, $f=10$ mc, $R_k=50$ ohms and the space charge capacitance characteristic has a slope K of 0.5 pF per 1 ma/v ($0.5\cdot 10^{-9}$ ohm farads), the frequency change $\Delta f/f$ is 20 ppm at 5 ma/v, and 0.4 ppm at 0.1 ma/v. In view of the smallness of this effect, the frequency change can be introduced in the frequency equation without considering second-order effects in other terms.

It is interesting to note that if the oscillator is operated in optimum conditions (27), the contributions to the total frequency change are $(1/2Q)\ (X_k/X_g)$ from each of the two first terms; $(1/2Q)\ (R_k/X_g)$, cancelling each other, from the two terms combined in term 3; and $(K\omega/2Q)\cdot(R_k/X_g)$ from space charge effects. If R_k and X_k are equal, all the four main terms are equal, indicating a well balanced design. The space charge term is at least one order smaller and therefore does not affect noticeably the frequency changes caused by cathode impedances. If the oscillator is run at higher values of g_0, this term increases.

The general conclusion is that in an oscillator, frequency changes caused by input capacitance variations due to changes of g are negligible, if the oscillator is operated near the optimum working conditions as given by (27). Frequency changes caused by random variations of the grid capacitance will have to be considered sometimes, but in general their effect on the frequency is much lower than the effect of the cathode interface impedance, if the latter is fully effective.

VII. The Effect of the Anode Capacitance

If the cathode impedance is outside the valve, the total capacitance from anode to cathode is of the same order as the grid capacitance and must be considered. The performance is determined by (22).

A comparison of (22) and (24) shows that the major difference between the expressions is in the second term, which is the result of the interposition of the cathode reactance between X_1 and X_g. Here the term $(X_2/X_1)\ (X_g/X_a)$ represents the effect due to the anode reactance X_a. If this term is made equal to 1, X_a cancels the effect of X_g and the frequency change is reduced to zero. The physical reason for the cancellation can be seen from Fig. 2. The impedances Z_1, Z_2, Z_g, and Z_a form a bridge network with Z_k as its diagonal. If $Z_1/Z_2 = Z_g/Z_a$ the bridge is balanced and there is no voltage across Z_k. Z_k can now have any value without affecting the currents flowing in the four branches and in the second diagonal containing Z. In this case, not only the third term of (22) can be reduced to zero but the second also. The remaining frequency change is due to the $g_0 X_k$ term.

Formally, the optimum working conditions can be written

$$g_0 X_k \simeq 0$$

$$\frac{X_1}{X_2}\left(1 - \frac{X_2}{X_1}\frac{X_g}{X_a}\right)^2 + \left(\frac{X_1 + X_2}{X_a}\right)^2 = 0$$

$$1 - \frac{B}{D}\frac{N_3}{N_4} - \frac{B}{D} g_0 R_k = 0 \tag{31}$$

[14] D. O. North, "Analysis of the effects of space charge on grid impedance," Proc. IRE, vol. 24, pp. 108–136; February, 1936.

[15] R. H. Booth, "Triode interelectrode capacitances," *Wireless Engr.*, vol. 26, p. 211; June, 1949.

from which

$$X_1/X_2 = X_g/X_a \quad \text{and} \quad g_0 \to 0 \text{ (mode 1).} \tag{32}$$

This is the method of operation which gives the smallest frequency change as far as the effect of a cathode impedance is concerned. All the major terms are minimized so that the frequency becomes very little dependent on the cathode impedance. However, the ratio X_1/X_2 must be adjusted to balance with the valve capacitances.

It may not always be possible to balance X_1, X_2, X_a, and X_g. In this case, the optimum working conditions are obtained by minimizing the sum of the two first terms of (22) as was done in the case $C_a \ll C_g$, instead of minimizing each separately. This gives

$$\frac{X_1}{X_2} = 1, \qquad g_0 = \left| \frac{1}{X_g} - \frac{1}{X_a} \right| \text{ (mode 2).} \tag{33}$$

The performance here is not quite as good as in mode 1, but the value of X_1/X_2 is independent of the valve. Now the optimum value of g_0 depends on X_g and X_a, but quite large deviations of g_0 from the optimum are permissible.

If the two modes of operation are compared, it will be seen that although mode 1 can give better performance, mode 2 will find a more general application in practice. It stabilizes the oscillator simultaneously for an interface impedance and for an external cathode impedance. The X_1/X_2 ratio is independent of the valve and has the convenient value of 1, generally favored in the design of high stability oscillators. The g_0 value is admittedly higher, but it may prove necessary to increase g_0 above the minimum in any case, in order to improve the short-term stability (Section VIII). If a good over-all performance is aimed at, mode 2 will be given preference.

Summarizing the results of this section, it can be stated that a finite anode capacitance usually improves the performance of the oscillator. If the elements of the tuned circuit form a balanced reactance bridge with the valve capacitances ($X_1/X_2 = X_g/X_a$), the influence of a cathode impedance on the frequency is smallest when $g_0 \to 0$. If it is not practicable to balance the capacitances, the X_1/X_2 ratio should be made equal to 1, and the value of g_0 chosen from (33). For any particular valve, the higher the frequency, the higher is the optimum value of g_0.

VIII. The Effect of Supply Voltages

In a linear oscillator, supply voltage changes produce frequency changes mainly by affecting the grid capacitance and the amplification factor of the valve.[8] The relative frequency change $\delta f/f$, due to a relative supply voltage change of $\delta V_s/V_s$, amounts to

$$\frac{\delta f}{f} \Big/ \frac{\delta V_s}{V_s} = \frac{1}{2Q}\left[\left(\frac{X_1}{X_2} + \frac{X_1 + X_2}{X_a}\right) \cdot \frac{1}{A g_0 X_g} \frac{\delta C_g}{C_g} + \frac{A}{D} \frac{g_0 X_2}{\mu} \frac{\delta \mu}{\mu}\right] \tag{34}$$

where δC_g and $\delta\mu$ are the changes of C_g and μ, respectively, caused by δV_s.

If a constant impedance is inserted in the cathode of the oscillator, the sensitivity to supply voltage changes is not affected significantly.[3] Therefore, all rules applying to oscillators in general will hold good in the presence of a cathode impedance also. From (34) it can be seen that, apart from choosing valves and operating points with low $\delta C_g/C_g$ and $\delta\mu/\mu$, it is advantageous to use high values of g and X_2/X_1 if the $\delta C_g/C_g$ term is prevalent (high frequencies), and high μ values operating at low g and low X_2/X_1 ratios when the $\delta\mu/\mu$ term prevails.

These conclusions do not apply fully to valves with an interface layer. The interface layer has the properties of a semiconductor: its impedance thus depends strongly on temperature. If the heater voltage changes, Z_k changes as well, and to the frequency coefficient given by (34), an additional term caused by ΔR_k and ΔX_k must be added; this additional frequency change can be obtained from (22) or (24) in which ΔR_k and ΔX_k is substituted for R_k and X_k, respectively. As can be seen from Table I, this contribution can be many times greater than the effect of C_g and μ changes. Means for reducing this effect are operation at low values of g_0 and at high cathode temperatures; *i.e.*, at the maximum permissible heater voltage. The second measure will, of course, speed up the growth of the interface layer.

Changes of the HT voltage do not affect the interface layer impedance; hence, the effect of anode voltage changes does not increase with the formation of the interface layer.

The optimum operating conditions, derived with the view to obtaining a small dependence on supply voltages, clash to a great extent with those giving small frequency changes with a change of the cathode impedance, and a compromise between conflicting factors must be made. In particular, good independence of Z_k calls for low g_0, and good short-term stability calls for a large g_0 if Z_k is constant, but for a rather low g_0 if Z_k is an interface impedance. The requirement for good short-term stability is, apart from random variations of C_g, the only factor which will induce the designer to adopt a higher g_0 value than given by (27), (32), or (33).

IX. Experimental Verification

A large number of tests was made in order to check the analysis. It was found that if the cathode impedance is considered in connection with the valve capacitance C_a and C_g, a satisfactory explanation of the behavior of an oscillator with cathode impedance can be obtained.[3]

Here, only the more important results will be given.

The tests were made by inserting a capacitor or a resistor into the cathode lead of a 3-mc oscillator and by measuring the resulting frequency change. In Fig. 4, the conditions were chosen so that the g_0X_k term was prevalent. This results in a linear relationship between Δf and C_1C_2. The calculated Δf line agrees with measured points very well, except for the smallest frequency changes where the effect of the valve capacitances becomes noticeable.

Fig. 5 shows the effect of valve capacitances which were increased intentionally by the addition of external capacitances in order to make their effect more pronounced. The measured curves (m) confirm the calculated characteristics (c) fairly well, especially as far as the positions of the Δf minima are concerned. The existing differences of the values of the minima are explained by difficulties in accurately measuring small capacitances, although the existence of factors other than those covered by the theory is not excluded.

Fig. 6 shows the performance in the presence of a cathode resistor. Frequency changes close to zero appear at $X_1/X_2=1$ and where $X_1/X_2=X_g/X_a$, as predicted by the analysis. Of particular interest is curve 1 which shows the advantage of using X_g/X_a close to 1; the X_1/X_2 ratio then becomes noncritical.

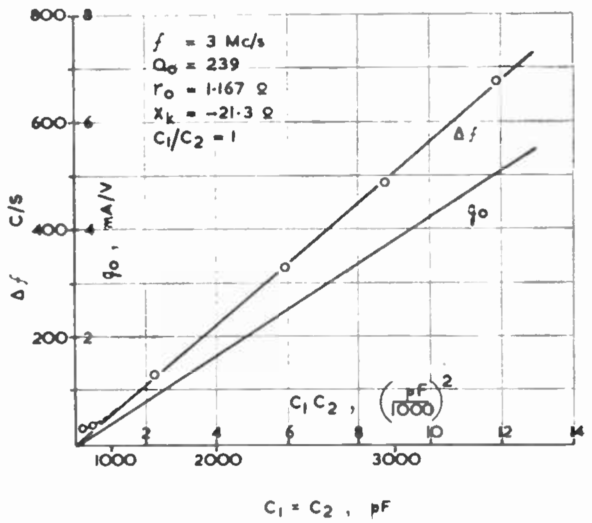

Fig. 4—The effect of the g_0X_k term on the frequency of oscillation as a function of $C_1C_2\sim g_0$ for $X_1/X_2=1$.

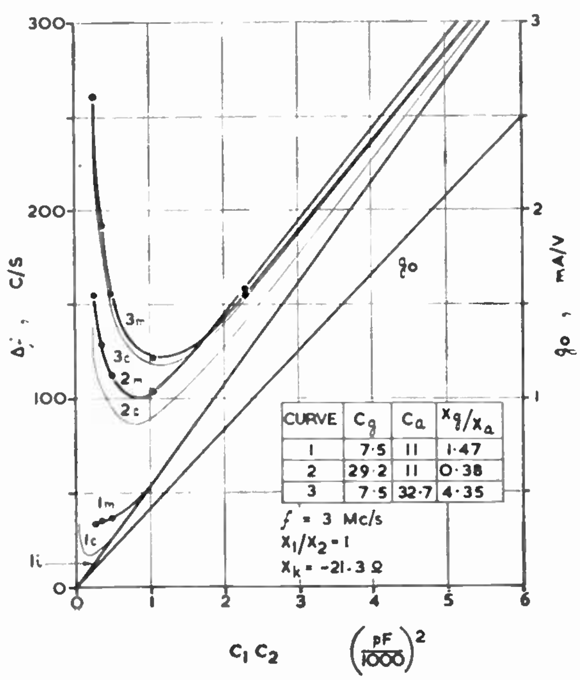

Fig. 5—The combined effect of the g_0X_k and the $X_k/g_0X_g^2$ terms on the frequency of oscillation for various ratios of X_gX_a at $X_1/X_2=1$, as a function of $C_1C_2\sim g_0$.

X. Conclusion

It has been shown that a cathode impedance considerably affects the frequency of an oscillator. If the cathode impedance is due to a growing interface layer, the resulting frequency drift can reach a few hundred ppm in 1000 hours. The growing interface layer is certainly the most important single valve factor giving rise to poor long-term and short-term stability; the latter is caused by the extreme sensitivity of the semiconductor material of the interface layer to temperature (*i.e.*, heater voltage) changes.

It is necessary, therefore, to revise the design of oscillators, which in the past has been based on the assumption that the frequency instability arising from the valve was mainly due to the valve capacitances. That this is not necessarily so can be seen from Table I, where the measured sensitivity to heater voltage changes is up to 10 times greater in a valve with a con-

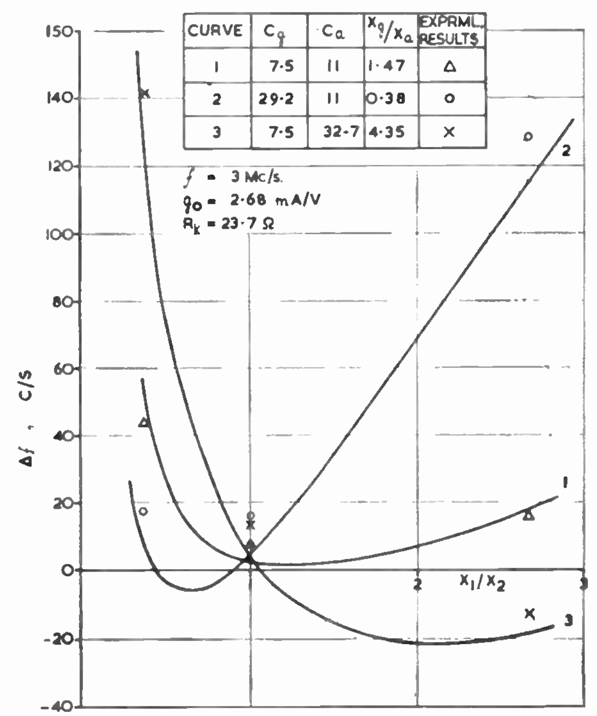

Fig. 6—The frequency change due to a cathode resistance as a function of X_1/X_2, for various values of X_g/X_a.

siderable interface layer impedance than in one which is free of it. The cathode impedance must be considered in the design with as much care as the valve capacitance.

In oscillators, the latter is shown not to be as harmful as generally feared. When ruggedized valves are used at reduced heater voltage (*e.g.*, 5 v instead of 6.3 v) and with low dissipation, the frequency changes due to this cause are, in oscillators of customary design, 2 to 3 orders lower than the possible frequency instability due to the interface layer, which thus becomes the principal source of instability.

Broadly speaking, two ways of operation can be recommended. The first consists in eliminating the interaction between the cathode impedance and the valve capacitances by making X_1/X_2 equal to X_g/X_a, and in reducing the effect of the cathode reactance by reducing the mutual conductance as far as possible. This method gives the smallest frequency instability for any specified change of the cathode impedance. The optimum working conditions are independent of the value of the cathode impedance, but depend on the valve capacitances which are different for an external cathode impedance and an interface layer impedance, and which may lead to an inconvenient X_1/X_2 ratio.

The second method consists in eliminating the interaction between the cathode resistance and the valve capacitances by making $X_1/X_2=1$, and in minimizing the effect of the cathode reactance by using a mutual conductance in accordance with (33). Here the frequency instability is larger, but X_1/X_2 is independent of the valve. The value of g_0 is higher here than in the first method, but it is still much lower than is customary in designs aiming at a reduction of the effects of grid capacitance changes alone. As this method works equally well with internal and external cathode impedances, and uses the convenient X_1/X_2 ratio of 1, it should prove the most successful general design.

The conditions for obtaining good frequency stability with supply voltage changes differ to a great extent from those just recommended. A compromise is necessary here.

The final procedure in the design of a linear oscillator will consist in determining three effects: those of the cathode impedance, of random grid capacitance variations, and of the effect of supply voltage variations, for various X_1/X_2 and g_0. As far as X_1/X_2 is concerned, the choice is between 1 and X_g/X_a; g_0 may take a wide range of values depending on the magnitude and importance of the variable parameters Z_k, X_g, and V_s. An example of a performance calculation is given in Table II which applies to an oscillator with $Q=200$, $L\omega=300$ ohms, $f=3$ mc, $\mu=65$, $C_g=6$ pF, $C_a=9$ pF. It is assumed that the valve can develop, in its lifetime, a cathode impedance of 20–j20 ohms, that the grid capacitance changes can amount to 0.01 pF, and that a change of the supply voltage by 10 per cent gives rise to $\delta C_g/C_g=1$ per cent and $\delta\mu/\mu=-10$ per cent.

TABLE II

CALCULATED OSCILLATOR PERFORMANCE FOR VARIOUS VALUES OF X_1/X_2 AND g_0

X_1/X_2	g_0 mA/V	$\Delta f/f$ in ppm caused by $Z_k=20-j20\Omega$	$\Delta C_g=0.01$pF	$\delta V_s/V_s=0.1$
1	5	250	0.1	0.6
	1	50	0.5	1.6
	0.2	10	2.2	7.1
	0.05	3.1	8.9	28
	0.028	2.6	16	50
1.5	0.05	2.4	13.3	42
	0.018	1.2	24	75
	0.01	0.5	66	210

If $X_1/X_2=1$ and g_0 is 5 ma/v, the long-term stability due to the growth of Z_k is very poor, while the short-term stability is excellent. If the g_0 value is chosen in accordance with (33) ($g_0=0.028$ ma/v, the long-term drift is negligible but C_g changes of 0.01 pF may cause frequency changes of 16 ppm. A fair compromise is obtained with $g_0=0.2$ ma/v; the long-term stability is 10 ppm, capacitance variations produce 2.2 ppm, and frequency changes due to supply voltage variation can be reduced to 0.7 ppm, if a voltage stabilizer capable of maintaining the supply voltage to an accuracy of 1 per cent is used.

If the same oscillator is operated at a lower frequency, for instance 100 kc, the ΔC_g and δC_g instabilities are reduced 30 times. X_1/X_2 can now be made 1.5 with $g_0=0.028$ ma/v which gives stabilities of 1.2, 0.8, and 0.25 ppm, respectively; no measurements, however, have been made to confirm this result.

In crystal oscillators, because of the high Q factor of the crystal, the effect of a cathode impedance is less pronounced. If for a 3-mc crystal, Q is 45,000 with $L=0.13$ H and a total shunt capacitance of 20 pF, and $C_1=C_2=40$ pF, the value of g_0 is 0.21 ma/v. In the same conditions as assumed in Table II the frequency changes become 0.12, 0.026, and 0.17 ppm, respectively. For technical applications this is excellent, but in standard frequency oscillators the drift caused by a growing interface layer would be excessive and the supply voltages would have to be stabilized. It is interesting to note that the drift caused by the growing interface layer goes in the same direction as frequency changes due to normal crystal aging, so that one can be mistaken for the other.

The peculiarities of interface layer impedances could not be dealt with in detail, although some of their properties have been pointed out. It was also impossible to give here an analysis of nonlinear operation in the presence of a cathode impedance, but a preliminary note on this subject can be found in Kohn.[2]

XI. ACKNOWLEDGMENT

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Multiple Diversity with Nonindependent Fading*

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Summary—**Previous analyses of diversity techniques are extended to include the performance of an optimum (maximal-ratio) combiner in the case of nonindependent signal fading fluctuations, for an arbitrary number of diversity branches. The analysis includes the general possibility of correlations among the quadrature components of the various signals. Some computational simplifications for certain cases of physical interest are given, as well as a specific application to two problems in digital communications.**

Introduction

DIVERSITY reception methods are in extensive use on fading radio circuits, especially as higher and higher reliabilities are demanded. Space diversity (separated antennas) is probably the most common form, but in addition, techniques of frequency, time, and angle-of-arrival diversity are either in use or being contemplated for future use.

In order to gain a significant advantage from the use of diversity there must be a sufficient degree of statistical independence in the fading of the several received signals. There are many situations where, as a practical matter, the correlation between the several signals is small enough so that it can be ignored. On the other hand, as the use of diversity techniques expands, there arise many cases where physical limitations prevent the correlation from being small. In these cases the diversity gain may be grossly overestimated if independent fading is assumed. Such situations include multiple space diversity where available space limits antenna separation (*e.g.*, wing-tip mounted antennas on an aircraft); frequency diversity where spectrum allocations limit frequency separation; time diversity for binary transmissions on slowly fading circuits where storage capabilities limit maximum time separation; and angle-of-arrival diversity where antenna design characteristics place an inherent limit on independence of signals on different beams from the same antenna structure.

In the past, statistics of the diversity-combined signal have been largely derived under the assumption of independent fading. Certain recent, more general analyses were reported in the special case of dual diversity. In this paper, we derive formulas for the probability density of the signal power after optimum (maximal-ratio) combining, for an arbitrary number of signals, with arbitrary correlations between all diversity branches, and arbitrary average powers in each. This result should provide a powerful analytical tool for realistic estimation of system performance. It is expected that it can lead to considerable savings in system design efforts, particularly in evaluating proposals for new or extended diversity methods.

Several forms of signal combining have been suggested in the past: pre- and post-detection, with or without weighting factors, "switch" diversity, and, for digital communications, various modifications of the others. Of these, Brennan[1] and Kahn[2] have shown that a form of weighted predetection combining yields the largest SNR possible with a linear combining network. It is this maximal-ratio combining which we have treated in this report. Maximal-ratio combining yields an output SNR equal to the sum of the SNR's in each of the several diversity branches.

In general, in analyzing diversity systems we can safely ignore the signal modulation and consider the signals as CW waveforms bearing the fading and corrupted by noise. Furthermore it is assumed that the effect of the process underlying the fading, in addition to producing fluctuating signal levels, is to completely randomize the RF phase, the fluctuations in the latter occurring at a rate similar to those in the signal level. Thus each signal may be regarded as composed of fading quadrature components, and experimental evidence tends to support the assumption (which also arises from physical reasoning) that the fluctuations in each of these quadrature components constitute a joint normal process. This implies, in fact, that the fading CW signals have the statistical attributes of narrow-band Gaussian noise. Thus, the resulting envelope distributions on scatter circuits and other long-distance links tend to follow the Rayleigh distribution over short measuring periods. In the absence of contrary evidence, this phenomenological model will be assumed to be valid.

Using the assumption of a joint normal process, Altman and Sichak[3] have derived the cumulative distribution of combined power when the several signals are independent. The case of dual diversity with nonindependent fading has also been treated previously for this model, by two different methods. The first[4] of these in-

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[1] D. G. Brennan, "On the maximum signal-to-noise ratio realizable from several noisy signals," PROC. IRE, vol. 43, p. 1530; October, 1955. Also, "Linear diversity combining techniques," PROC. IRE, vol. 47, pp. 1075–1102; June, 1959.

[2] L. Kahn, "Ratio squarer," PROC. IRE, vol. 42, p. 1704; November, 1954.

[3] F. J. Altman and W. Sichak, "Simplified diversity communication system for beyond-the-horizon links," *Elec. Commun.*, vol. 33, pp. 151–160; June, 1956 (reprinted with revisions from IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-4, pp. 50–55; March, 1956). The result also appears in H. Staras, "The statistics of combiner diversity," PROC. IRE, vol. 44, pp. 1057–1058; August, 1956.

[4] G. B. Parrent and M. J. Beran, "Space diversification for receiving antennas in scatter communication systems," presented at Third Aero. Comm. Symp., Utica, N. Y.; November, 1957.

volves Rice's[5] result for the joint density function of two envelopes and cannot be extended since higher order joint envelope distributions are not known except in the case of the Markov process; the second method, due to Packard,[6] involves resolution of the signals into sums of independent Gaussian components—a formulation which could presumably be extended but which would become extremely unwieldy.

The principal result of this paper assumes a complete knowledge of the variances and covariances among all possible pairs of signal components from all the diversity branches. Although a mathematically simple concept, it is a difficult engineering problem to obtain these correlations by direct measurement and, in fact, we do not know of any such extensive experimental data. For this reason, we discuss the use of data on envelope correlations, as are extensively measured. Use of these envelope correlations will usually be only an approximation to the true performance, since essentially half of the pertinent information has been discarded. Nevertheless, performance estimation from envelope data should be valuable in that it provides a considerably more accurate picture than any similar estimate based on the assumption of independent signals.

From a point of view of computational convenience, it is desirable to simplify or eliminate the required matrix calculations wherever possible. A section of the paper is devoted to this topic, and shows the simplifications possible in certain special cases of physical interest. The previously known results for independent fading and for dual diversity are also reviewed as special cases of the more general form.

The final section of the paper applies the general result to performance analysis of two typical digital systems, one with linear combination and detection, the other with envelope detection and post-detection combining.

The Moment Matrix

We will refer our analysis to a space diversity system for simplicity in the ensuing discussion. We will assume, in line with other investigators, that the fading carrier has the properties of a narrowband Gaussian noise.

Suppose that there are N antennas, not necessarily identical, in the diversity array. The signal from the nth antenna can be written as

$$e_n(t) = I_{cn}(t) \cos \omega t + I_{sn}(t) \sin \omega t, \tag{1}$$

where $\cos \omega t$ is a reference cosinusoid at the nominal carrier frequency. I_{cn} and I_{sn} are the slowly varying in-phase and quadrature, or cosine and sine, components of the signal, and are normally distributed with zero mean.

We will define the power level at the nth antenna as

$$S_n = \tfrac{1}{2}(I_{cn}^2 + I_{sn}^2). \tag{2}$$

This is the average of instantaneous power over a few cycles at the carrier frequency; it is also sometimes spoken of as the *envelope* power since it is proportional to the square of the envelope of the fading signal. Furthermore, with little loss in generality we take the average noise power in each receiver to be the same[7] so that the diversity combination results in an addition of signal powers. Denoting the sum of the received powers by S, we have

$$S = \sum_{n=1}^{N} S_n, \tag{3}$$

and in the following sections, we derive the density and distribution functions of S.

We order the $2N$ components from the N antennas in a row matrix,

$$X = [I_{c1}, I_{s1}, I_{c2}, I_{s2}, \cdots, I_{cN}, I_{sN}]. \tag{4}$$

The N antenna currents constitute a joint normal process (by assumption) which is completely specified by a $(2N \times 2N)$ *moment* matrix or *covariance* matrix:

$$M = \begin{bmatrix} \overline{I_{c1}^2} & \overline{I_{c1}I_{s1}} & \cdots & \overline{I_{c1}I_{sN}} \\ \overline{I_{s1}I_{c1}} & \overline{I_{s1}^2} & \cdots & \overline{I_{s1}I_{sN}} \\ \overline{I_{c2}I_{c1}} & \overline{I_{c2}I_{s1}} & \cdots & \overline{I_{c2}I_{sN}} \\ \cdots & \cdots & \cdots & \cdot \\ & & \cdots & \\ \overline{I_{sN}I_{c1}} & \overline{I_{sN}I_{s1}} & \cdots & \overline{I_{sN}^2} \end{bmatrix}. \tag{5}$$

We note that M is symmetric and furthermore that it must have a non-negative definite quadratic form.[8]

Consider now the expansion of e_n with respect to the same reference frequency but with a new reference phase θ:

$$e_n(t) = I_{cn}'(t) \cos (\omega t + \theta) + I_{sn}'(t) \sin (\omega t + \theta). \tag{6a}$$

The primed components are related to the original ones by

$$\begin{aligned} I_{cn}' &= I_{cn} \cos \theta - I_{sn} \sin \theta, \\ I_{sn}' &= I_{cn} \sin \theta + I_{sn} \cos \theta. \end{aligned} \tag{6b}$$

Looking at just one of the four possible moments between the components of two signals, e_n and e_m, we find that

[5] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. J.*, vol. 24, pp. 77–78; January, 1945.

[6] K. S. Packard, "Effect of correlation on combiner diversity," Proc. IRE, vol. 46, pp. 362–363; January, 1958.

[7] For unequal noise powers, we can imagine the gain of each receiver to be altered so as to make all noise powers equal. The signal voltages and the correlation coefficients used in the calculations below must then be altered by corresponding factors. It is emphasized that this is a matter of mathematical convenience only since the maximal-ratio combining rule holds without any restrictions on relative receiver noise levels.

[8] For example see H. Cramer, "Mathematical Methods of Statistics," Princeton University Press, Princeton, N. J., p. 295, sec. 22.3; 1946.

$$\overline{I_{cn}'I_{cm}'} = \overline{(I_{cn}\cos\theta - I_{sn}\sin\theta)(I_{cm}\cos\theta - I_{sm}\sin\theta)}$$
$$= \overline{I_{cn}I_{cm}}\cos^2\theta + \overline{I_{sn}I_{sm}}\sin^2\theta$$
$$- (\overline{I_{cm}I_{sn}} + \overline{I_{cn}I_{sm}})\sin\theta\cos\theta.$$

Since the moments should be invariant with respect to the reference phase, we must have

$$\overline{I_{cn}'I_{cm}'} = \overline{I_{cn}I_{cm}} \text{ for all } \theta. \tag{7}$$

Setting $d(\overline{I_{cn}'I_{cm}'})/d\theta = 0$, we obtain

$$\overline{I_{cn}I_{cm}} = \overline{I_{sn}I_{sm}},$$
$$\overline{I_{cn}I_{sm}} = -\overline{I_{sn}I_{cm}}, \tag{8a}$$

and in particular,

$$\overline{I_{cn}^2} = \overline{I_{sn}^2}, \qquad \overline{I_{cn}I_{sn}} = 0. \tag{8b}$$

Therefore the moment matrix can be written in the partitioned form

$$M = \begin{bmatrix} A_1 & B_{12} & B_{13} & \cdots & B_{1N} \\ B_{12}^t & A_2 & B_{23} & \cdots & B_{2N} \\ B_{13}^t & B_{23}^t & A_3 & \cdots & B_{3N} \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ B_{1N}^t & B_{2N}^t & B_{3N}^t & \cdots & A_N \end{bmatrix} \tag{9a}$$

where the superscript t denotes the transpose of a matrix, and A_n and B_{nm} are 2×2 matrices of the form

$$A_n = \begin{bmatrix} a_n & 0 \\ 0 & a_n \end{bmatrix}, \qquad B_{nm} = \begin{bmatrix} b_{nm} & \beta_{nm} \\ -\beta_{nm} & b_{nm} \end{bmatrix}, \tag{9b}$$

with the constants being given by (5) and (8) as

$$a_n = \overline{I_{cn}^2}, \qquad b_{nm} = \overline{I_{cn}I_{cm}}, \qquad \beta_{nm} = \overline{I_{cn}I_{sm}}. \tag{9c}$$

Assuming M is nonsingular, the joint normal density function is then[9]

$$p(X) = (2\pi)^{-N} | M |^{-1/2} \exp(-\tfrac{1}{2}XM^{-1}X^t), \tag{10}$$

where M^{-1} is the inverse of M, and $|M|$ is the determinant.

Direct Evaluation of Moments

It should be noted that in terms of the signals themselves,

$$\overline{e_n(t)e_m(t)} = \overline{I_{cn}I_{cm}}\cos^2\omega t + \overline{I_{sn}I_{sm}}\sin^2\omega t$$
$$+ (\overline{I_{cn}I_{sm}} + \overline{I_{sn}I_{cm}})\sin\omega t\cos\omega t$$
$$= \overline{I_{cn}I_{cm}},$$

or

$$b_{nm} = \overline{e_ne_m} \tag{11a}$$

and

$$a_n = \overline{e_n^2}.$$

[9] *Ibid.*, p. 311, sec. 24.2.

Similarly, if we subject $e_m(t)$ to a 90° phase shift, the result would be

$$e_m'(t) = I_{sm}(t)\cos\omega t - I_{cm}(t)\sin\omega t,$$

so that

$$\overline{e_n(t)e_m'(t)} = \overline{I_{cn}I_{sm}}\cos^2\omega t - \overline{I_{sn}I_{cm}}\sin^2\omega t$$
$$+ (\overline{I_{sn}I_{sm}} - \overline{I_{cn}I_{cm}})\cos\omega t\sin\omega t$$
$$= \overline{I_{cn}I_{sm}},$$

or

$$\beta_{nm} = \overline{e_ne_m'}. \tag{11b}$$

Thus the coefficients of the moment matrix may be found directly from the signals, with and without phase shifting, by multiplication and averaging without any necessity for resolving the several signals into their sine and cosine components.

It is well to mention here that the averages in (11a) and (11b) are to be construed as finite time averages; in particular, the length of time over which the averaging is performed should not be longer than a few seconds to a moderate fraction of an hour, depending on the type of circuit considered (whether tropospheric or ionospheric, for example). On most fading circuits the signal fluctuations appear to be a combination of a rapidly-varying Rayleigh-distributed component, and a slower-varying component apparently due to long-term variations in the transmission path. Diversity is intended principally to take advantage of lack of correlations in the rapid fluctuations of several signals. The averages described in (11a) and (11b) must therefore be taken over a time interval short enough to prevent inclusion of slow variations in the correlations.

The method described above, if practical at all, is probably usable only at frequencies below 50 mc. From an engineering viewpoint, it would be much more desirable to use envelope correlations where possible. We defer further discussion on this to a later and more appropriate section.

Density Function for the Combined Power

We note first that (3) can be rewritten as

$$S = \tfrac{1}{2}XIX^t \tag{12}$$

where I is the identity matrix. We use a technique similar to the characteristic function method to find $p(S)$, the density function for S.

For convenience, the Laplace transform of the density function $p(S)$ will be used since $p(S)=0$ for $S<0$. We write this transform as $P(z)$ with z the variable of the transformation:

$$P(z) = \int_0^\infty \exp(-zS)p(S)dS. \tag{13}$$

But this can be rewritten as a *mean value* or *expectation* E of the exponential:

$$\begin{aligned} P(z) &= E[\exp(-zS)] \\ &= E[\exp(-\tfrac{1}{2}zXIX^t)] \\ &= \int \cdots \int \exp(-\tfrac{1}{2}zXIX^t)p(X)dI_{c1}\cdots dI_{cN} \\ &= \int_{-\infty}^{\infty}\cdots\int_{-\infty}^{\infty}(2\pi)^{-N}|M|^{-1/2} \\ &\quad\cdot\exp[-\tfrac{1}{2}X(M^{-1}+zI)X^t]dI_{c1}\cdots dI_{sN}. \end{aligned}$$

Now $XM^{-1}X^t$ is a positive definite quadratic form, as is XIX^t so that $X(M^{-1}+zI)X^t$ is certainly positive definite for z real and non-negative. Therefore,[10]

$$\begin{aligned} P(z) &= |M|^{-1/2}|M^{-1}+zI|^{-1/2} \\ &= (|M||M^{-1}+zI|)^{-1/2} \\ &= |M(M^{-1}+zI)|^{-1/2} = |I+zM|^{-1/2}. \end{aligned}$$

However, it is seen from Appendix I that the determinant is a perfect square and that the result can be expressed as

$$P(z) = |I+zL|^{-1} \tag{14a}$$

where

$$L = \begin{bmatrix} a_1 & B_{12}^* & \cdots & B_{1N}^* \\ B_{12} & a_2 & \cdots & B_{2N}^* \\ B_{13} & B_{23} & \cdots & B_{3N}^* \\ \cdots & \cdots & \cdots & \cdots \\ B_{1N} & B_{3N} & \cdots & a_N \end{bmatrix} \tag{14b}$$

and

$$B_{jk} = b_{jk} + i\beta_{jk}.$$

It may be noted that the L matrix is Hermitian (equals its conjugate transpose). Thus $P(z)$ is a rational function of the form

$$P(z) = \left(\sum_{n=0}^{N} c_n z^n\right)^{-1}, \tag{14c}$$

where the coefficients c_n are those that result from evaluating the determinant.

If $P(z)$ is put in the form

$$P(z) = (-\lambda)^N|L-\lambda I|^{-1} \quad \text{where} \quad \lambda = -1/z, \tag{15}$$

we see that the poles of $P(z)$ are the negative reciprocals of the eigenvalues or latent roots of L. Since by the Hermitian property of L its eigenvalues are all real, and must also be positive because of the assumed positive definite nature of M, the poles of $P(z)$ are all negative real.[11]

To obtain $p(S)$ we have recourse to any of the usual inverse transform techniques. In general, since $P(z)$ has only poles, we can write

$$p(S) = \frac{1}{2\pi i}\oint \exp(zS)P(z)dz \tag{16}$$

where the contour runs counterclockwise and encloses all the poles of $P(z)$. It will usually be necessary to find the poles of $P(z)$ first by numerical means. In the simple general case where $P(z)$ has only simple poles, the result will be in the form

$$p(S) = \sum_{n=1}^{N} d_n \exp[z_n S] \tag{17}$$

where the z_n are the poles of $P(z)$, and the residues d_n are given by

$$d_n = \lim_{z\to z_n}[(z-z_n)P(z)]. \tag{18}$$

Thus, since the z_n are negative real, $p(S)$ is a sum of negative exponentials. For multiple poles, the corresponding results will again contain negative exponentials, but with polynomial multipliers. (Also note (44) later.)

In certain special cases, it will be sufficient to know only the Maclaurin series for $p(S)$ for computations involving small numerical values of S. This can be found without determining the poles of $P(z)$ and will frequently yield more accurate results in numerical work. We will write the Maclaurin series in the form

$$p(S) = \sum_{n=0}^{\infty}\frac{1}{n!}p^{(n)}(0+)S^n, \tag{19}$$

where

$$p^{(n)}(0+) = \lim_{S\to 0+}\left[\frac{d^n p(S)}{dS^n}\right]. \tag{20}$$

Both the limiting operation and differentiation can be performed using the transform and the rules

$$\mathcal{L}\left[\frac{df(S)}{dS}\right] = z\mathcal{L}[f(S)] - f(0+), \tag{21}$$

$$\lim_{S\to 0+} f(S) = \lim_{z\to\infty} zF(z).$$

[10] *Ibid.*, p. 119, sec. 11.12.

[11] A result similar to (14a) was described by R. Price (in a verbal communication) by specializing a result of Kac and Siegert. See especially M. Kac and A. J. F. Siegert, "On the theory of noise in radio receivers with square law detectors," *J. Appl. Phys.*, vol. 18, p. 396; April, 1947.

The situation here can be taken to be analogous to narrow-band noise passed through a square-law detector and thence through a post-detection filter whose impulse response consists of N impulses spaced in time. The autocorrelation in time of the noise corresponds to the spatial autocorrelation of the fading signal; the square-law detector (after removing high-frequency components) produces an output proportional to the slowly varying power level of the noise, directly analogous to the variations in space of the diversity signals; finally, the multiple impulse response of the filter corresponds to summing the power at multiple points in space.

Moreover, we see that (14b) can be rewritten as a Laurent expansion valid for all z of magnitude larger than the largest pole of $P(z)$, and of the form

$$P(z) = z^{-N}\left[\sum_{n=0}^{\infty} k_n \left(\frac{1}{z}\right)^n\right] \tag{22}$$

(obtainable, for example, by ordinary long division). Applying the differentiation and limit rules, (21) gives

$$p^{(n)}(0+) = \begin{cases} 0 & n < N-1 \\ k_{n-N+1} & n \geq N-1 \end{cases}$$

and

$$p(S) = \sum_{n=N-1}^{\infty} \frac{1}{n!} k_{n-N+1} S^n. \tag{23}$$

Distribution Function and Mean Power

The mean power can be found by recalling that

$$\frac{d}{dz}\mathcal{L}[p(S)] = -\int_0^\infty Se^{-zS}p(S)dS, \tag{24}$$

$$-\left.\frac{dP(z)}{dz}\right|_{z=0} = \int_0^\infty Sp(S)dS = \bar{S}. \tag{25}$$

Hence the average received power can be found quite simply from the transform.

The cumulative distribution function can be obtained by integrating (19) or (23); the latter may be truncated to yield a polynomial approximation. Thus

$$\text{Prob}\,(S < S_t) = \int_0^{S_t} p(S)dS. \tag{26}$$

Performance Estimation From Envelope Correlations

It was pointed out earlier that it would be desirable to make use of envelope correlations rather than component correlations both because of the greater availability of data on the envelope correlations, and because of the difficulty of performing the measurements necessary to determine component correlations. In this section we indicate how this envelope data can be used. Let us denote the envelope amplitude of the various signals as R_j, defined by

$$R_j = \sqrt{2S_j}. \tag{27}$$

The normalized correlations of the power and envelope respectively are then

$$\mu_S(j,k) = \overline{\left[\frac{S_j - \bar{S}_j}{\sqrt{\overline{(S_j - \bar{S}_j)^2}}}\right]\left[\frac{S_k - \bar{S}_k}{\sqrt{\overline{(S_k - \bar{S}_k)^2}}}\right]} \tag{28a}$$

$$\mu_R(j,k) = \overline{\left[\frac{R_j - \bar{R}_j}{\sqrt{\overline{(R_j - \bar{R}_j)^2}}}\right]\left[\frac{R_k - \bar{R}_k}{\sqrt{\overline{(R_k - \bar{R}_k)^2}}}\right]}. \tag{28b}$$

Since

$$p(S_j) = \frac{1}{\bar{S}_j}\exp(-S_j/\bar{S}_j)$$

$$p(R_j) = \frac{R_j}{\bar{S}_j}\exp(-R_j^2/2\bar{S}_j),$$

it is easily shown that

$$\mu_S(j,k) = \frac{\overline{S_jS_k}}{\bar{S}_j\bar{S}_k} - 1 \tag{29a}$$

$$\mu_R(j,k) = \frac{\overline{R_jR_k} - \frac{\pi}{2}\sqrt{\bar{S}_j\cdot\bar{S}_k}}{\left(2 - \frac{\pi}{2}\right)\sqrt{\bar{S}_j\cdot\bar{S}_k}}. \tag{29b}$$

If we combine (29a) with previously known results,[12] we find

$$\mu_S(j,k) = \frac{b_{jk}^2 + \beta_{jk}^2}{a_ja_k}, \tag{30a}$$

or in terms of the notation in the complex matrix L, in (14b),

$$\mu_S(j,k) = \frac{|\boldsymbol{B}_{jk}|^2}{a_ja_k}. \tag{30b}$$

However, by a trivial extension of a result given by Lawson and Uhlenbeck[13] we can express the envelope correlation as a function of the power correlation:

$$\mu_R(j,k) = f[\mu_S(j,k)]. \tag{31}$$

This function is given in tabular form in Appendix V. It is seen that for all practical purposes the two correlations may be taken as approximately equal. (That this is true for small values of the correlation can be shown directly from power series expansion of the functional form in (31). The table, however, shows the statement to be true over the entire range of the normalized correlation.)

For convenience in later work we define

$$\rho_{jk} \equiv \sqrt{\mu_S(j,k)} \tag{32a}$$

so that

$$\boldsymbol{B}_{jk} = \rho_{jk}\sqrt{a_ja_k}\exp(i\phi_{jk}), \tag{32b}$$

or equivalently,

$$\begin{aligned} b_{jk} &= \rho_{jk}\sqrt{a_ja_k}\cos\phi_{jk} \\ \beta_{jk} &= \rho_{jk}\sqrt{a_ja_k}\sin\phi_{jk}. \end{aligned} \tag{32c}$$

Thus, given envelope correlations from which we can find power correlations by using Appendix V, the complex correlations $\boldsymbol{B}_{jk}$ are known to within an arbitrary angle ϕ_{jk}.

[12] D. E. Kerr, Ed., "Propagation of Short Radio Waves," McGraw-Hill Book Co., Inc., New York, N. Y., p. 562; 1947.

[13] J. A. Lawson and G. E. Uhlenbeck, "Threshold Signals," McGraw-Hill Book Co., Inc., New York, N. Y., p. 62; 1952.

For engineering purposes we can suggest that if envelope correlations only are known, the phase angles, ϕ_{jk}, should be assumed to be zero. Thus

$$B_{jk}(=)\rho_{jk}\sqrt{a_j a_k}, \text{ when } \phi_{jk} \text{ is unknown,} \tag{33}$$

where the (=) is to be read "assumed equal to."

The effect of the assumption (33) is difficult to evaluate. For time diversity no error is introduced in the final result (14) if the spectrum of the fading is symmetrical, and (33) is used. It can also be shown that if the L matrix is completely specified, no change in (14) will result if a row of L is multiplied by $\exp(i\theta)$ and the corresponding column by $\exp(-i\theta)$. Thus the end result is unaltered by the assumption of zero phase angles if this can be brought about by a number of such row and column operations. (This last remark is simply a restatement of the fact that the eigenvalues of M are invariant under a rigid rotation of coordinates.) The permissibility of these changes also shows that the phase angle of $\boldsymbol{B}$ is immaterial in dual diversity.

As an example of a situation where the phase angle *is* significant, consider the matrix

$$L = \begin{bmatrix} 1 & \rho_{12} & \rho_{13}e^{-i\phi} \\ \rho_{12} & 1 & \rho_{23} \\ \rho_{13}e^{i\phi} & \rho_{23} & 1 \end{bmatrix}$$

for which

$$|I + zL| = (z+1)^3 - z^2(z+1)(\rho_{12}^2 + \rho_{13}^2 + \rho_{23}^2) + 2z^3\rho_{12}\rho_{23}\rho_{13}\cos\phi.$$

It is readily seen that the roots of this polynomial are a function of ϕ.

Finally, as a further argument on the plausibility of (33), we note that for certain values of the ρ_{jk}, not all choices of ϕ_{jk} lead to permissible correlation matrices; on the other hand, taking $\phi_{jk}=0$, they always will lead to a permissible correlation matrix when the ρ_{jk} are determined from measured correlations.

In all cases it is felt that the assumption (33) will lead to a more realistic estimate of diversity performance than does the assumption of independent fading.

EXAMPLES

Example 1: Independent Fading

For independent fading with identical antennas, the coefficients (9c) are given by

$$a_n = S_0, \qquad b_{nm} = \beta_{nm} = 0, \tag{34}$$

where S_0 is the mean power at any receiver, so that

$$|I + zL| = (1 + zS_0)^N$$

and

$$p(S) = \frac{1}{2\pi i}\oint \frac{\exp(zS)dz}{(1+zS_0)^N} \tag{35}$$

which gives

$$p(S) = \frac{S^{N-1}\exp(-S/S_0)}{(N-1)!S_0^N}. \tag{36}$$

This is exactly the well-known result cited earlier[3] for n-fold maximal-ratio combining of statistically independent signals which has also been recognized as a representation for the Chi-Square process of order $2N$.

Example 2: Dual Diversity

Let the reduced, L, matrix for the variates be given by

$$L = \begin{bmatrix} S_0 & b + i\beta \\ b - i\beta & S_0 \end{bmatrix} \tag{37}$$

where we again take the received powers as S_0 in either receiver. Then

$$|I + zL| = [(1 + zS_0)^2 - z^2(b^2 + \beta^2)] \tag{38}$$

or

$$P(z) = \frac{1}{(1 + zS_0 + z\rho S_0)(1 + zS_0 - z\rho S_0)} \tag{39}$$

where

$$\rho = \sqrt{b^2 + \beta^2}/S_0.$$

[The coefficient ρ is defined as in (32a).] Expanding this in partial fractions,

$$P(z) = \frac{(1+\rho)/2\rho}{1 + z(1+\rho)S_0} - \frac{(1-\rho)/2\rho}{1 + z(1-\rho)S_0}$$

and

$$p(S) = \frac{\exp[-S/(1+\rho)S_0] - \exp[-S/(1-\rho)S_0]}{2\rho S_0}. \tag{40}$$

This coincides with the previously published result for this special case[6] and shows also that indeed, for dual diversity, the statement of (33) is always applicable.

Example 3: A Typical Numerical Example

Suppose a one-way ionospheric scatter circuit employing a high gain transmitting antenna is to be converted to two-way service by using the high gain antenna for receiving, supplemented by low gain antennas on each side of the high gain antenna. Two low gain units are temporarily placed adjacent to the large array and at a weak signal time of the year, the following moment matrix is measured:

$$M = \begin{bmatrix} 1/9 & 0 & 1/6 & 1/6 & 1/30 & 1/90 \\ 0 & 1/9 & -1/6 & 1/6 & -1/90 & 1/30 \\ 1/6 & -1/6 & 9/8 & 0 & 7/30 & -1/30 \\ 1/6 & 1/6 & 0 & 9/8 & 1/30 & 7/30 \\ 1/30 & -1/90 & 7/30 & 1/30 & 1/9 & 0 \\ 1/90 & 1/30 & -1/30 & 7/30 & 0 & 1/9 \end{bmatrix}.$$

(The moments are all in units of $m\mu\mu$ watts in a 50 ohm load. Actual experimental values would of course be in decimal values but for convenience of hand calculation fractions are used here.) The determinant (14) is evaluated and found to be

$$|I+zL| = \left(\frac{6480 + 8730z + 972z^2 + 25z^3}{6480}\right).$$

The roots z_a are found by trial and error and $P(z)$ factored into

$$P(z) = 1/\left(1+\frac{z}{25.67}\right)\left(1+\frac{z}{12.40}\right)\left(1+\frac{z}{.8146}\right)$$
$$= \frac{0.03060}{1+\dfrac{z}{25.67}} - \frac{0.1360}{1+\dfrac{z}{12.40}} + \frac{1.105}{1+\dfrac{z}{0.8146}};$$

from which we find

$$p(S) = +0.7856 \exp(-25.67S) - 1.686 \exp(-12.40S) + 0.9005 \exp(-0.8146S).$$

Alternatively, the Laurent expansion of $P(z)$ may be found:

$$P(z) = \frac{259.2}{z^3}\left[1 - 3.888\left(\frac{10}{z}\right) + 11.62\left(\frac{10}{z}\right)^2 - 31.88\left(\frac{10}{z}\right)^3 + 84.36\left(\frac{10}{z}\right)^4 - 219.7\left(\frac{10}{z}\right)^5 \cdots\right]$$

which yields the power series expansion of $p(S)$:

$$p(S) = 1.296[(10S)^2 - 1.296(10S)^3 + 0.9687(10S)^4 - 0.5313(10S)^5 + 0.2343(10S)^6 - 0.08717(10S)^7 \cdots]$$

In both of the expressions S is in $m\mu\mu$ watts. The series expansion truncated to the seventh degree polynomial shown will yield results accurate to within 0.5 per cent for S as large as 1/20 $m\mu\mu$ watt or 14.3 db below mean power. Further, since

$$P(z) = \frac{6480}{6480 + 8730z + 972z^2 + 25z^3},$$

we have directly that

$$P'(z) = \frac{-6480(8730 + 1944z + 75z^2)}{(6480 + 8730z + 972z^2 + 25z^3)^2}$$

and

$$\bar{S} = -P'(z)\Big|_{z=0} = \frac{8730}{6480} = 1.347 \ m\mu\mu \text{ watts}$$

or slightly more (0.8 db) than the high gain array alone.

The distribution function may be found exactly as

$$\text{Prob}(S<S_t) = 1 - 0.03060 \exp(-26.67S_t) + 0.1360 \exp(-12.40S_t) - 1.105 \exp(-0.8146S_t),$$

or approximately as

$$\text{Prob}(S<S_t) = 43.2(S_t^3 - 9.72S_t^4 + 58.02S_t^5 - 265.7S_t^6 + 1004S_t^7 - 3269S_t^8 \cdots).$$

Suppose we have a receiving system, to be connected to this diversity array, with a threshold of 1/20 $m\mu\mu$ watt (−163 dbw). The polynomial approximation yields sufficiently accurate results and we obtain Prob($S<1/20$) = 0.00341, or an outage time of 0.34 per cent. The high gain array alone will give an outage time of 4.35 per cent, so on paper, at least, our imaginary system has been improved considerably.

At the other extreme, to get an indication of dynamic range required, we might want to find the percentage of time the received power exceeds ten times the average power. In this case we would certainly use the exact expression and obtain

$$\text{Prob}(S<13.47) = 1 - 1.9\times 10^{-5},$$
$$p(S>13.47) = 0.0019 \text{ per cent.}$$

(In passing it should be remarked that the Rayleigh approximation may itself be inaccurate at very large signal levels.)

Simplified Calculation Methods in Special Cases

In the results obtained, we have seen that the basic information about the distributions in question lies in the poles of the Laplace transform, $P(z)$. As seen already, in many cases it will be most convenient to determine these poles by routinely evaluating the form $|I+zL|^{-1}$ (or possibly $|I+zM|^{-1/2}$ if computation of the real matrix is more convenient), and numerically solving for the denominator roots. In certain special cases, however, simpler forms may be available, or it may also be simpler to determine the poles of $P(z)$ as eigenvalues of the matrix when the latter has special properties. We indicate below some examples of such cases.

Symmetrical Configurations

A simplification in computations is available when the diversity "array" (be it diversity in space, frequency, time, or whatever) has a physical symmetry. Specifically, it is required that the correlations among the signals be the same whether they be taken in their first order, or in completely reverse order. Mathematically, if the diversity signals are ordered $i=1, \cdots, N$ (2 quadrature components for each), the requirement is that in the definitions of (9)

$$b_{mn} = b_{N+1-m,N+1-n}$$
$$\beta_{mn} = \beta_{N+1-m,N+1-n}. \tag{41}$$

Physically, such a symmetry in the diversity array will often be the case in practice.[14] It is shown in Appendix II that under the conditions stated, $P(z)$ can further be written in the form

$$P(z) = |I + zL^{(e)}|^{-1}|I + zL^{(o)}|^{-1} \tag{42}$$

where each of the determinants is now a polynomial approximately of order $N/2$, whereas the matrix L was order $N \times N$. Specifically the elements of $L^{(e)}$ and $L^{(o)}$ are defined in terms of those of L as follows:

$$\text{If } N \text{ even: } \quad L^{(e)} \text{ is } \left(\frac{N}{2} \times \frac{N}{2}\right) \text{ matrix}$$

$$L^{(o)} \text{ is } \left(\frac{N}{2} \times \frac{N}{2}\right) \text{ matrix}$$

and

$$L_{mn}^{(e)} = L_{mn} + L_{m,N+1-n}, (m, n) \leq N/2$$

$$L_{mn}^{(o)} = L_{mn} - L_{m,N+1-n}, (m, n) \leq N/2. \tag{43a}$$

$$\text{If } N \text{ odd: } \quad L^{(e)} \text{ is } \left(\frac{N+1}{2} \times \frac{N+1}{2}\right) \text{ matrix}$$

$$L^{(o)} \text{ is } \left(\frac{N-1}{2} \times \frac{N-1}{2}\right) \text{ matrix}$$

and

$$L_{mn}^{(e)} = \begin{Bmatrix} L_{mn} + L_{m,N+1-n}, n < (N+1)/2 \\ L_{mn} \qquad\qquad\quad, n = (N+1)/2 \end{Bmatrix}$$

$$L_{mn}^{(o)} = L_{mn} - L_{m,N+1-n}, (m, n) \leq (N-1)/2. \tag{43b}$$

Solution for the Eigenvalues

As has been suggested earlier, the transform $P(z)$ is completely determined by the eigenvalues of the matrix L. In matrix notation, this can be expressed as follows, where $\lambda_j (j = 1, \cdots, N)$ are the N eigenvalues of L.

If we assume Q is the unitary matrix which diagonalizes L (and which must exist due to the Hermitian nature of L), we have

$$\begin{aligned} P(z) &= |I + zL|^{-1} = \{|Q|^{-1}|I + zL||Q|\}^{-1} \\ &= |Q^{-1}Q + zQ^{-1}LQ|^{-1} = |I + z\Lambda|^{-1} \\ &= \prod_{j=1}^{N} (1 + z\lambda_j)^{-1}. \end{aligned} \tag{44}$$

[14] It has been pointed out that since $\beta_{mn} = -\beta_{nm}$ always, the conditions stated include the statement that $\beta_{mn} = \beta_{nm} = 0$ whenever $m+n = N+1$. This is, of course, also implied physically by the symmetry assumption.

The last step follows since $\Lambda = Q^{-1}LQ$ has only elements along the main diagonal and these are individually exactly the eigenvalues of L.

If, now, these eigenvalues can be given explicitly for a particular form of the matrix, L, there will obviously be much labor saved towards obtaining general information about the distribution of the diversity-combined resultant signal. We will indicate two particular cases, which appear to be of more than passing engineering interest, and for which direct information about the eigenvalues can be obtained:

Equi-Spaced Diversity: Often the diversity signals will be taken from a configuration which, in some physical sense, represents equispaced sampling (either in time, space, frequency, etc.). From the stationary nature of the over-all process, and assuming that all signals are statistically equivalent, we would expect in such a system to have

$$b_{nm} = f[|n - m|\tau]$$
$$\beta_{nm} = g[|n - m|\tau]$$

where τ is some parameter describing the "spacing." The L matrix then has only n independent elements, since each slant line parallel to the main diagonal has all its elements the same. Although the general closed-form solution for the eigenvalues of even such a matrix is not known, it is possible to give a solution for the particular case of a modified exponential correlation. Namely, if

$$b_{nm} = \sigma^2 \exp[-\alpha|n - m|\tau] \cos[\beta|n - m|\tau]$$

and

$$\beta_{nm} = \sigma^2 \exp[-\alpha|n - m|\tau] \sin[\beta|n - m|\tau], \tag{45}$$

we have

$$L_{mn} = \begin{cases} b_{nm} + i\beta_{nm} = \sigma^2 \exp[-(\alpha - i\beta)|n - m|\tau] & m \geq n \\ b_{nm} - i\beta_{nm} = \sigma^2 \exp[-(\alpha + i\beta)|n - m|\tau] & m \leq n. \end{cases} \tag{46}$$

Or, if we set

$$q = \exp[-(\alpha - i\beta)\tau],$$

$$L_{mn} = \begin{cases} \sigma^2 q^{|n-m|} & m \geq n \\ \sigma^2 (q^*)^{|n-m|} & m \leq n \end{cases} \tag{47}$$

where the asterisk denotes a complex conjugate.

This correlation function is closely related to that for a Gaussian Markov process, and in fact represents a type of Markov process among the n band-pass signals under consideration. In Appendix III, we show that the eigenvalues of L in this case can be explicitly stated, in the form:

$$\lambda_i = \sigma^2 \frac{(1 - |q|^2)}{1 + |q|^2 + 2|q| \cos \phi_i} \tag{48a}$$

where the ϕ_i are the solutions of the transcendental equation

$$\tan N\phi = \frac{-\sin\phi}{\cos\phi\left(\frac{1+|q|^2}{1-|q|^2}\right)+\frac{2|q|}{1-|q|^2}}. \quad (48b)$$

This result is a slight generalization of one published earlier[15] without proof.

Correlation Only Among Adjacent Signals: In engineering diversity applications, the "array" is usually designed to approach the ideal of statistical independence among the individual signals. If there is residual correlation, it will often be significant only among physically adjacent members of the array. For this case, when additionally all signals are of equal level, it is shown in Appendix III that the eigenvalues are given by

$$\lambda_k = \sigma^2\left[1 - 2|q|\cos\frac{k\pi}{N+1}\right], \quad k = 1, \cdots, N. \quad (49)$$

As discussed in Appendix III, this result has no significance for $|q| > \frac{1}{2}$. As a practical matter, moreover, we would probably require $|q| < 0.1$ for the assumed matrix to be realistic.

Two Applications to Digital Communications

We will consider two methods of binary communication: phase reversal keying with coherent diversity combination reception and frequency shift keying with square law diversity at the receiver. Matched filter detection is presumed with fading sufficiently slow so that output samples from the filters have statistics similar to those of a continuous wave.

It has been shown[16] that with *independent* Rayleigh fading, and with samples uncorrelated from pulse to pulse the square-law combination method is optimum for FSK if the receiver is unable to extract phase information. Similarly, by extension of analyses[16,17] for coherent detection of FSK, it can be shown that Brennan's method of weighting signals is also maximally efficient for diversity reception of phase reversal keying if signal amplitude is known exactly, and if signal phase is known exactly for either binary element.

It is perhaps questionable whether either of these two systems would be used on a particular fading circuit; the success of the phase-reversal system is dependent on the integration time available for determining expected amplitude and phase, while for FSK, the square-law combination is not optimum under either sufficiently slow fading or under nonindependent fading in the several diversity branches. Nevertheless, computation of error rates for these special cases is valuable since they provide bounds on the performance of an actual binary system to the extent that one need not do worse than with FSK and square-law combination, and cannot do better than phase reversal keying with coherent combination. Additionally, the FSK system is representative of a large class of orthogonal pulse systems including certain bandspreading techniques with time diversity.

Both of these systems prove to have error rates which are functions only of the sum of the several signals powers and therefore can be handled using the results up to this point.

We define

T = pulse length, seconds
n_0 = noise density, watts per cps.

Phase Reversal Keying

Reiger[18] gives the error probability in a phase-reversal system as

$$p(e \mid S) = \frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/n_0}}^{\infty} \exp(-t^2)\,dt. \quad (50)$$

To obtain average error rate we take the expected value of this conditional error probability:

$$p(e) = \int_0^{\infty} p(e \mid S)p(S)dS. \quad (51)$$

which on substituting (17) becomes

$$p(e) = \int_0^{\infty}\left[\frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/n_0}}^{\infty}\exp(-t^2)dt\right]\left[\sum_{n=1}^{N} d_n \exp(z_nS)\right]dS,$$

when the z_n's are all negative and distinct. Thus

$$p(e) = \sum_{n=1}^{N} d_n \int_0^{\infty} dS\exp(z_nS)\frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/n_0}}^{\infty}\exp(-t^2)dt$$
$$= \sum_{n=1}^{N} d_n\left(-\frac{1}{2z_n} + \frac{1}{2z_n\sqrt{1-\frac{z_nn_0}{T}}}\right),$$

by integrating by parts, so that

$$p(e) = \sum_{n=1}^{N}\frac{d_n}{-2z_n}\left(1 - \frac{1}{\sqrt{1-\frac{z_nn_0}{T}}}\right). \quad (52)$$

In the case of repeated roots, we will have terms in $p(S)$ of the form

$$p(S) = d_{mn}S^m\exp(z_nS) + \cdots. \quad (53a)$$

The contribution to average error rate of these terms can be evaluated by

[15] S. Stein and J. E. Storer, "Generating a Gaussian sample," IRE Trans. on Information Theory, vol. IT-2, pp. 87-90; June, 1956.

[16] J. N. Pierce, "Theoretical diversity improvement in frequency-shift keying," Proc. IRE, vol. 46, pp. 903–910; May, 1958.

[17] H. B. Law, "The detectability of fading radio telegraph signals in noise," *Proc. IEE*, vol. 104, pt. B, pp. 130–140; March, 1957.

[18] S. Reiger, "Error probabilities of binary data transmission systems in the presence of random noise," 1953 IRE Convention Record, pt. 8, pp. 72–79.

$$\int_0^\infty S^m \exp(z_n S)\left[\frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/n_0}}^\infty \exp(-t^2)dt\right]dS$$

$$= \left[\frac{d^m}{dz^m}\int_0^\infty dS \exp(zS)\frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/n_0}}^\infty \exp(-t^2)dt\right]_{z=z_n}$$

$$= \left[\frac{d^m}{dz^m}\left(-\frac{1}{2z}+\frac{1}{2z\sqrt{1-\frac{zn_0}{T}}}\right)\right]_{z=z_n}. \quad (53b)$$

On the other hand, we can obtain an expansion valid for very small error rates, without the necessity of first finding the z_n by using the series (23):

$$p(e) = \int_0^\infty\left[\sum_{n=N-1}^\infty \frac{1}{n!}k_{n-N+1}S^n\right]$$

$$\cdot\left[\frac{1}{\sqrt{\pi}}\int_{\sqrt{ST/n_0}}^\infty \exp(-t^2)dt\right]dS. \quad (54a)$$

The double integral can be evaluated by standard means, so that finally,

$$p(e) = \frac{1}{2}\sum_{n=N}^\infty \frac{(2n)!}{4^n(n!)^2}k_{n-N}\left(\frac{n_0}{T}\right)^n. \quad (54b)$$

The convergence of this series, as well as the approximation properties of its partial sums, is not nearly as satisfying as other series discussed here. For large n the coefficients are approximately

$$\frac{(2n)!}{4^n(n!)^2}k_{n-N} \approx \frac{k_{n-N}}{\sqrt{\pi n}},$$

so that the convergence of (54b) is essentially the same as that of (22). Hence we require at least that

$$\left(\frac{n_0}{T}\right) < \frac{1}{\mathrm{Max}(-z_n)}; \quad (54c)$$

in many cases, then, (54) will only be valid for error rates which are so small as to be unrealizable on other grounds. [It will be recognized that (54c) is necessary and sufficient to justify the interchange of summation and integration in the original integral, by dominant convergence. The partial sums of the first bracket are dominated in absolute value by $C_1 \exp[S\,\mathrm{Max}(-z_n)]$; the second bracket is dominated by $C_2 \exp(-ST/n_0)$, where C_1, C_2 are constants, so that the result follows.]

Example: Suppose that we are to use the fictitious system of the previous numerical examples to receive phase-reversal keying with a pulse length of $T=1/2$ msec and with a noise density of 10^{-4} mμμ watts/cps. Substituting these values in (52) we obtain

$$p(e) = \frac{1}{2}\left[0.03060\left(1-\frac{1}{\sqrt{1+25.67(1/5)}}\right)\right.$$

$$-0.1360\left(1-\frac{1}{\sqrt{1+12.40(1/5)}}\right)$$

$$\left.+01.105\left(1-\frac{1}{\sqrt{1+0.8146(1/5)}}\right)\right]$$

$$= 0.0178$$

Frequency-Shift Keying: Square-Law Combination

It is shown in Appendix IV that the error probability for FSK with square-law combining is given by

$$p(e\mid S) = \exp\left(-\frac{ST}{2n_0}\right)\sum_{m=0}^{N-1}\frac{1}{m!}\gamma_m\left(\frac{ST}{2n_0}\right)^m, \quad (55a)$$

with

$$\gamma_m = \sum_{n=m}^{N-1}\frac{(N-1+n)!2^{-N-n}}{(N-1+m)!(n-m)!}. \quad (55b)$$

The result is in terms of S, the sum of the several signal powers. It should be stressed that the density function (16), for example, is the density function for the sum of the received powers no matter how they are combined—in fact, it is valid even if the various signals are not combined at all. Hence, we write

$$p(e) = \int_0^\infty p(e/S)_p(S)dS$$

$$= \int_0^\infty p(S)\exp\left(-\frac{ST}{2n_0}\right)\left[\sum_{m=0}^{N-1}\frac{1}{m!}\gamma_m\left(\frac{ST}{2n_0}\right)^m\right]dS$$

$$= \sum_{m=0}^{N-1}\frac{1}{m!}\gamma_m\left(\frac{T}{2n_0}\right)^m\int_0^\infty S^m\exp\left(-\frac{ST}{2n_0}\right)p(S)ds$$

$$= \sum_{m=0}^{N-1}\frac{1}{m!}\gamma_m\left(\frac{T}{2n_0}\right)^m$$

$$\left[\left(-\frac{d}{dz}\right)^m\int_0^\infty \exp(-zS)p(S)dS\right]_{z=T/2n_0},$$

so that the end result is

$$p(e) = \sum_{m=0}^{N-1}\frac{1}{m!}\gamma_m\left(-\frac{T}{2n_0}\right)^m\left|P^{(m)}(T/2n_0)\right|, \quad (56)$$

$P^{(m)}$ being the mth derivative of $P(z)$.

This result is much simpler in that we need not find the poles of $P(z)$ or evaluate $p(S)$ to obtain an exact result. This exact result can be obtained by substituting (14) in (56).

Example: Suppose that with our previously assumed parameters we again take $n_0=10^{-4}$ mμμ watts/cps but this time we solve for the error rate using FSK with square-law combination and a pulse length of 2 msec. Normalized to the signal power we have $T/2n_0=10$ and since for $N=3$, we have $\gamma_0=1/2$, $\gamma_1=3/16$, $\gamma_2=1/32$,

$$p(e) = \frac{1}{2} P(10) - \frac{30}{16} P'(10) + \frac{100}{64} P''(10) = 0.0261,$$

or an error rate of 2.61 per cent.

Summary

A general analysis has been outlined for estimation of multiple-diversity performance in the case of nonindependent fading among the signals. Numerical computations are obtained from the moment matrix, whose elements are the covariances among the signals, and which are based either on theoretical predictions or on direct physical measurements. The possibility of using only envelope correlation data, as is more usually measured, has also been discussed. Various numerical "short-cuts" have been indicated for obtaining certain types of data, and results for some special cases given in full. Other basic properties of the calculations have been described in terms of eigenvalues of the moment matrix, and examples given of the nature of such eigenvalues in special cases. Finally, applications of the results have been described for error-rate calculations on two digital systems—phase reversal keying with optimum coherent diversity combining and FSK with square-law combination.

Appendix I

Reduction of the Moment Matrix

The eigenvalues of the moment matrix are determined from the characteristic determinantal equation

$$|M - \lambda I| = 0$$

where the matrix $M-\lambda I$, is of the form

$$M - \lambda I = \begin{bmatrix} A_1 - \lambda I & B_{12} & \cdots & B_{1N} \\ B_{12}' & A_2 - \lambda I & \cdots & B_{2N} \\ B_{13}' & B_{23}' & \cdots & B_{3N} \\ \cdots & \cdots & \cdots & \cdots \\ B_{1N}' & B_{2N}' & \cdots & A_N - \lambda I \end{bmatrix} \tag{57}$$

with

$$A_j - \lambda I = \begin{bmatrix} a_j - \lambda & 0 \\ 0 & a_j - \lambda \end{bmatrix}, \qquad B_{jk} = \begin{bmatrix} b_{jk} & \beta_{jk} \\ -\beta_{jk} & b_{jk} \end{bmatrix}.$$

Let

$$Q = \begin{bmatrix} C & 0 & 0 & \cdots & 0 \\ 0 & C & 0 & \cdots & 0 \\ 0 & 0 & C & \cdots & 0 \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ 0 & 0 & 0 & \cdots & C \end{bmatrix}$$

$$\text{with} \quad C = \sqrt{1/2} \begin{bmatrix} 1 & i \\ i & 1 \end{bmatrix},$$

$$i^2 = -1. \tag{58}$$

Now $|Q| = 1$ so $|Q(M-\lambda I)Q^{-1}| = |M-\lambda I|$. But $Q(M - \lambda I)Q^{-1} =$

$$\begin{bmatrix} A_1 - \lambda I & D_{12} & D_{13} & \cdots & D_{1N} \\ D_{12}^* & A_2 - \lambda I & D_{23} & \cdots & D_{2N} \\ D_{13}^* & D_{23}^* & A_3 - \lambda I & \cdots & D_{3N} \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ D_{1N}^* & D_{2N}^* & D_{3N}^* & \cdots & A_N - \lambda I \end{bmatrix} \tag{59}$$

where the asterisk indicates complex conjugation, and

$$D_{jk} = \begin{bmatrix} (b_{jk} - i\beta_{jk}) & 0 \\ 0 & (b_{jk} + i\beta_{jk}) \end{bmatrix}. \tag{60}$$

Thus $Q(M-\lambda I)Q^{-1}$ has a checkerboard array of zeros, so that by permuting rows and columns we can obtain

$$|Q(M - \lambda I)Q^{-1}| = \begin{vmatrix} L - \lambda I & 0 \\ 0 & (L - \lambda I)' \end{vmatrix} \tag{61}$$

where

$$L = \begin{bmatrix} a_1 & \mathbf{B}_{12}^* & \mathbf{B}_{13}^* & \cdots & \mathbf{B}_{1N}^* \\ \mathbf{B}_{12} & a_2 & \mathbf{B}_{23}^* & \cdots & \mathbf{B}_{2N}^* \\ \mathbf{B}_{13} & \mathbf{B}_{23} & a_3 & \cdots & \mathbf{B}_{3N}^* \\ \cdots & \cdots & \cdots & \cdots & \cdots \\ \mathbf{B}_{1N} & \mathbf{B}_{2N} & \mathbf{B}_{3N} & \cdots & a_N \end{bmatrix} \tag{62}$$

and $\mathbf{B}_{jk} = b_{jk} + i\beta_{jk}$. Finally, therefore,

$$|M - \lambda I| = |L - \lambda I|^2$$

and (63)

$$|I + zM| = |I + zL|^2$$

thus showing that the eigenvalues are all double, and also giving an alternate computational form for $P(z)$.

An alternate proof can be given by adding and subtracting rows, and columns, so that the first two columns and first two rows of the determinant are zero except for the diagonal elements. The resulting determinant can be shown to be of the same form as the original one so that the desired result can be proven by induction on N.

Appendix II

Added Properties of Physically Symmetric Array

The eigenvalues $\lambda^{(k)}$ and the normalized column eigenvectors $x^{(k)}$ for the L matrix satisfy the matrix equation

$$Lx^{(k)} = \lambda^{(k)} x^{(k)}. \tag{64}$$

Or, in terms of components,

$$\sum_{j=1}^{N} L_{mj} x_j^{(k)} = \lambda^{(k)} x_m^{(k)}, \quad m = 1, \cdots, N. \tag{65}$$

Let us now consider a column vector $y^{(k)}$ defined by reversing the components of $x^{(k)}$, *i.e.*,

$$y_j^{(k)} = x_{N+1-j}^{(k)}. \tag{66}$$

Then if we perform the multiplication

$$Ly^{(k)} = z^{(k)}, \tag{67}$$

we have

$$z_m^{(k)} = \sum_{j=1}^{N} L_{mj}y_j^{(k)} = \sum_{j=1}^{N} L_{mj}x_{N+1-j}^{(k)}$$
$$= \sum_{p=1}^{N} L_{m,N+1-p}x_p^{(k)}. \tag{68}$$

Thus, if we now assume our physical symmetry property that for all m, n

$$L_{mn} = L_{N+1-m,N+1-n}, \tag{69}$$

we have

$$L_{m,N+1-p} = L_{N+1-m,p} \tag{70}$$

and

$$z_m^{(k)} = \sum_{p=1}^{N} L_{N+1-m,p}x_p^{(k)}. \tag{71}$$

We have in fact thus proven the result

$$z_m^{(k)} = \lambda^{(k)}x_{N+1-m}^{(k)} \tag{72}$$

or

$$Ly^{(k)} = \lambda^{(k)}y^{(k)}; \tag{73}$$

i.e., both $y^{(k)}$ and $x^{(k)}$ are possible eigenvectors of L, with the same eigenvalue. Now, this can be a trivial statement; namely, if the $x^{(k)}$ eigenvector itself inherently contains a symmetry such that $x_{N+1-j}^{(k)} = x_j^{(k)}$, or $x_{N+1-j}^{(k)} = -x_j^{(k)}$ for all j, then indeed $y^{(k)}$ is just $x^{(k)}$ to within an irrelevant change of sign. On the other hand, if $\lambda^{(k)}$ can be a degenerate eigenvalue, then $x^{(k)}$ and $y^{(k)}$ can be linearly independent and neither need be symmetric. But in this case, it is then readily shown that the vectors

$$u^{(k)} = \frac{x^{(k)} + y^{(k)}}{\sqrt{2}}, \qquad v^{(k)} = \frac{x^{(k)} - y^{(k)}}{\sqrt{2}} \tag{74}$$

are also linearly independent, and hence are also legitimate alternative eigenvectors for the degenerate eigenvalue $\lambda^{(k)}$. But $u_j^{(k)} = u_{N+1-j}^{(k)}$ and $v_j^{(k)} = -v_{N+1-j}^{(k)}$, by inspection. Hence we have proven that all eigenvectors of the L matrix can be represented as having either an odd symmetry

$$x_{N+1-j}^{(k)} = -x_j^{(k)} \tag{75}$$

or an even symmetry

$$x_{N+1-j}^{(k)} = x_j^{(k)}.$$

Let us now consider the defining equation

$$\sum_{j=1}^{N} L_{mj}x_j^{(k)} = \lambda^{(k)}x_m^{(k)}. \tag{76}$$

For the even symmetry case, this can immediately be rewritten as

$$\sum_{j=1}^{M^e} L_{mj}^{(e)}x_j^{(k)} = \lambda^{(k)}x_m^{(k)} \tag{77}$$

where

$$N \text{ even: } L_{mj}^{(e)} = L_{mj} + L_{m,N+1-j}$$
$$M^e = N/2$$
$$N \text{ odd: } L_{mj}^{(e)} = \begin{cases} L_{mj} + L_{m,N+1-j}; & j \neq (N+1)/2 \\ L_{mj} & ; j = (N+1)/2 \end{cases}$$
$$M^e = (N+1)/2. \tag{78}$$

Obviously the same equation is obtained if we replace m by $N+1-m$. Hence only $m \leq M^e$ need be considered and in fact we have at least partially replaced our eigenvalue problem for the $N \times N$ matrix L by one involving the $M^e \times M^e$ matrix $L^{(e)}$ of approximately half the order [$M^e = N/2$ or $(N+1)/2$],

$$L^{(e)}x^{(k)} = \lambda^{(k)}x^{(k)}. \tag{79}$$

On the other hand, for the odd symmetry cases we have

$$\sum_{j=1}^{M} L_{mj}^{(o)}x_j^{(k)} = \lambda^{(k)}x_m^{(k)} \tag{80}$$

where

$$L_{mj}^{(o)} = L_{mj} - L_{m,N+1-j} \tag{81}$$

and for

$$N \text{ even: } \quad M^o = N/2 \tag{82}$$
$$N \text{ odd: } \quad M^o = (N-1)/2,$$

(the latter because for odd symmetry of $x^{(k)}$, and N odd, the center component of $x^{(k)}$, namely $x_{(N+1)/2}^{(k)}$, must vanish). We see that in either case, *i.e.*, N even or odd, $M^e + M^o = N$, and the two sets of eigenvalues totally yield the N eigenvalues of L.

It then follows that since the determinant $|I+zL|$ is a polynomial in z with roots given by the negative reciprocal eigenvalues of L, it must be represented to within a constant factor by the product

$$|I + zL^{(e)}|\,|I + zL^{(o)}|$$

since the roots of the two polynomial factors represented in the latter are similarly related to exactly the same eigenvalues. That the constant factor is unity is evident by noting that the constant term is exactly +1 in each case. We thus have

$$P(z) = |I + zL|^{-1} = \{|I + zL^{(e)}||I + zL^{(o)}|\}^{-1}. \quad (83)$$

Appendix III

Two Special Closed-Form Solutions

The Eigenvalues for Equispaced Diversity with Exponential Correlation

The matrix, as defined in the text is in the form

$$L = \sigma^2 \begin{bmatrix} 1 & q^* & q^{*2} & & \cdots & q^{*N-1} \\ q & 1 & q^* & & \cdots & q^{*N-2} \\ q^2 & q & 1 & & \cdots & q^{*N-3} \\ & & & 1 & & \\ \cdots & \cdots & \cdots & & \ddots & \cdots \\ q^{N-1} & q^{N-2} & q^{N-3} & & \cdots & 1 \end{bmatrix} \quad (84)$$

where

$$q = \exp[-(\alpha - i\beta)\tau]. \quad (85)$$

The secular determinantal equation is

$$|L - \lambda I| = 0 \quad (86)$$

or, dividing by σ^2,

$$\begin{vmatrix} 1-\frac{\lambda}{\sigma^2} & q^* & q^{*2} & & \cdots & q^{*N-1} \\ q & 1-\frac{\lambda}{\sigma^2} & q^* & & \cdots & q^{*N-2} \\ q^2 & q & 1-\frac{\lambda}{\sigma^2} & & \cdots & q^{*N-3} \\ & & & 1-\frac{\lambda}{\sigma^2} & & \\ \cdots & \cdots & \cdots & & \ddots & \cdots \\ q^{N-1} & q^{N-2} & q^{N-3} & & \cdots & 1-\frac{\lambda}{\sigma^2} \end{vmatrix} = 0. \quad (87)$$

We will only outline the further steps, which are readily repeated. The determinant can be simplified as follows: Multiply column 2 by q and subtract from column 1; likewise multiply column 3 by q and subtract from column 2; and so forth across all columns. This places zeros in all elements for which $m > n+1$. Similarly, next multiply row 2 by q^* and subtract from row 1; multiply row 3 by q^* and subtract from row 2; and so forth. This places zeros in all elements for which $m < n+1$, and leaves an equation which can be written in the form

$$\left(\frac{\lambda}{\sigma^2}\right)^N \begin{vmatrix} y & q^* & 0 & 0 & \cdots & & & & \\ q & y & q^* & 0 & \cdots & & & & \\ 0 & q & y & & \cdots & & & & \\ 0 & 0 & & & \cdots & & & & \\ & & & & \cdots & & & & \\ & & & & \cdots & & & 0 & 0 \\ & & & & \cdots & & y & q^* & 0 \\ & & & & \cdots & 0 & q & y & q^* \\ & & & & \cdots & 0 & 0 & q & \frac{\sigma^2-\lambda}{\lambda} \end{vmatrix} = 0 \quad (88)$$

where

$$y = \sigma^2 \frac{(1 - |q|^2)}{\lambda} - (1 + |q|^2). \quad (89)$$

Let us now consider the $k \times k$ determinant D_k, which has the form of the determinant above, except that the $(\sigma^2 - \lambda)/\lambda$ in the lower corner is also replaced by y. An expansion along a column then gives the equation

$$D_k = yD_{k-1} - |q|^2 D_{k-2} \quad (90)$$

which may be regarded as a second order difference equation with constant coefficients, with boundary conditions defined by the obvious forms

$$D_1 = y$$
$$D_2 = y^2 - |q|^2. \quad (91)$$

By the usual techniques of using a trial solution of form $\exp(ik\psi)$, it is readily shown that the general solution for D_k is

$$D_k = |q|^k \frac{\sin(k+1)\phi}{\sin\phi} \quad (92)$$

where

$$\cos\phi = \frac{y}{2|q|} = \frac{1}{2|q|}\left[\sigma^2 \frac{1-|q|^2}{\lambda} - (1+|q|^2)\right]. \quad (93)$$

If we return to our determinantal equation, we see that expansion along the last column gives the form

$$\left(\frac{\lambda}{\sigma^2}\right)^N \left\{\frac{(\sigma^2-\lambda)}{\lambda} D_{N-1} - |q|^2 D_{N-2}\right\} = 0. \quad (94)$$

This becomes

$$\frac{1}{\sin\phi}\left(\frac{\lambda}{\sigma^2}|q|\right)^N \left\{\frac{(\sigma^2-\lambda)}{\lambda|q|}\sin N\phi - \sin(N-1)\phi\right\} = 0. \quad (95)$$

The possibility $\lambda = 0$ can be discarded since the matrix was assumed nonsingular at the start. Likewise the possibility $\phi = p\pi$, p an integer, is found to give a nonzero limiting value for the form in (95) and hence is not a solution. The quantity in braces in (95) therefore vanishes and may be recast in the form (using also (93) to replace λ by ϕ)

$$\tan N\phi = \frac{-\sin\phi}{\cos\phi\left(\dfrac{1+|q|^2}{1-|q|^2}\right)+\dfrac{2|q|}{1-|q|^2}}. \tag{96}$$

The N solutions of this transcendental equation, ϕ_i, give the eigenvalues through (93), which is

$$\lambda_i = \sigma^2 \frac{1-|q|^2}{1+|q|^2+2|q|\cos\phi_i}. \tag{97}$$

The eigenvalues are all numerically different, although they are clustered around the value σ^2, since it is readily observed from (97) that the values must lie in the range

$$\sigma^2\left(\frac{1-|q|}{1+|q|}\right) \le \lambda \le \sigma^2\left(\frac{1+|q|}{1-|q|}\right). \tag{98}$$

(Since the trace of a matrix is invariant under diagonalizing transformations, it is true for *all* covariance matrices for which $R_n(0)=\sigma^2$ for all n, that

$$\sum_{i=1}^{N}\lambda_i = \sum_{i=1}^{N} R_i(0) = N\sigma^2 \tag{99}$$

so that, independent of the form of the correlation function, the average value of the λ_i is exactly σ^2.)

The Eigenvalues When Only Adjacent Signals Are Correlated

As a corollary of the investigation above, we note that if we consider a new secular equation

$$(\sigma^2)^N \begin{vmatrix} 1-\dfrac{\lambda}{\sigma^2} & r^* & & & & 0 \\ r & 1-\dfrac{\lambda}{\sigma^2} & r^* & & & \\ & r & 1-\dfrac{\lambda}{\sigma^2} & \ddots & & \\ & & r & \ddots & \ddots & \\ & & & \ddots & 1-\dfrac{\lambda}{\sigma^2} & r^* \\ 0 & & & & r & 1-\dfrac{\lambda}{\sigma^2} \end{vmatrix} = 0, \tag{100}$$

then we have an $N\times N$ determinant completely analogous to the D_N form defined following (88) and (89). The equation can thus be written immediately as

$$|r|^N \frac{\sin(N+1)\theta}{\sin\theta} = 0 \tag{101}$$

where

$$\cos\theta = \frac{1-(\lambda/\sigma^2)}{2|r|}. \tag{102}$$

The solutions to (101) are simply

$$\theta = \frac{p\pi}{N+1}, \qquad p = 1, \cdots, N. \tag{103}$$

Or hence

$$1-\frac{\lambda_p}{\sigma^2} = 2|r|\cos\frac{p\pi}{N+1}$$

$$\lambda_p = \sigma^2\left[1-2|r|\cos\frac{p\pi}{N+1}\right]. \tag{104}$$

It is clear that all eigenvalues are different.

With respect to the L-matrix defined in Appendix 1, it is seen that the eigenvalues defined by the secular equation (100) are those for a covariance matrix representing the situation in which, in the ordering of the signals, each is correlated only to the adjacent signal on each side, and to no others. The mean square of each signal is σ^2 (assumed identical for all signals), and the nonvanishing b_{nm} and β_{nm} (as defined in Appendix 1) are the real and imaginary parts of $\sigma^2 r$.

However, we may readily calculate that the L-matrix involved in (100) has as its determinant

$$|r|^k \frac{\sin(k+1)\phi}{\sin\phi}$$

where $\cos\phi = 1/2|r|$. Thus L is singular (the determinant vanishes) for values of $|r|$ given by

$$\phi = \frac{p\pi}{N+1}, \qquad p = \text{integer},$$

or, hence,

$$|r| = \frac{1}{2\cos\dfrac{p\pi}{N+1}}, \qquad p = \text{integer}.$$

If L and all the submatrices of L which are correlation matrices among subsets of the N signals are to be nonsingular and nonnegative, it is clear from this last equation that the results above can be valid only for

$$|r| \le \frac{1}{2\cos\dfrac{\pi}{N+1}}.$$

For large N, this is approximately $|r|\le\frac{1}{2}$. Of course, physically, we would probably not attempt to assume this form unless we had $|r|\le 0.1$, or so.

Appendix IV

Probability of Error in Diversity Detection of Frequency-Shift Keying with Square-Law Combination

In the analysis below, we postulate a receiver which operates on the incoming signal and noise in the following manner. It is first presumed that the transmitted signal will consist of pulses of length T, with the binary symbol "1" being transmitted with a steady sinusoid of one frequency, and the symbol "0" being transmitted with a different frequency; in teletype terminology these are the *mark* and *space* frequencies. The repetition rate of the pulses is $1/T$ so that there is no dead time between successive bits. The two frequencies are chosen so that the possible transmitted waveforms are orthogonal over a pulse interval; this is accomplished either by choosing the separation to be an integral multiple of $1/T$ or by choosing the frequency separation so large that the residual correlation is negligible.

We next presume that the fading is sufficiently slow so that at the receiver terminals the amplitude and phase of the mark or space frequencies are essentially constant over a pulse length. Each receiver in a diversity scheme is identical: the signal plus noise at the input to the receiver is passed into two narrow band filters or resonators, one centered on the mark frequency, and one on the space frequency. These filters are gated synchronously with the incoming pulse train in such a manner that the filters have zero energy storage at the beginning of each pulse, and have their stored energy discharged at the end of the pulse.

Following each of the two filters is an envelope detector followed by a square-law device. The output of the square-law device is sampled just prior to discharge of the filter.

The diversity combination and decision is accomplished by comparing the sum of the mark channel samples from all receivers with the sum of the space channel samples from all receivers. The larger sum determines the more likely transmitted symbol.

It can be shown[16] that in the presence of independent fading in the several diversity branches, with fading sufficiently rapid that the amplitude of successive pulses is approximately independent, this diversity combination and decision method is the optimum statistical inference receiver. As pointed out in the body of the text, it will no longer be optimum in the presence of very slow fading or correlated signals, but the error rate for this method provides a useful upper bound on possibly more sophisticated methods.

Following the notation used in this paper we let

S_n = received power during one pulse in the nth diversity branch

S = total received power in all diversity branches

n_0 = spectral noise density in power/cps, presumed identical at each receiver

$\alpha = T/n_0$.

A fuller explanation of the derivation of the equations which follow can be found in Pierce.[16] Denoting the mark channel square-law output by x_n, the space channel output by y_n, these variates have density functions

$$p(x_n) = \exp(-x_n - \alpha S_n) I_0(2\sqrt{\alpha S_n x_n}),$$

$$p(y_n) = \exp(-y_n).$$

In these expressions we are considering *fixed* signal powers in the diversity branches; the statistical variability in x_n, y_n is due solely to thermal noise. After determining the *conditional* error probability as a function of these N fixed signal powers, we average over their joint density function describing the signal fading to determine the average error probability.

Let $X = \Sigma x_n$, $Y = \Sigma y_n$. Taking z as the variable of a Laplace transformation, since $x_1, x_2, \cdots, x_N$ are independent for fixed signal powers,

$$\mathcal{L}[p(X)] = \overline{\exp(-zX)} = \overline{\exp(-zx_1 - zx_2 \cdots - zx_N)};$$

$$\mathcal{L}[p(X)] = \overline{\exp(-zx_1)}\,\overline{\exp(-zx_2)} \cdots \overline{\exp(-zx_N)}$$
$$= \mathcal{L}[p(x_1)]\mathcal{L}[p(x_2)] \cdots \mathcal{L}[p(x_N)];$$

and similarly,

$$\mathcal{L}[p(Y)] = \mathcal{L}[p(y_1)]\mathcal{L}[p(y_2)] \cdots \mathcal{L}[p(y_N)].$$

But[19]

$$\mathcal{L}[p(x_n)] = \int_0^\infty \exp(-zx_n) \exp(-x_n - \alpha S_n) I_0(2\sqrt{\alpha S_n x_n})\, dx_n$$
$$= \frac{1}{z+1} \exp(-\alpha S_n) \exp\left(\frac{\alpha S_n}{z+1}\right),$$

so that

$$\mathcal{L}[p(X)] = (z+1)^{-N} \exp(-\alpha S) \exp\left(\frac{\alpha S}{z+1}\right)$$

$$\mathcal{L}[p(Y)] = (z+1)^{-N}$$

and

$$p(Y) = \frac{Y^{N-1}}{(N-1)!} \exp(-Y).$$

The error rate is given by

$$p(e) = \int_0^\infty p(X)\,dX \int_X^\infty p(Y)\,dY$$
$$= \int_0^\infty p(X)\left[\exp(-X) \sum_{n=0}^{N-1} X^n/n!\right] dX$$

or

[19] Erdelyi, *et al.*, "Tables of Integral Transforms," McGraw-Hill Book Co., Inc., New York, N. Y., vol. 1, p. 197, eq. 4.16.14; 1954.

$$p(e) = \sum_{n=0}^{N-1} \frac{1}{n!} \int_0^\infty dX \exp(-X) X^n p(X)$$

$$= \sum_{n=0}^{N-1} \frac{1}{n!} \left[\left(-\frac{d}{dz} \right)^n \mathcal{L}[p(X)] \right]_{z=1}$$

$$= \exp(-\alpha S) \sum_{n=0}^{N-1} \frac{1}{n!}$$

$$\left[\left(-\frac{d}{dz} \right)^n \frac{1}{(z+1)^N} \exp\left(\frac{\alpha S}{z+1} \right) \right]_{z=1}$$

$$= \exp(-\alpha S) \sum_{n=0}^{N-1} \frac{1}{n!}$$

$$\left[\left(-\frac{d}{dz} \right)^n z^{-N} \exp(\alpha S z^{-1}) \right]_{z=2}.$$

Or, letting

$$z = v^{-1}, \qquad \frac{d}{dz} = \frac{dv}{dz}\frac{d}{dv} = -v^2 \frac{d}{dv},$$

the expansion becomes

$$p(e) = \exp(-\alpha S) \sum_{n=0}^{N-1} \frac{1}{n!} \left[\left(v^2 \frac{d}{dv} \right)^n v^N \exp(\alpha S v) \right]_{v=1/2}$$

$$= \exp(-\alpha S) \sum_{n=0}^{N-1} \frac{1}{n!} \left[\exp(\alpha S v) \right.$$

$$\left. \sum_{m=0}^{n} \frac{(N-1+n)! n! (\alpha S)^m v^{N+n+m}}{(N-1+m)!(n-m)!m!} \right]_{v=1/2}$$

$$= \exp(-\tfrac{1}{2}\alpha S) \sum_{n=0}^{N-1} \sum_{m=0}^{n} \frac{(N-1+n)! 2^{-(N+n+m)} (\alpha S)^m}{(N-1+m)!(n-m)!m!}$$

$$= \exp(-\tfrac{1}{2}\alpha S) \sum_{m=0}^{N-1} \frac{1}{m!} \left(\frac{\alpha S}{2} \right)^m$$

$$\sum_{n=m}^{N-1} \frac{(N-1+n)! 2^{-N-n}}{(N-1+m)!(n-m)!}.$$

Appendix V

Comparison of Normalized Correlation Coefficients for Instantaneous Envelope and Power

Power	Envelope	Power	Envelope	Power	Envelope
0.00	0.0000	0.40	0.376	0.80	0.780
0.01	0.0092	0.41	0.386	0.81	0.791
0.02	0.0183	0.42	0.396	0.82	0.801
0.03	0.0275	0.43	0.405	0.83	0.812
0.04	0.0367	0.44	0.415	0.84	0.823
0.05	0.0459	0.45	0.425	0.85	0.833
0.06	0.0551	0.46	0.435	0.86	0.844
0.07	0.0643	0.47	0.445	0.87	0.855
0.08	0.0736	0.48	0.454	0.88	0.866
0.09	0.0828	0.49	0.464	0.89	0.877
0.10	0.0921	0.50	0.474	0.90	0.888
0.11	0.101	0.51	0.484	0.91	0.899
0.12	0.111	0.52	0.494	0.92	0.910
0.13	0.120	0.53	0.504	0.93	0.921
0.14	0.129	0.54	0.514	0.94	0.932
0.15	0.138	0.55	0.524	0.95	0.943
0.16	0.148	0.56	0.534	0.96	0.954
0.17	0.157	0.57	0.544	0.97	0.966
0.18	0.167	0.58	0.554	0.98	0.977
0.19	0.176	0.59	0.564	0.99	0.988
0.20	0.185	0.60	0.574	1.00	1.00
0.21	0.195	0.61	0.584		
0.22	0.204	0.62	0.594		
0.23	0.214	0.63	0.604		
0.24	0.223	0.64	0.614		
0.25	0.233	0.65	0.624		
0.26	0.242	0.66	0.635		
0.27	0.252	0.67	0.645		
0.28	0.261	0.68	0.655		
0.29	0.271	0.69	0.665		
0.30	0.280	0.70	0.676		
0.31	0.290	0.71	0.686		
0.32	0.299	0.72	0.696		
0.33	0.309	0.73	0.707		
0.34	0.318	0.74	0.717		
0.35	0.328	0.75	0.727		
0.36	0.338	0.76	0.738		
0.37	0.347	0.77	0.748		
0.38	0.357	0.78	0.759		
0.39	0.367	0.79	0.769		

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Correspondence

National Standards of Time and Frequency in the United States*

Time

Time is one of the independent quantities chosen as a basis for the measuring system of science. To establish a time scale, that is, to specify a time coordinate, we must first establish an origin and a constant unit of time and lay off the unit at least as far as all points of interest, in exact analogy to the specification of position. The orbital motion of the earth about the sun provides a time scale, called Ephemeris Time (ET) which is suitable in these respects, and also constitutes a standard from which the unit may always be obtained. In October, 1956, the International Committee on Weights and Measures, with representation from the United States, effectively adopted the second of ET as the fundamental unit of time by resolving that "the second is the fraction 1/31,556,925.9747 of the tropical year for January 0, 1900 at 12 hours Ephemeris Time."[1] The tropical year for any given epoch (or moment) is the interval, taken symmetrically about that epoch, necessary for the mean longitude of the sun to increase by 360°, as measured along the ecliptic from the vernal equinox. This definition of the second has been one of many attempts to define a unit of time 1) which remains constant with epoch, 2) which is physically realizable to high precision, and 3) which is permanently available for observation.

Ephemeris Time is obtained in practice by observations on the position of the moon and reference to tables giving this position as a function of ET. As the fundamental unit, the second of ET replaced the second of Universal Time (UT), a time scale based on the rotation of the earth. The delay in the determination of UT from the astronomical observations is of the order of a month; the delay in the determination of ET to any useful degree of accuracy is of the order of several years.

The maintaining of a national standard of time consists of the process of determining the time scale in the United States from the international standard and making this information available to the user. Often, however, as with position, it is only time differences or time intervals which are of interest. This need for convenient interval determination has led to the common use of other time scales with easier realizability than ET.

Measurements often have been, and still are, expressed in terms of time scales other than ET. These other scales are all related or relatable to ET and to each other through conversion factors with various degrees of precision as to size of unit at least, and in some cases as to origin also. The situation is analogous to using a measuring rod of different length than the standard rod but of known calibration. For example, UT0 is the astronomically observed Universal Time or mean solar time, uncorrected for polar variation and annual fluctuation in the earth's speed of rotation; UT1 is UT0 corrected for polar variation; UT2 is UT1 corrected for annual variation. None of the UT seconds is constant. They are variously determined by several nations. Various other nominal "seconds" are defined by the duration of a specified number of periods of various oscillators or resonators, such as quartz or atomic devices. These all have the outstanding characteristic that they are more readily observable than ET. The atomic time unit, furthermore, is assumed for the present to be as constant as the unit of ET.

The United States Frequency Standard

A national standard of frequency called the United States Frequency Standard (USFS) is maintained at the Boulder Laboratories, National Bureau of Standards, for the purpose of making immediately and continuously available, through the Standard Frequency Broadcasts discussed below, a provisional time scale with which to make measurements. This provisional scale is sufficiently accurate for all civil and most scientific uses. Nevertheless, for the highest possible accuracy, this scale must be continually corrected to or calibrated against the national time scale as finally determined, and it must be relatable to other frequency and time scales in common use.

The USFS in general consists of the weighted value of the outputs (reduced of course to a common basis) of several actual oscillators and resonators maintained at the Boulder Laboratories, these devices being among the best available at any current state of the frequency control art. In particular, the USFS is presently stable to 2 parts in 10^{10} or better over intervals from about 1 to 10^3 minutes; over longer intervals, its value has been maintained as constant as possible prior to October 9, 1957 with respect to the UT2 second as determined by the U. S. Naval Observatory, and since October 9, 1957 with respect to atomic frequency standards. These atomic standards have been compared with other atomic standards via a network comparison. Such atomic standards have been shown to be in agreement and to remain constant with respect to each other to 5 parts in 10^{10} or better.[2,3]

Standard Frequency Broadcasts

The United States Frequency Standard is distributed to interested users by means of standard broadcasts of Radio Stations WWV and WWVH. The frequencies of these stations are kept in agreement with respect to each other and have been maintained as constant as possible with respect to the United States Frequency Standard since December 1, 1957. The nominal broadcast frequencies should, for the purpose of highly accurate scientific measurements, for establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected either to the United States Frequency Standard, as indicated in the monthly tables published in the IRE,[4] or to a particular time scale as determined by the Naval Observatory, with adequate limits assigned for propagation errors.

Time Signals

Time signals, sufficiently accurate for all civil and most scientific uses, are also carried by the standard broadcasts. The WWV and WWVH time signals are also kept in agreement with each other. They are locked to the nominal frequency of the transmissions, and consequently may depart continuously from other time scales such as UT2. Corrections expressed as the times of reception of the WWV time signals on the UT2 scale are determined and distributed by the U. S. Naval Observatory.[5] Recently, the Observatory has established a particular atomic time scale, A.1, with unit determined by 9,192,631,770 periods of Cs at zero field and origin coinciding with UT2 on January 1, 1958, and has published times of reception of WWV on this scale also.[6] Agreement with time on the UT2 scale, or simply agreement with UT2, within ±30 msec at all times, has been maintained by making step adjustments in time of precisely plus or minus twenty milliseconds on Wednesdays at 1900 UT when necessary. Beginning January 1, 1960, the broadcast frequencies will be offset from the United States frequency standard by a different amount than heretofore in order to establish a unit in substantial agreement with the current value of the unit of UT2. Thus the time signals, locked to the broadcast frequency, will require less frequent adjustment than in the past.

Corrections of the USFS and the Standard Frequency Broadcasts

The method of final correction of the

* Received by the IRE, October 15, 1959.

[1] Comite International des Poids et Mesures, "Proces-Verbaux de Seances," 1956 Session, Ser. 12, vol. 25, Gauthier-Villars, Paris; 1957.

[2] Essen, Parry, Holloway, Mainburger, Reder, and Winkler, "Comparison of cesium frequency standards of different construction," *Nature*, vol. 182, pp. 41–42; July, 1958.

[3] Mockler, Beehler, and Barnes, "An evaluation of a cesium beam frequency standard," Symposium Record, Office of Naval Research Symposium on Quantum Electronics, Bloomingburg, N. Y.; September 14–16, 1959.

[4] W. D. George, "WWV standard frequency transmissions," PROC. IRE, vol. 46, pp. 910–911; May, 1958, and subsequent months.

[5] "U. S. Naval Observatory Time Signals, I. Preliminary Times of Reception, UT2," Bulletin B, U. S. Naval Observatory, Washington, D. C. (unpublished).

"U. S. Naval Observatory Time Signals, Final Times of Reception, UT2," Bulletin A, U. S. Naval Observatory, Washington, D. C. (unpublished).

[6] "Time Service Notice No. 6," U. S. Naval Observatory, Washington, D. C., January 1, 1959 (unpublished).

USFS and the Standard Frequency Broadcasts to the national time scale has changed over the years as improvements in the measurement of time and frequency have been made.

Fig. 1 shows schematically the relation between the frequencies (*i.e.*, cycles per common unit of time) of hypothetical oscillators which oscillate such as to mark off the unit of the time scales indicated after exactly the same number of cycles in each case. The frequencies are shown relative to the value of the hypothetical oscillator marking off the defined unit of ET. The figure is equivalently interpreted as the reciprocal of the size of the units of the indicated time scale relative to the unit of ET.

The Standard Frequency Broadcasts and Time Signals have been and are related to the UT and A.1 time scales as explained above. From these observations, frequency corrections of the Standard Frequency Broadcasts to these scales are easily made; and in addition, conversion from the UT scale at a given epoch to the ET scale is available after reduction of astronomical observations.[7–9]

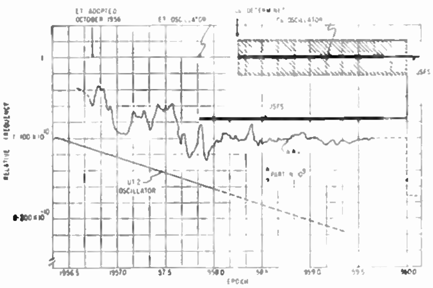

Fig. 1—Relative frequencies of oscillators keeping the time scales indicated; or equivalently, reciprocals of the lengths of unit intervals on the indicated time scales. Width of Cs and USFS lines indicates precision; cross hatching shows assigned accuracy of Cs frequency. The UT2 scale is derived from smoothed data.[5] The UT2 unit is considerably less precise than the other units. The WWV scale is taken from a U. S. Naval Observatory Notice.[6] Values of USFS and WWV transmissions will be altered as indicated on January 1, 1960.

Prior to October 9, 1957, the final value of the USFS was assigned retrospectively on the UT scale from the determination of the frequency of the Standard Frequency Broadcasts on this scale and the known relation between the Standard Frequency Broadcasts and the USFS. These values have not been published, but are a matter of scientific record at the National Bureau of Standards.

With improvement in stability and constancy realized in the maintenance of the USFS by the availability of atomic frequency standards, monthly publications of corrections of the Standard Frequency Broadcasts to the USFS were made from October 9, 1957, to realize the advantage of rapid correction to a more constant frequency than that based on the unit of UT2.

It is thus also necessary to relate the USFS and Standard Frequency Broadcasts to frequencies provided by atomic frequency standards such as cesium. Since the relation of an atomic time unit defined by a given number of periods, N, of an atomic oscillation to the unit of ET has not yet been adopted internationally, we are required to state explicitly an assumed value of N when comparing frequencies or times to an atomic scale. In comparing the USFS to an atomic time scale, the following equality was adopted as of October 9, 1957, and holds to a precision of 1 part in 10^{10}; 100,000,000 . . . periods of the USFS equals N_1 periods of the zero-field $(4, 0) \leftrightarrow (3, 0)$ Cs transition, where N_1 equals 9,192,631,838.

If N_0 is the number of Cs oscillations per unit of ET, then the ratio N_0/N_1 may be used to convert the USFS to the ET scale as soon as N_0 is determined. Markowitz, *et al.*[8] have measured N_0 to be 9,192,631,770 ± 20 as the number of Cs oscillations per unit of ET, providing the best and only value expected for several years. Effective January 1, 1960, the value of the USFS will be corrected so that N_1 equals 9,192,631,770, and the published corrections to the Standard Frequency Broadcasts will thus be given with respect to the unit of ET as realized by atomic standards to a precision of the intercomparison of atomic standards (a few parts in 10^{10}) and to an accuracy determined by the work of Markowitz, *et al.* (± 22 parts in 10^{10}).

The final correction of the Standard Frequency Broadcasts, and hence the USFS, to the national time scale, will be effected by the Naval Observatory by the observation of the times of reception as described.

In summary, to express a given time interval or frequency in terms of the international standard, it must be referred to a standard broadcast, or some atomic standard with proper regard for the precision of this comparison, and the latter standard must be referred to the ephemeris second as indicated herein with proper regard for the uncertainties involved.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

[7] Essen, Parry, Markowitz, and Hall, "Variation in the speed of rotation of the earth since June, 1955," *Nature*, vol. 181, p. 1054; April, 1958.
[8] Markowitz, Hall, Essen, and Parry, "Frequency of cesium in terms of Ephemeris Time," *Phys. Rev. Letters*, vol. 1, pp. 105–106; August, 1958.
[9] Brouwer, "A study of the changes in the rate of rotation of the earth," *Astron. J.*, vol. 57, pp. 125–146; September, 1952.

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 U.S.A. Frequency Standard was 1.4 parts in 10^9 high with respect to the frequency derived from the UT2 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 U.S.A. Frequency Standard, as indicated in the table; values are given as parts in 10^{10}. This correction is *not* with respect to the current value of frequency based on UT2. A minus sign indicates that the broadcast frequency was low.

The WWV and WWVH time signals are synchronized; however, they may gradually depart from UT2 (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.

WWV and WWVH time signals are maintained in close agreement with UT2 by making step adjustments in time of precisely plus or minus twenty milliseconds on Wednesdays at 1900 UT when necessary; no time change or adjustment was made at WWV and WWVH during October. Retarding time adjustments were made on November 4 and 18, 1959.

* Received by the IRE, November 27, 1959.

WWV FREQUENCY†

1959		#1	#2	#3
October	1	−23	−29	−26
	2‡	−22	−24	−20
	3	−24	−23	
	4‡	−24	−19	
	5	−34	−29	−24
	6‡	−32	−19	−17
	7	−38	−31	−27
	8	−38	−32	−28
	9	−37	−35	−31
	10	−36	−39	
	11	−36	−36	
	12	−35	−34	−29
	13	−34	−35	−30
	14	−33	−35	−31
	15	−33	−35	−30
	16	−32	−35	−30
	17	−32	−35	
	18	−31	−35	
	19	−31	−34	−30
	20	−30	−35	−30
	21	−30	−35	−29
	22	−30	−35	−29
	23	−30	−34	−30
	24	−30	−35	
	25	−30	−34	
	26	−30	−34	−29
	27	−30	−34	−30
	28	−30	−35	−31
	29	−31	−35	−29
	30	−31	−35	−29
	31	−31	−35	
November	1		−34	
	2		−34	−29
	3		−38	−33
	4		−33	−28
	5		−33	−28
	6		−33	−28
	7		−33	
	8		−32	
	9		−33	−27
	10		−33	−28
	11		−33	
	12		−32	−28
	13		−32	−28

† WWVH frequency is synchronized with that of WWV.

Column #1 Vs NBS‡ atomic standards, Boulder, Colo., 30-day moving average seconds pulses at 15 mc.
Column #2 Vs atomichron at WWV, measuring time one hour at 2.5 mc.
Column #3 Vs atomichron at the U. S. Naval Research Laboratory, Washington, D. C., measuring time 56 minutes at 2.5 mc.

‡ Method of averaging is such that an adjustment of frequency of the control oscillator appears on the day it is made. The following frequency adjustments were made:

October 2—minus 3×10^{-10};
October 4—minus 9×10^{-10};
October 6—minus 7×10^{-10}.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

The Optimum Noise Performance of Tunnel-Diode Amplifiers*

Introduction

The noise in a tunnel diode[1] has been found to be essentially of the shot-effect type. The magnitude of the excess noise factor, which is defined as the noise factor F minus unity, is

$$F - 1 = \left(\frac{G_1}{G_g} + \frac{G_L}{G_g} + \frac{G_e}{G_g}\right)\frac{T}{T_0}, \quad (1)$$

where G_g is the source conductance at the reference temperature T_0, G_n is the loss conductance which contributes thermal noise, G_L is the load conductance, and G_e is the equivalent shot-noise conductance of the diode at the ambient temperature T.

If $R(=1/G)$ is the negative resistance of the diode corresponding to the operating diode current I, and g_p is the power gain, it follows that

$$(A)(F-1)$$
$$= \left(G_1|R| + G_L|R| + \frac{eI}{2kT}|R|\right) \quad (2)$$

where

$$A = (|R|G_g)\frac{T_0}{T} = \left(1 + \frac{G_L}{G_g} + \frac{G_1}{G_g}\right.$$
$$\left. - \frac{2}{\sqrt{g_p}}\sqrt{\frac{G_L}{G_g}}\right)^{-1}\left(\frac{T_0}{T}\right) \quad (3)$$

$$g_p = \frac{4G_gG_L}{(G_1 + G_g + G_L - |G|)^2}. \quad (4)$$

Here $A(F-1)$, the product of the excess noise and a constant depending on gain, is a measure of the noise-and-gain behavior of the tunnel diode amplifier. A small value of $A(F-1)$ means small noise factor and/or large power gain g_p.

The purpose of this note is to show analytically that with a given I-V characteristic of a tunnel diode, a minimum value of $A(F-1)$ exists at a unique point on the characteristic. A further object is to suggest an ideal I-V characteristic which can be a guide for the design of tunnel-diode amplifiers with low values of $A(F-1)$.

Analysis

The I-V characteristic of a typical tunnel diode is shown in Fig. 1(a). The negative-slope portion of the curve lies between $V=V_1$ and $V=V_2$, where the slopes are zero. An inflection point (V_0, I_0) exists between these two points.

The negative-slope portion of the curve can be represented by the function:

$$I - I_0 = f(V - V_0). \quad (5)$$

Then, its derivative with respect to V is

$$I' = \frac{dI}{dV} = f'(V - V_0) = \frac{1}{R}. \quad (6)$$

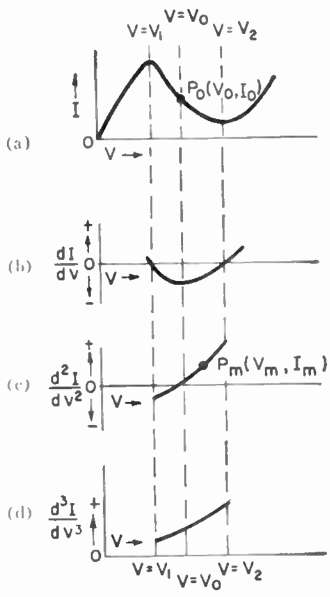

Fig. 1—Characteristic curves of tunnel diodes.

where R, the reciprocal of the slope, is the negative resistance of the diode. Consider first the case in which the thermal noise term $G_1|R|$ and the load noise term $G_L|R|$ in (2) are ignored. The IR product

$$(=ACF-1)2KT/e)$$

is

$$IR = \frac{f(V - V_0) + I_0}{f'(V - V_0)}. \quad (7)$$

IR is a maximum or a minimum at the point (V_m, I_m) where

$$\frac{d(IR)}{d(V - V_0)} = 0; \quad (8)$$

that is

$$\frac{[f'(V_m - V_0)]^2 - \{f(V_m - V_0) + I_0\}f''(V - V_0)}{[f'(V_m - V_0)]^2} = 0, \quad (9)$$

since

$$V_1 < V_m < V_2 \quad (10)$$

by Fig. 1(a),

$$f'(V_m - V_0) \neq 0. \quad (11)$$

Eq. (9) thus becomes

$$[f'(V_m - V_0)]^2$$
$$= \{f(V_m - V_0) + I_0\}f''(V_m - V_0). \quad (12)$$

It follows from (7) that

$$(IR)\ \text{max or min} = \frac{f'(V_m - V_0)}{f''(V_m - V_0)}. \quad (13)$$

That the IR product (13) is a minimum rather than a maximum will now be shown.

By making use of (12), the second derivative of (7) at the point (V_m, I_m) is

$$\left(\frac{d^2(IR)}{dV^2}\right)_{\substack{V=V_m \\ I=I_m}} = \frac{f'(V_m - V_0)f''(V_m - V_0) - \{f(V_m - V_0) + I_0\}f'''(V_m - V_0)}{-[f'(V_m - V_0)]^2}. \quad (14)$$

Since $f(_m - V_0)+I_0$ is always positive, $f''(V_m - V_0)$ is positive according to (12). Thus, by the graph of $f''(V-V_0)$ [Fig. 1(c)], V_m must be greater than V_0; that is, the maximum or minimum IR product occurs *beyond* the inflection point (V_0, I_0).

Also

$$f'(V_m - V_0) < 0 \ \text{[by Fig. 1(b)]} \quad (15)$$

$$f'''(V_m - V_0) > 0 \ \text{[by Fig. 1(d)]}. \quad (16)$$

In fact, for the I-V characteristic of the tunnel diode, it can be shown graphically that the numerator of (14) is always negative and hence the second derivative of IR is always positive. Therefore, the absolute value of IR given by (13) is a ***minimum***, and a minimum value of $A(F-1)$ does exist.

If the thermal-noise term $G_1|R|$ and the load-noise term $G_L|R|$ are included in (2) then the point of optimum noise factor (V_m, I_m) shifts towards the inflection point (V_0, I_0). The actual minimum value of $A(F-1)$ at this new point is the same as that obtained by the present analysis provided I_0 is replaced by $I_0 + (G_1 + G_L)(2KT/e)$.

Example

Assume

$$I = I_0 - G_0V_0 \sin\frac{(V - V_0)}{V_0}, \quad (17)$$

where $-G_0$ is the slope at the inflection point. This is a crude but simple approximation for the characteristics of a tunnel diode. It follows that

$$I' = -G_0 \cos\frac{(V - V_0)}{V_0} \quad (18)$$

$$I'' = \frac{G_0}{V_0}\sin\frac{(V - V_0)}{V_0}. \quad (19)$$

Using (12), the voltage which yields a minimum IR is

$$V_m = \left\{\left[\sin^{-1}\left(\frac{V_0}{I_0}G_0\right)\right] + 1\right\}V_0. \quad (20)$$

From (13)

$$(IR)_{\min} = \frac{I_0}{G_0}\sqrt{1 - \left(\frac{V_0}{I_0}G_0\right)^2} \quad (21)$$

where $(V_0G_0/I_0)^2 < 1$ according to (10). Accordingly, IR can be made small by making $(V_0)^2(G_0)^2/I_0^2$ approach unity.

According to (13), the minimum IR product is smallest for small values of $f'(V_m - V_0)$ and large values of $f''(V_m - V_0)$. This, of course, requires a small value of the minimizing current I_m by (12). An I-V characteristic (Fig. 2), which has a steep drop (portion a–b) followed by a slow decay (portion b–c) toward a very small current minimum, is, for example, an ideal curve to fulfill these requirements for a low minimum IR product. The slow decay may offer an additional advantage of a wide

* Received by the IRE, October 13, 1959. This research was sponsored in part by the Electronic Res. Directorate, AF Cambridge Res. Center, under Contract AF-19-(604)-4980.

[1] H. S. Sommers, "Tunnel diodes as high-frequency devices," Proc. IRE, vol. 47, pp. 1201–1206; July, 1959. Also K. K. N. Chang, "Low-noise tunnel-diode amplifier," Proc. IRE, vol. 47, pp. 1268–1269, July, 1959.

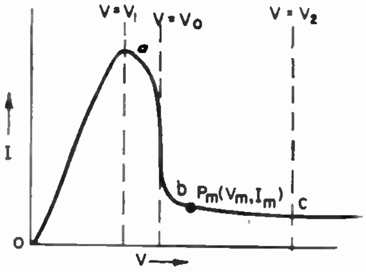

Fig. 2—Ideal I-V characteristic.

choice of stable operating points of low slopes.

In practice, however, low operating slopes result in high operating negative resistances which can be realized only by small capacitances at high operating frequencies. This is probably the restriction for obtaining a low IR product for low-noise amplification.

It is a pleasure to thank Dr. S. Bloom for many enlightening discussions.

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Maser Operation with Signal Frequency Higher than Pump Frequency*

An X-band solid-state maser utilizing the four Zeeman levels in ruby has been successfully operated with a signal frequency (10,590 mc) higher than the pump frequency (9595 mc). The method of operation depends upon the mechanism of harmonic spin-coupling[1,2] which is closely allied to the cross-relaxation mechanism treated by Bloembergen, *et al.*[3]

Continuous-wave operation was obtained at a magnetic field of 1675 oersteds oriented at 90 degrees to the ruby C axis. The energy level diagram is shown in Fig. 1, where the energy levels are numbered in order of increasing energy. For essentially the same operating conditions, anomalous excitation of a maser with signal frequency $f_{12} \approx 1000$ mc using a pump at $f_{23} \approx 9650$ mc had previously been reported by Higa[1] and Geusic.[2] The explanation for the anomalous operation is given in terms of harmonic spin-coupling between the f_{23} and f_{14} transitions, since $f_{14} = 2f_{23}$ for this operating point.

* Received by the IRE, October 28, 1959. This work is part of a doctoral dissertation at Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.

[1] W. H. Higa, "Anomalous Excitation of a Solid-State Maser," presented at the Conf. on Electron Tube Res., Mexico City, Mexico, June 24, 1959; to be published in *Proc. Conf. Quantum Electronics—Resonance Phenomena*, Columbia University Press, New York, N. Y.

[2] J. E. Geusic, "Maser Action in Ruby in the Frequency Range 800 to 9999 Mc, and Harmonic Spin Coupling in Ruby," presented at the Conf. on Electron Tube Res., Mexico City, Mexico; June 24, 1959.

[3] N. Bloembergen, S. Shapiro, P. Pershan, and J. O. Artman, "Cross-relaxation in spin systems," *Phys. Rev.*, vol. 144, pp. 445–459; April 15, 1959.

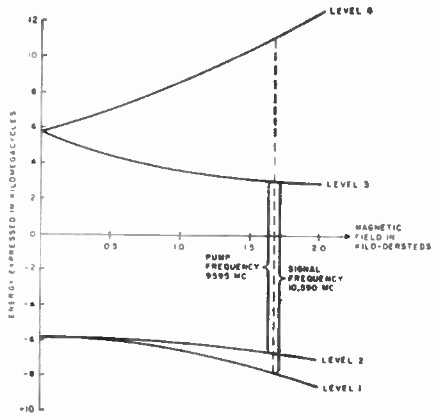

Fig. 1—Energy levels used in obtaining maser operation at a signal frequency higher than the pump frequency.

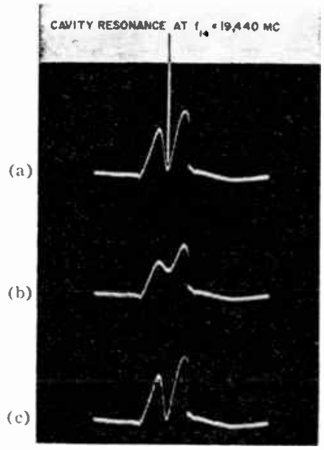

Fig. 2—Oscilloscope display of power reflected from maser cavity as a function of frequency. (a) Magnetic field off, pump power off; (b) magnetic field on, pump power off; (c) magnetic field on, pump power on at 9650 mc, showing saturation.

Solution of the rate equations shows that f_{34} acts as the idler in this type of operation.

We have established experimentally that the f_{14} transition is saturated due to the pump power applied at frequency f_{23} (Fig. 2). Furthermore, the frequency ratio need not be exactly integral. It was possible, using a constant pump frequency of $f_{23} = 9650$ mc. to saturate the f_{14} transition over a 600-mc frequency range (centered near 19,300 mc) with optimized magnetic field.

The negative temperature at f_{12} and the saturated f_{23} transition combine to obtain a negative temperature at a frequency f_{13} that is 1000 mc higher than the applied pump frequency f_{23}. This was verified experimentally, and strong CW maser amplification and oscillations were obtained.

The experimental setup consisted of a doubly-resonant tunable cavity, operating in the TE_{10} waveguide mode. The ruby crystal had a 0.05 per cent residual chromium content. The pump power was 8 milliwatts at a liquid helium bath temperature of 4.2°K. To be sure that the saturation of the f_{14} transition was primarily due to harmonic spin-coupling, the second-harmonic power output of the pump power source in the 18- to 26-kmc range was measured and was found to be less than 1 microwatt. Further measurements, such as gain-bandwidth product and tuning range, are in progress.

Harmonic spin coupling and cross-relaxation effects play an important role in masers and have relaxation times normally short compared with spin-lattice relaxation times.[4,5] Higher-order cross-relaxation effects[6] are probably responsible for transient maser action reported in hyperfine structure of $Cu(NH_4)_2(SO_4)_2 \cdot 6H_2O$ with a signal frequency 200 mc higher than the pump frequency.[7] In addition to the presently-described mode of operation, other interesting schemes for maser operation can be obtained,[8] in which harmonic spin-coupling effects are of importance. By taking advantage of this mechanism, it can be expected that CW millimeter-wave maser amplifiers using low-frequency pump sources (and frequency ratios higher than 2) will become available in the foreseeable future.

Stimulating discussions with Prof. M. Birnbaum, Polytechnic Institute of Brooklyn, are gratefully acknowledged. The author also wishes to thank Prof. A. E. Siegman for discussions and an advance copy of the paper by Chang.[5]

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[4] S. Shapiro and N. Bloembergen, "Relaxation Effects in a Maser Material, $K_3(CoCr)(CN)_6$," Cruft Lab., Harvard University, Cambridge, Mass., Tech. Rept. No. 306; June, 1959.

[5] W. S. C. Chang, "Spin-Lattice Relaxation via Harmonic Coupling," *Proc. Conf. Quantum Electronics—Resonance Phenomena*, Bloomingburg, N. Y., September 14, 1959; to be published by Columbia University Press, New York, N. Y.

[6] P. Sorokin, G. Lasher, and I. Gelles, "Cross-Relaxation and Maser Pumping by a Quadruple Spin Flip Mechanism," *Proc. Conf. Quantum Electronics—Resonance Phenomena*, Bloomingburg, N. Y., September 14, 1959; to be published by Columbia University Press, New York, N. Y.

[7] F. R. Nash and E. Rosenvasser, "Cross-Relaxation and Maser Action in $Cu(NH_4)_2(SO_4)_2 \cdot 6H_2O$," *Proc. Conf. Quantum Electronics—Resonance Phenomena*, Bloomingburg, N. Y., September 14, 1959; to be published by Columbia University Press, New York, N. Y.

[8] F. R. Arams, "Low-field X-band ruby maser," Proc. IRE, vol. 46, p. 1373; August, 1959.

Experimental Verification of Parametric Amplifier Excess Noise Using Transformer Coupling*

A method for obtaining the excess noise of a parametric amplifier by using transformer coupling between it and the second stage has been experimentally verified. By proper adjustment of the transformer turns ratio and the negative conductance of the amplifier, it is possible to minimize the noise contribution of the second stage. This is accomplished at the expense of the increased bandwidth and gain stability obtainable with a circulator. It has proved to be a very useful tool for measurement of the effective noise temperature of parametric amplifiers operating in regions where circulators are not available.

* Received by the IRE, September 10, 1959.

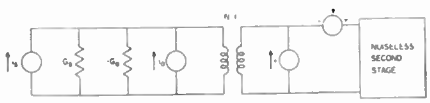

Fig. 1—Equivalent circuit of the transformer-coupled parametric amplifier (at resonance).

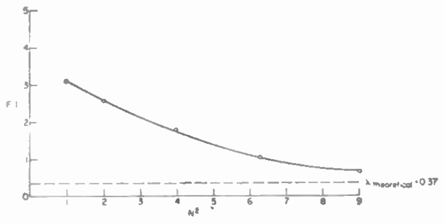

Fig. 2—Excess noise as a function of transformer turns ratio squared (excluding transformer noise contribution).

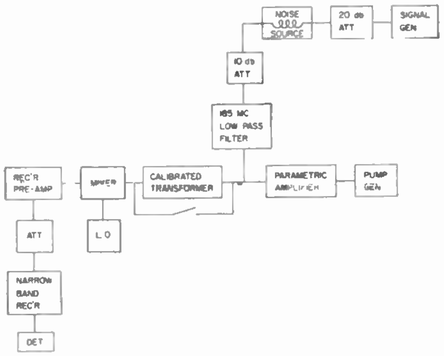

Fig. 3—Block diagram for noise measurements.

The expression for the excess noise of a transformer-coupled, non-degenerate parametric amplifier is given by[1]

$$F - 1 = \lambda\left(-\frac{G_0}{G_s}\right) + \frac{1}{n^2}\frac{G_l}{G_s} + n^2\frac{R_e}{G_s}(G_s + G_0)^2 + \frac{G_s + G_0}{G_s}\gamma^2 R_e G_l, \qquad (1)$$

where λ is the minimum excess noise of the parametric amplifier in a circulator connection. [For the nondegenerate amplifier, this excess noise is given by[2]

$$\lambda = \text{limit}\,(F - 1)_{G_l \to \infty} = \frac{T}{T_0}\left(\frac{G_l}{G_S} + \frac{f_s}{f_i}\right).]$$

G_l and G_e are the equivalent noise conductance and resistance of the second stage, γ is the correlation coefficient $(-1 \leq \gamma \leq 1)$, and n is the transformer turns ratio. G_l is the unloaded conductance of the signal tank circuit, T is the operating temperature, $T_0 = 290°$ K, G_s is the source conductance, and G_0 is the negative conductance of the amplifier.

The equivalent circuit for this connection is shown in Fig. 1 with the second stage noise sources included. The over-all excess noise of the cascade can be made to approach the value of λ by setting $G_0 = -G_s$ and making the transformer turns ratio, n, large. The transducer gain of the parametric amplifier is given by

$$G_t = \frac{4n^2 G_s Gl}{(G_0 + Gl + G_s)^2}, \qquad (2)$$

where G_l is the input conductance of the second stage. For $G_l = G_s$ and $n = 1$, the negative conductance of the amplifier, G_0, equals G_s when the transducer gain is adjusted for 6 db. If the transformer turns ratio is now increased while maintaining the pump power constant, the excess noise given by (1) becomes

$$F - 1 = \lambda + \frac{1}{n^2}\frac{G_l}{G_s}. \qquad (3)$$

The excess noise decreases with increasing turns ratio and asymptotically approaches the limiting value achievable with a circulator connection.

This relationship was checked on a coaxial cavity parametric amplifier operating at a signal frequency, f_s, of 140 mc and pump frequency, f_p, of 615 mc. The second stage noise figure was measured as a function of source impedance using standard noise measuring techniques and G_l/G_s was found to be 2.63. The plot of excess noise vs n^2 is shown in Fig. 2.

The limiting value for λ could not be achieved because of the gain instability caused by the change in VSWR of the noise source during the measurement. A block diagram of the noise-measurement setup is shown in Fig. 3. The signal generator was used for monitoring the gain variation caused by the gas discharge noise lamp.

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[1] H. A. Haus, "Theory of Noise Measurements on Solid-State Amplifier of Transmission Cavity Type," Raytheon Co., Waltham, Mass., Tech. Memo. T-138, T-138A.

[2] H. Heffner and G. Wade, "Gain, bandwidth and noise characteristics of the variable-parameter amplifier," *J. Appl. Phys.*, vol. 29, pp. 1321–1331; September, 1958.

The Backfire Antenna, a New Type of Directional Line Source*

Endfire antennas are usually parasitically-excited slow wave structures; for example, the Yagi antenna [see Fig. 1(a)]. Excited at one end of the array by the feed F (in combination with reflector R), a surface wave travels along the array of directors D with a phase velocity smaller than that of light and with its energy concentrated in a "wave channel" surrounding the array axis. At the termination of the structure, the surface wave radiates from a virtual aperture VA. The total radiation pattern is the sum of this surface wave radiation from VA and some direct radiation from F.

If a plane reflector M is mounted at the termination of the slow wave structure, Fig. 1(b), the surface wave incident from F will be reflected at M and traverse the structure a second time. The energy is now radiated from a virtual aperture VA located at the feed end. We term this structure a "backfire" antenna because it radiates in the reverse direction from ordinary endfire.

* Received by the IRE, August 28, 1959.

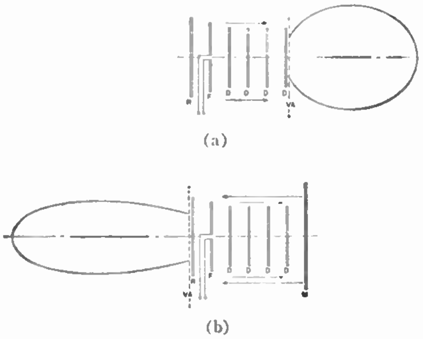

Fig. 1—(a) Principle of Yagi type endfire antenna; (b) principle of "backfire" antenna.

The maximum gain obtainable from endfire antennas is directly proportional to their lengths, provided the phase velocity is progressively adjusted to its optimum value as the antenna is lengthened. The optimum value increases with the increasing length of the antenna, approaching the velocity of light for very long structures. The gain of an ordinary endfire antenna under this optimum condition[1–3] is about 10 $l/2$ above isotropic, where l is the length of the structure. Because the physical length is traversed twice in the case of the backfire antenna, this antenna acts like an ordinary endfire antenna of double length, and the phase velocity must be adjusted accordingly. We then expect the backfire antenna to have a gain 3 db higher than an ordinary endfire antenna of equal length; and conversely, to achieve a prescribed gain, the backfire antenna need be only half as long as in the normal endfire case.

The surface-wave reflector M is plane because the surface-wave phase fronts are plane, and it should be large enough to intercept the bulk of the energy traveling in the wave channel along the array axis. The channel cross section increases with increasing phase velocity and thus with increasing length of structure. The surface wave reflector must, therefore, also increase in size as the antenna length and gain go up. It can be shown, in fact, that the diameter of the surface wave reflector is approximately the same as that of a parabolic dish whose gain equals that of the backfire antenna.

Therefore, the principal applications of the backfire antenna, as indeed of any kind of endfire line source, lie in situations in which dishes are not competitive, especially when the gain is less than 20 db. In competitive situations, the structural advantage of a plane reflector over a parabolic dish must be weighed against the

[1] W. W. Hansen and J. R. Woodyard, "A new principle in directional antenna design," PROC. IRE, vol. 26, pp. 333–345; March, 1938.

[2] D. G. Reid, "The gain of an idealized Yagi array," *J. IEE*, vol. 93, pp. 564–566; August, 1946.

[3] H. W. Ehrenspeck and H. Poehler, "A New Method for Obtaining Maximum Gain from Yagi Antennas," ASTIA Document No. AD160761; August, 1956.

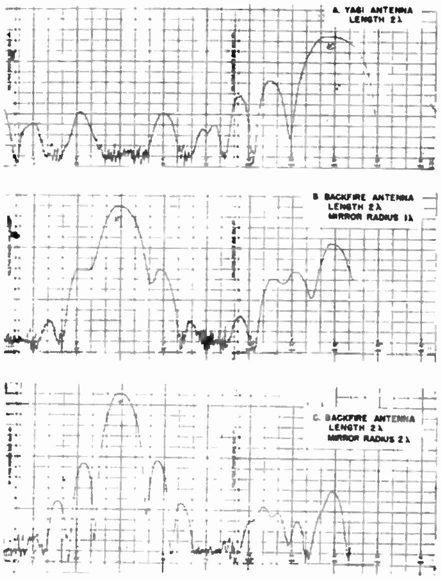

Fig. 2—Far field patterns of Yagi and "backfire" antenna.

added complication of the row of directors whose length increases with increased gain.

A series of measurements has verified the backfire principle. Far field patterns of three antennas, each two wavelengths long, are presented in Fig. 2. The Yagi antenna (A) has a beamwidth of 41° in azimuth, and the first sidelobe is 9 db below maximum. The backfire antenna (B) uses a plane reflector (called "mirror" in Fig. 2) two wavelengths in diameter which gives a gain increase of 3 db over A. The beamwidth in B is only 25°, and the first sidelobe (a shoulder) is 12 db below maximum. We note a relatively high backlobe that is only 7 db below maximum, which shows that evidently too much energy still bypasses the surface wave reflector.

The backfire antenna with doubled reflector diameter (C) has a beamwidth of about 20°, the first sidelobes are at 13 db, and the backlobe is now down to 18 db. The pattern bandwidth of C was found to be in excess of 20 per cent.

Since the beamwidth in elevation differs only slightly from that in azimuth, the 20° beamwidth of antenna (C) implies a gain that is higher than the expected 3 db above ordinary endfire. It has been found, in fact, that the gain can be considerably increased beyond 3 db by optimizing the efficiency of excitation of the surface wave in the feed system, and by increasing the diameter of the surface wave reflector to an optimum value.

To give an example, a backfire antenna 2.5 wavelengths long with a reflector diameter of 4 wavelengths produced a half-power beamwidth of 16° and a gain of 20.8 db above isotropic (about 6 db above a 2.5-wavelengths long ordinary endfire antenna).

The principle of the backfire antenna is applicable not only to Yagi-type endfire antennas, but to other slow wave stuctures as well.

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Circulators at 70 and 140 KMC*

The symmetric Y junction seems to be a particularly suitable geometry for circulators in the millimeter wavelength region. Carlin[1] has pointed out that any lossless, nonreciprocal, three-port, microwave junction which is matched, is a perfect circulator. The scattering matrix for a lossless, nonreciprocal, symmetric, three-port junction may be written:

$$\begin{pmatrix} \alpha & \beta & \gamma \\ \gamma & \alpha & \beta \\ \beta & \gamma & \alpha \end{pmatrix}$$

where α is the reflection coefficient at each port and β and γ are the transmission coefficients at the other two ports. From conservation of energy, one must have

$$|\alpha|^2 + |\beta|^2 + |\gamma|^2 = 1$$

and

$$\alpha\beta^* + \beta\gamma^* + \gamma\alpha^* = 0.$$

It is obvious from the above relations that if $\alpha = 0$ (matched condition), then either $\beta = 0$ and $|\gamma| = 1$ or $\gamma = 0$ and $|\beta| = 1$, the conditions for a perfect circulator. A particular three-port turnstile circulator utilizing a ferrite rod has been analyzed in detail by Fowler.[2]

For a lossless, but not perfectly matched symmetrical three-port junction with $|\beta|$ close to unity and $|\alpha|$ and $|\gamma|$ small, the above relations show that

$$|\alpha| \approx |\gamma| \text{ and } |\beta| \approx \sqrt{1 - 2|\alpha|^2}.$$

Thus, minimum insertion loss corresponds to both maximum isolation and minimum VSWR looking into any of the three ports.

Circulators at both 70 and 140 kmc have been constructed by symmetrically loading H-plane Y junctions with Ferramic R-1 with an applied field perpendicular to the plane of the junction as shown in Fig. 1. This design is similar to that of Petterson[3] and Chait,[4] except that here the ferrite is inserted in a copper sleeve which serves to match the junction symmetrically.

The 70 kmc circulator uses a Ferramic R-1 cylinder 0.050 inch in diameter by 0.094 inch long, inserted in a 0.005-inch wall copper sleeve with about 0.010 inch of the ferrite protruding from the end of the sleeve. The sleeve is then inserted into a 0.060-inch hole, carefully centered in the junction. The applied magnetic field and length of sleeve inserted into the junction are then simultaneously adjusted for minimum insertion loss. The characteristics for a typical circulator are shown in Fig. 2.

* Received by the IRE, October 6, 1959. The work reported in this paper was performed at Lincoln Laboratory, Lexington, Mass., a center for research operated by Massachusetts Institute of Technology with the joint support of the U. S. Army, Navy, and Air Force.

[1] H. J. Carlin, "Principles of gyrator networks," *Proc. Symp. on Modern Advances in Microwave Techniques*, Polytechnic Institute of Brooklyn, Brooklyn, N. Y., pp. 175–204; November 8–10, 1954.

[2] H. Fowler, "Reciprocity and the Scattering Matrix of Ferrite Devices," presented at the Symp. on Microwave Properties and Applications of Ferrites, Harvard University, Cambridge, Mass.; April, 1956.

[3] T. Schaug-Petterson, private communication; January, 1957.

[4] H. N. Chait and T. R. Curry, "Y circulator," *J. Appl. Phys.*, vol. 30, pp. 152S–153S; April, 1959.

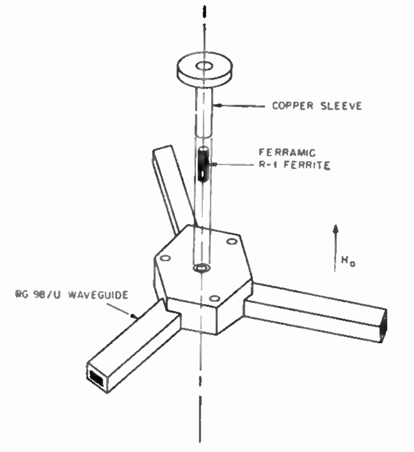

Fig. 1—Construction of 70-kmc circulator.

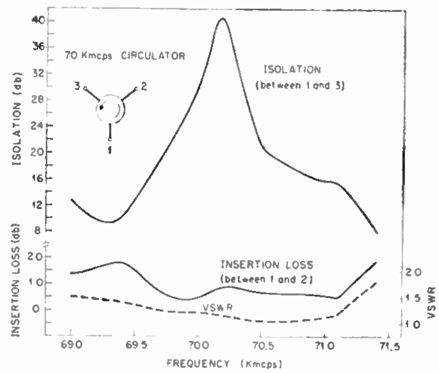

Fig. 2—Characteristics of a typical 70-kmc circulator.

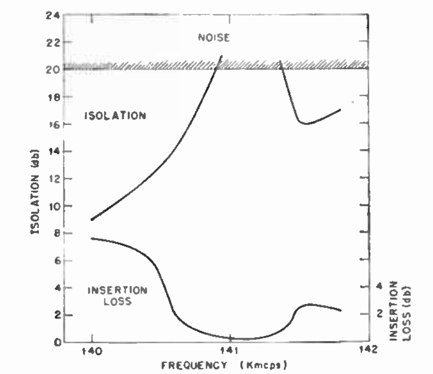

Fig. 3—Characteristics of a typical 140-kmc circulator.

At 140 kmc, the construction is similar using G-band waveguide with a Ferramic R-1 cylinder 0.035 inch in diameter by 0.066 inch long. The ferrite is held in a 0.002-inch wall copper sleeve with about 0.007 inch of the ferrite protruding from the sleeve. Typical characteristics for this circulator are shown in Fig. 3. Lack of matched loads, VSWR measuring equipment, and millimeter power at these frequencies prevent the measurement of the maximum isolation of this circulator.

The applied magnetic field required for both circulators was about 200 gauss. The characteristics are not very sensitive to magnetic field variations and the fringing field of a small alnico magnet is sufficiently homogeneous to operate both of the circulators described here.

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Frequency Dependence of the Noise and the Current Amplification Factor of Silicon Transistors*

The theory of noise in junction transistors has been well verified experimentally in the case of germanium transistors at ordinary operating temperatures and low injection levels.[1-3] There have been reports of discrepancies between theory and experiment, however, in the case of germanium transistors at low temperatures[4] (say that of liquid nitrogen) and in the case of silicon transistors.[5] The purpose of this paper is to present some recent experimental evidence concerning these discrepancies.

The shot noise in a junction transistor may be represented by an equivalent noise EMF e_e in series with the emitter junction and an equivalent noise current generator i in parallel with the collector junction.[6] Measurement of the latter generator, which reflects physical processes occurring at both the emitter and collector junctions, provides a very accurate method of checking the theory. The measurement can be carried out by operating the transistor in the grounded-base connection with the emitter circuit "ac open-circuited" and then comparing the output noise with that of a standard noise diode; this conveniently gives the equivalent saturated diode current I_{eq} of the generator i.

The early theory gave the following expression for I_{eq}:

$$I_{eq} = I_e[\alpha_0 - |\alpha|^2] \tag{1}$$

where α_0 is the low-frequency value of the ac current amplification factor α and I_e is the emitter current. At low frequencies this equation reduces to

$$I_{eq} = \alpha_0(1 - \alpha_0)I_e \tag{2}$$

and at high frequencies it becomes

$$I_{eq} = \alpha_0 I_e. \tag{3}$$

The results reported here were obtained from measurements on several TI type 2N332 *n-p-n* grown junction silicon transistors. All of the transistors tested showed essentially the same behavior.

Fig. 1 shows the results of measurements of I_{eq} plotted as a function of emitter current with frequency as a parameter. Also shown are three theoretical curves: One corresponds to the value $\alpha_0(1-\alpha_0)I_e$, the second to the value $\alpha_{dc}(1-\alpha_{dc})I_e$, and the third to the value $\alpha_{dc}I_e$; here α_{dc} is the dc current amplification factor. In general, for silicon transistors $\alpha_{dc} \neq \alpha_0$. The $\alpha_{dc}I_e$ curve forms an excellent low-current asymptote for the experimental curves. The $\alpha_{dc}(1-\alpha_{dc})I_e$ curve forms an excellent "high-current" asymptote, providing a better fit to the data than does the $\alpha_0(1-\alpha_0)I_e$ curve predicted by the early theory. At each frequency the curve giving I_{eq} shows the transition from the "high-current" asymptote to the low-current asymptote. This behavior would agree with (1) if the high-frequency behavior of the current amplification factor α were strongly current dependent. That this is the case for these silicon transistors is shown in Fig. 2 which gives the normalized magnitude of α in db as a function of frequency with the emitter current as a parameter. For comparison, the average alpha cutoff frequency of these transistors at an emitter current of 1 ma is about 4 mc.

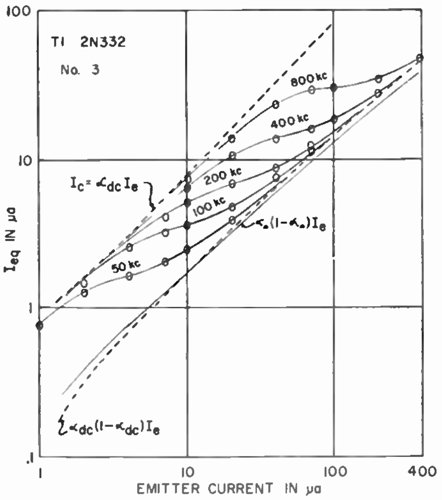

Fig. 1—Equivalent saturated diode current I_{eq} as a function of emitter current at several frequencies. Also shown are three theoretical curves.

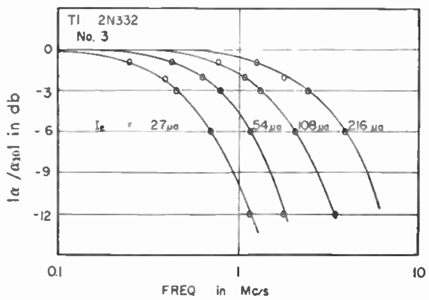

Fig. 2—Normalized magnitude of grounded base current gain α in db as a function of frequency with emitter current as a parameter.

Taking account of the difference between α_0 and α_{dc} van der Ziel has shown[7]

$$I_{eq} = I_e\left[\alpha_{dc} + |\alpha|^2 - \frac{2|\alpha|^2\alpha_{dc}}{\alpha_0}\right] \tag{4}$$

which reduces to

$$I_{eq} = I_e[\alpha_{dc}(1 - \alpha_{dc}) + (\alpha_0 - \alpha_{dc})^2] \tag{5}$$

for low frequencies; this in turn is equal to $\alpha_{dc}(1-\alpha_{dc})I_e$ in reasonable approximation since $(\alpha_0-\alpha_{dc})^2$ is a small quantity. At high frequencies $|\alpha|^2 \sim 0$ and hence $I_{eq}=\alpha_{dc}I_e$. If $\alpha_{dc}=\alpha_0$, (4) reduces to (1).

Fig. 3 compares some of the data shown in Fig. 1 with theoretical curves of I_{eq} as a function of frequency calculated by means of (4). Eq. (1) predicts the same behavior, but (4) gives a better fit to the data, especially as far as the asymptotes are concerned.

The conclusion is, therefore, that the transistor noise theory can adequately explain the observed data if the current dependence of the high-frequency behavior of α is taken into account and if a relatively small correction is made to account for the difference between α_{dc} and α_0. This small

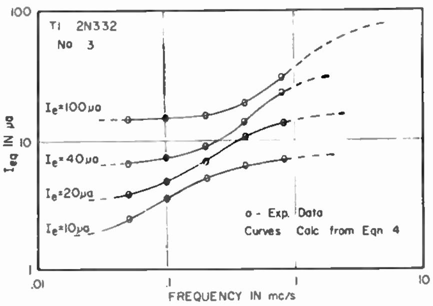

Fig. 3—Comparison of experimental data with theoretical curves of I_{eq} as a function of frequency for several values of emitter current.

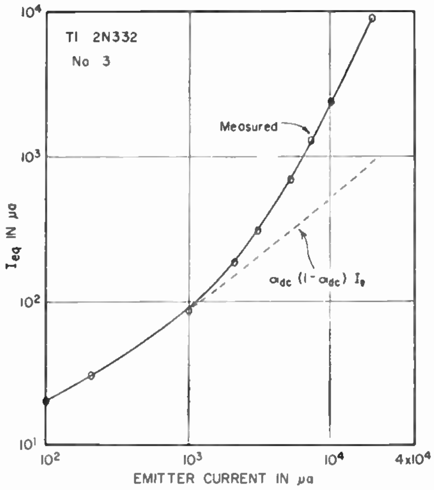

Fig. 4—Comparison of the observed performance of I_{eq} at relatively high emitter current with that predicted by the theory.

* Received by the IRE, May 4, 1959. Supported by U. S. Signal Corps Contract.

[1] G. H. Hanson and A. van der Ziel, "Shot noise in transistors," Proc. IRE, vol. 45, pp. 1538–1542; November, 1957.

[2] W. Guggenbuehl and M. J. O. Strutt, "Theory and experiments on shot noise in semiconductor diodes and transistors," Proc. IRE, vol. 45, pp. 839–854; June, 1957.

[3] E. G. Nielsen, "Behavior of noise figure in junction transistors," Proc. IRE, vol. 45, pp. 957–963; July, 1957.

[4] C. A. Lee and G. Kaminsky, "Temperature Dependence of Noise in Transistor Structures," presented before the American Physical Society, New York, N. Y.; January 29–February 1, 1958.

[5] B. Schneider and M. J. O. Strutt, "Theory and experiments on shot noise in silicon *p-n* junction diodes and transistors," Proc. IRE, vol. 47, pp. 546–554; April, 1959.

[6] A. van der Ziel, "Noise in junction transistors," Proc. IRE, vol. 46, pp. 1019–1038; June, 1958.

[7] A. van der Ziel, "High-frequency noise in silicon junction diodes and transistors," to be published.

correction is made necessary by trapping effects in the emitter junction. An investigation of the influence of trapping effects upon the emitter noise EMF e_e is in progress.

Fig. 4, which is included for the sake of completeness, compares the behavior of I_{eq} at high injection levels with that predicted by the present theory. Agreement is excellent up to $I_e \sim 2$ ma but there is an important discrepancy between theory and experiment at higher current levels. The theory as it now stands is thus correct for relatively low current densities only. A further study of high injection level effects is in progress.

The author wishes to express his gratitude to Dr. A. van der Ziel for suggesting this experiment and for helpful discussions.

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A Note on Scatter Propagation*

Some of our work in the field of scattering of electromagnetic energy from a periodic dielectric perturbation in the atmosphere[1] shows some interesting correlation to over-the-horizon scattering from random turbulence.

In this work, the theoretical predictions for the magnitude of the scattered electromagnetic energy from the periodic perturbation gives a first order effect of

$$\frac{P_r}{P_t} = \eta^2\delta^2 f(\theta) \sum_{n=1}^{a} b_n(k_{\sigma_n}) \tag{1}$$

where P_r and P_t are the reflected power and incident power, respectively, δ is the magnitude of the dielectric perturbation about a mean ϵ_0, and n is the number of periodic scatterers in the wave train. The function $f(\theta)$ expresses the geometrical dependency and will be left in the general form for this discussion, and the series defines the scatterer coefficients.

The theoretical work, which has been confirmed by experiments, predicts that the scattering effect is greatly enhanced if a certain wavelength ratio exists between the periodic disturbance and the electromagnetic energy. This ratio is given by

$$\frac{\lambda e}{\lambda_{\sigma_n}} = \frac{2}{a} g(\phi) \tag{2}$$

or in terms of wave numbers

$$\frac{k_{\sigma_n}}{k_e} = \frac{2}{a} g(\phi) \tag{3}$$

where the subscript σ_n refers to the scatterer and e to the electromagnetic energy. The factor a is an integer with values from 1 to ∞, and $g(\phi)$ describes the vector relationship between the electromagnetic energy and the dielectric perturbation.

If we consider two different electromagnetic frequencies and investigate the wavelength dependency of the scattered energy, it is found that for identical geometrical configuration

$$\frac{P_{r_1}}{P_{r_2}} = \frac{\delta_1^2 n_1^2}{\delta_2^2 n_2^2} \tag{4}$$

and

$$\frac{(k_\sigma)_1}{(k_e)_1} = \frac{(k_\sigma)_2}{(k_e)_2} = \frac{2}{a} g(\phi) = (2, 1, \tfrac{2}{3}, \cdots) g(\phi) \tag{5}$$

Thus

$$\frac{(k_\sigma)_1}{(K_\sigma)_2} = \frac{(k_e)_1}{(K_e)_2} \tag{6}$$

at each integral value of a.

If (3) is used to examine the dielectric turbulence spectra $L(k)$, which is defined as

$$L(k) = ck^{-m}, \tag{7}$$

it will be seen that there are numerous k_{σ_n} that are possible scatters. It is this point which is of importance, as it differs from the wave number accepted by other workers by the additional $1/a$ factor. Villar and Weisskopf give the wave number as

$$k_\sigma = 2k_e J(\phi). \tag{8}$$

Our experimental studies have shown that the values of $a = 2, 3, \cdots, 6$ are of importance when solving for the total scattered energy.

Fig. 1 indicates the multimodes which are possible scatterers. Since the total energy scattered is contributed by many wave numbers, the instantaneous magnitude of each must be considered. Such a consideration can also be used to establish the fading rate, a much more complex dependency than that now used. The same argument also applies when taking a specific value of m from (7) to predict wavelength dependency, at least for other than long time averages.

The power scattered for one electromagnetic frequency based on this multimode process has not yet been solved but is under study. It is easy, however, to use the above information to calculate the magnitude ratio of the scattered power for two electromagnetic wavelengths; say λe_1 and λe_2. All that is required is a consideration of δ and n at the two frequencies. The δ is proportional to the dielectric spectra, except near $k=0$ and is

$$\delta = aL(k) \tag{9}$$

or

$$\frac{\delta_1}{\delta_2} = \frac{L_1(k)}{L_2(k)} = \frac{k_1\sigma_n{}^{-m}}{k_2\sigma_n{}^{-m}} \tag{10}$$

which becomes

$$\frac{\delta_1}{\delta_2} = \left(\frac{k_{e_1}}{K_{e_2}}\right)^{-m} = \left(\frac{\lambda e_2}{\lambda e_1}\right)^{-m}. \tag{11}$$

The calculation for n is similar and can be found if a small increment of the scattering volume is selected. The vertical height of that volume is

$$h = n\lambda_{\sigma_n} \tag{12}$$

where λ_{σ_n} refers to any wave number, provided n also refers to that wave number.

For the same interaction volume height we find

$$\frac{n_1}{n_2} = \frac{(\lambda_\sigma)_2}{(\lambda_\sigma)_1} = \frac{ke_1}{Ke_2} = \frac{\lambda e_2}{\lambda e_1}. \tag{13}$$

Substituting (11) and (13) into (4) we have

$$\frac{P_{r_1}}{P_{r_2}} = \left(\frac{\lambda e_2}{\lambda e_1}\right)^{-2m} \left(\frac{\lambda e_2}{\lambda e_1}\right)^2 = \left(\frac{\lambda e_1}{\lambda e_2}\right)^{2m-2}. \tag{14}$$

The statistical theory of turbulence indicates that the exponent m should be 5/3 for the spectrum, but this does not indicate the time and space correlation between the various k_{σ_n}'s which is also required. Since the low-wave number components as well as the high-wave number components are involved, convective forces as well as others will play a part in determining the effective m, although the smaller k_{σ_n} contributes less energy than the larger values.

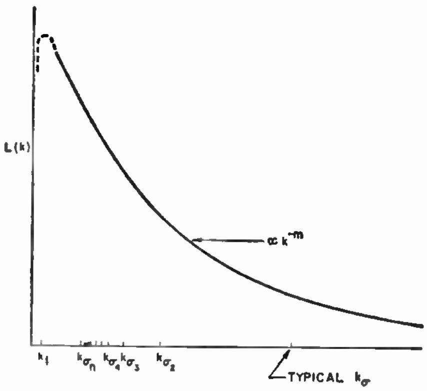

Fig. 1—Mean-square refractive index fluctuation spectrum.

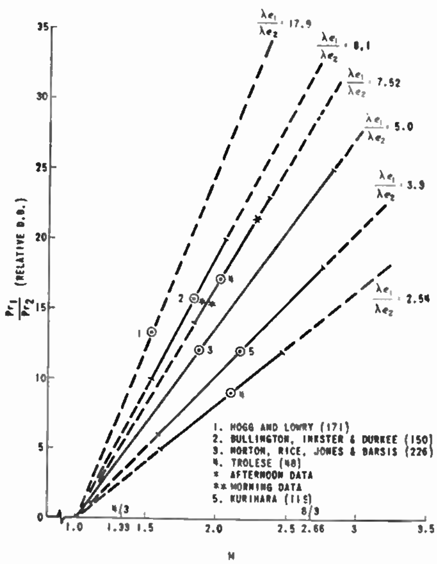

Fig. 2—Relative power ratio for different wavelengths.

Eq. (14) indicates the magnitude of the scattered power for a wavelength λe_1 compared to the scattered power at λe_2, where the interaction volume, antenna gain, etc., are considered equal. This equation is illustrated in Fig. 2 for various wavelength ratios. The broken curves are calculated for wavelengths at which experimental data had been published. Average data are shown by

* Received by the IRE, March 24, 1959.
[1] This work was under contract to the Office of Naval Res., Nonr. 1670(00), Subcontract FW1001 to the Bell Helicopter Corp.

circled points, with reported variation indicated by the solid part of the curves. The correlation between the data and the value of m is in part indicated by the path length, terrain, and altitude reported by several of the workers. For rough terrain and short path-length propagation, a high value of m could be expected. The lower values of m would hold for a flat terrain and long path-length scattering. A corresponding change in m is true for data reported for daytime and nighttime scattering. During the late part of the day, convective forces are more dominant than during the early morning.

Further data are required to test the theory more thoroughly and to complete the calculation of scattered energy. The author requests that workers having data on scattered energy at two frequencies communicate with him.

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Duplexing a Solid-State Ruby Maser in an X-Band Radar System*

Previous attempts to use maser preamplifiers in conventional pulse radars have been unsuccessful because the leak-through power from standard duplexing circuits can greatly reduce the amplifying characteristic of the maser. The required excess of spin population in the excited state is reduced by strong leak-through signals from the transmitter at the maser amplifying frequency. The purpose of this paper is to describe how this problem was solved for an X-band radar having a peak pulse power of 150 kw.

Leak-through power in a conventional TR circuit was minimized by careful selection of components. The peak value varied from 1 to 10 milliwatts, depending on a number of parameters, including frequency and temperature. In addition, pulse-to-pulse variations of up to 25 per cent were observed. A maser gain of 10 db was achieved under these conditions, but it was marginal and subject to fluctuations caused by variations in the leak-through power. The operation of the maser was judged unsatisfactory without further leak-through protection. Laboratory measurements indicated that an additional 30 db of isolation, which reduces the peak leak-through power to less than 10 μw, would allow near-optimum maser performance (25–30-db gain for a pulse repetition rate of 416 PPS and a pulse duration of 2.35 μsec). A direct method of providing the required additional isolation is to use a conventional TR circuit and to insert some RF switching device between the TR tube and the maser. (See Fig. 1.) It is necessary, however, to keep the insertion loss ahead of the maser as low as possible because of the effect such loss has on the noise temperature of the system.

* Received by the IRE, May 21, 1959.

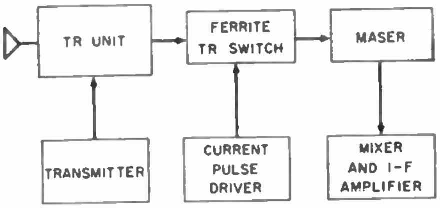

Fig. 1.

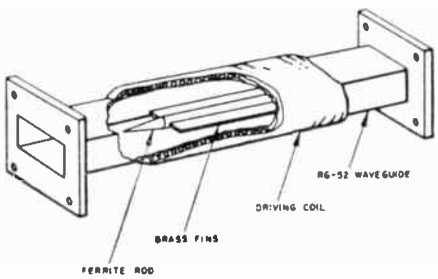

Fig. 2.

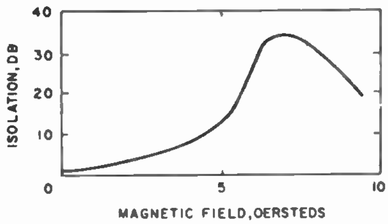

Fig. 3.

The problem of obtaining successful operation of the maser preamplifier in the radar was thus resolved into that of designing a high-speed, electronically operated switch having at least 30 db of attenuation in one state and an absolute minimum of insertion loss in the other state. Since 0.1 db of loss is approximately equivalent to 7°K in noise temperature, the insertion loss of the switch should preferably be no more than a few tenths of a decibel.

Currently available ferrite switches have between 0.5- and 1.0-db loss and require a larger switching pulse than is desired. Consideration of the specific requirements of the TR problem led to the conception and development of a reflective type of ferrite switch having an insertion loss of 0.25 db and an isolation of more than 30 db over a 120-mc band. A cutaway view of the switch is shown in Fig. 2. The isolation provided by the switch as a function of applied dc magnetic field is plotted in Fig. 3. The switch is pulsed to its attenuating condition when the transmitter is on, and it operates with very low loss during the receiving interval. The ferrite rod in the center of the waveguide has a conducting fin that extends from each side parallel to the broad wall. Without a dc magnetic field, the ferrite rod acts essentially as a low-loss dielectric. The tapered ends prevent reflections due to mismatch, and the fins do not disturb the transmission of the normal mode because they are in an equipotential plane. Thus the switch transmits with very low loss. When a longitudinal magnetic field is applied, the wave is converted into a mode which does not propagate through the structure. This results in the creation of a very high reflection at the input, and an attenuation of more than 30 db is obtained over a 120-mc band during the energizing pulse.

With the additional protection provided by the ferrite TR switch, the measured noise temperature of the maser, circulator, mixer, and IF amplifier combined was 65°K. The noise temperature of the over-all receiver, including that due to losses in the ferrite TR switch, the TR tube, waveguide, and rotary joints was 173°K, an excellent value compared with the 1500° to 2500°K noise temperature of a good X-band radar receiver without a maser preamplifier.

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Nomographs for Designing Elliptic-Function Filters*

The two nomographs in Henderson's article[1] can be simplified considerably. Every functional relation which can be expressed as

$$f(x_1) + g(x_2) + h(x_3) = 0 \tag{1}$$

can be represented in the form of the nomographs of Figs. 1 and 2. Fig. 1 corresponds to the nomograph of Fig. 2 of Henderson's article and represents

$$\Delta^4 = \frac{10^{.1a_{\max}} - 1}{10^{.1a_{\min}} - 1} \tag{2}$$

or

$$4 \log \Delta = \log (10^{.1a_{\max}} - 1) - \log (10^{.1a_{\min}} - 1). \tag{2a}$$

The connection between our notation and those used by Henderson is given by

$$\bar{\rho} = a_{\max} \quad [\text{db}]$$
$$\bar{\alpha} = a_{\min} \quad [\text{db}]$$
$$k_1 = \Delta^2$$
$$k_2 = \left(1 + \frac{\Delta\omega}{\omega_L}\right)^{-1}. \tag{3}$$

The use of Δ instead of k_1 is justified by the fact that this parameter has been tabulated.[2]

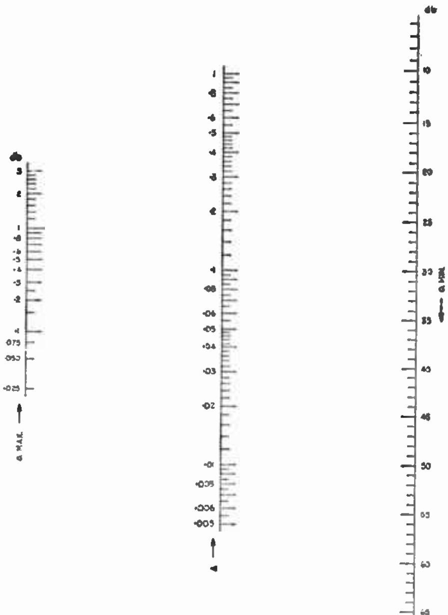
Our Fig. 2 corresponds to Fig. 3 of Henderson and the corresponding (11) of the cited article is already in the form of our (1).

The use of Fig. 1 can be extended for greater $a_{\min}$ values by using $a'_{\min} = a_{\min} - 40n$ db and multiplying the Δ value obtained for $a'_{\min}$ by 10^{-n}. Fig. 2 covers the entire range which may occur in practice. For small Δ values Henderson gives an approxi-

* Received by the IRE, May 7, 1959.

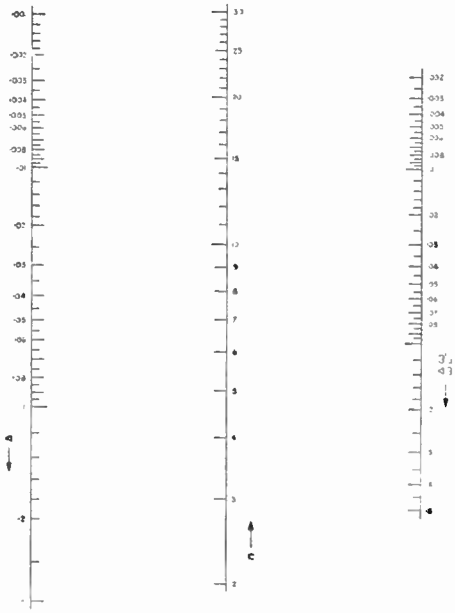
[1] K. W. Henderson, "Nomographs for designing elliptic-function filters," PROC. IRE, vol. 46, pp. 1860–1864; November, 1958.

[2] E. Glowatzki, "Sechsstellige Tafel der Cauer-Parameter," *Abhandl. bayer. Akad. Wiss.*, Heft 67, pp. 1–37; 1955. (Munich, Ger.)



$$\Delta = \sqrt{\frac{\epsilon^{20\ \max.} - 1}{\epsilon^{20\ \min.} - 1}}$$

Fig. 1.



$$q(\Delta^2) = \left[q\left(\frac{1}{1 + \frac{\Delta\omega}{\omega_L}} \right) \right]^n$$

Fig. 2.

mate formula. A corresponding formula for small $\Delta\omega/\omega_L$ values is given by

$$\log_{10} q(k_2) \doteq -\frac{\frac{(\pi \log_{10} e)^2}{16}}{\log_{10} \frac{1}{1 - k_2^2}}$$

$$\doteq -\frac{(\pi \log_{10} e)^2}{\log_{10} \frac{8}{\frac{\Delta\omega}{\omega_L}}}. \quad (4)$$

This leads to the approximate formula:

$$n \doteq 7.446(\log_{10} 2/\Delta)\left(\log_{10} 8 \middle/ \frac{\Delta\omega}{\omega_L}\right) \quad (5)$$

which is valid if

$$\Delta \ll 1; \quad \frac{\Delta\omega}{\omega_L} \ll 1$$

but gives sufficiently accurate estimates even with moderately small parameters.

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Shot Noise in Transistors*

The theory of shot noise in transistors presented in an earlier paper[1] has to be modified for silicon transistors to take account of trapping effects in the emitter space-charge region.[2,3] It is the aim of this paper to discuss this modification.

Let I_e and I_c be the dc emitter and collector current, respectively, then

$$I_c = \alpha_{dc} I_e + (I_c)_{sat} \quad (1)$$

where α_{dc} is the dc current amplification factor and $(I_c)_{sat}$ the collector saturated current. In silicon transistors $(I_c)_{sat} \simeq 0$, so that $\alpha_{dc} = I_c/I_e$. The low-frequency current amplification factor α_0 is thus

$$\alpha_0 = \partial I_c/\partial I_e = \alpha_{dc} + I_e \partial\alpha_{dc}/\partial I_e. \quad (2)$$

Because trapping effects in the emitter space-charge region influence I_e but do not affect I_c, α_{dc} depends upon I_e; consequently α_0 and α_{dc} will differ.

If v_e is the ac voltage across the emitter junction, then the ac emitter current is $i_e = Y_e v_e$ and the signal transfer properties of the transistor are described by a current generator $Y_{ce} v_e = \alpha i_e$ in parallel to the collector junction. Consequently, the current amplification factor is

$$\alpha = \frac{Y_{ce}}{Y_e} \quad \text{and} \quad \alpha_0 = \frac{G_{ceo}}{G_{eo}} \quad (3)$$

where G_{ceo} and G_{eo} are the low-frequency values of Y_{ce} and Y_e, respectively:

$$G_{ceo} = \frac{\partial I_c}{\partial V_e} = \frac{eI_c}{kT}, \quad G_{eo} = \frac{\partial I_e}{\partial V_e} = \frac{eI_e}{m_e kT}. \quad (4)$$

The first equation follows from the fact that $I_c = \text{const.} \exp(eV_e/kT)$ whereas the factor m_e in the second equation takes the trapping effects in the emitter transition region into account.[2] According to (4):

$$m_e = \frac{G_{ceo}}{G_{eo}} \cdot \frac{I_e}{I_c} = \frac{\alpha_0}{\alpha_{dc}}. \quad (4a)$$

We now turn to the noise. As usual,[1] we represent the noise first by a noise current generator i_1 across the emitter junction and a noise current generator i_2 across the collector junction. If the transistor is an n-p-n transistor, the noise is caused by five different processes (Fig. 1):

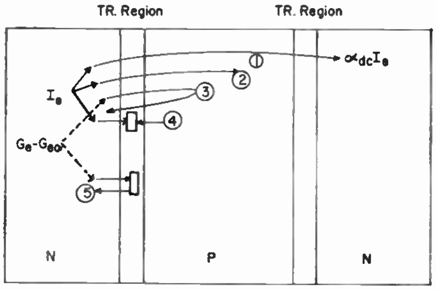

Fig. 1.

1) Noise due to electrons going from emitter to collector,
2) noise due to electrons going from emitter to base,
3) noise due to electrons injected into the base and returning to the emitter,
4) noise due to electrons trapped in the emitter space-charge region and recombining with holes coming from the base, and
5) noise due to electrons trapped in the emitter space-charge region and returning to the emitter after being detrapped thermally.

All five processes contribute to i_1 but only process 1) contributed to i_2. Full shot noise should be attributed to the currents associated with processes 1) and 2). Because 3) and 5) are thermal processes that give a contribution $(G_e - G_{eo})$ to the high-frequency emitter conductance G_e, full thermal noise should be associated with $(G_e - G_{eo})$. In process 4) there is a random time delay τ between the trapping of an electron and the subsequent trapping of a hole. At low frequencies $(\omega\tau \ll 1)$ the trapping of the electron and the hole can be considered to occur simultaneously; since the combined event transfers a full charge e across the junction and the individual events are independent and occur at random, process 4) also should give full shot noise of the current associated with it. At high frequencies, however, the two events can be considered as independent and the combined effect results in less noise.[2]

Ignoring the high-frequency effect in process 4) one would thus expect

$$\overline{i_1^2} = 2eI_e\Delta f + 4kT(G_e - G_{eo})\Delta f \quad (5)$$

$$\overline{i_2^2} = 2eI_c\Delta f \quad (6)$$

$$\overline{i_1^* i_2} = 2kTY_{ce}\Delta f = 2kT\alpha Y_e \Delta f, \quad (7)$$

(the asterisk denotes the conjugate complex quantity) in agreement with the earlier theory;[1] the proof of (7) is the same as before. Because of a different assumption about the noise of process 4), Schneider and Strutt's (23) differs from our (5) whereas their (24)

* Received by the IRE, May 16, 1959. Supported by U. S. Signal Corps Contract.

[1] A. van der Ziel and A. G. T. Becking, "Theory of junction diode and junction transistor noise," PROC. IRE, vol. 46, pp. 589–594; March, 1958.

[2] B. Schneider and M. J. O. Strutt, "Theory and experiments on shot noise in silicon p-n junction diodes and transistors," PROC. IRE, vol. 47, pp. 546–554; April, 1959.

[3] C. T. Sah, R. N. Noyce, and W. Shockley, "Carrier generation and recombination in p-n junctions and p-n junction characteristics," PROC. IRE, vol. 45, pp. 1228–1243; September, 1957.

and (24a) agree with our (6) and (7), respectively.

The two noise current generators i_1 and i_2 are now replaced by a noise emf $e_e = i_1/Y_e$ in series with the emitter and a noise current generator $i = (i_2 - \alpha i_1)$ in parallel to the collector junction.[1] Consequently, if "Re" stands for real part:

$$\begin{aligned}\overline{i^2} &= \overline{(i_2^* - \alpha^* i_1^*)(i_2 - \alpha i_1)}\\ &= \overline{i_2^2} - 2\,\mathrm{Re}\,(\alpha^* \overline{i_1^* i_2}) + |\alpha|^2 \overline{i_1^2}\\ &= 2e\left[I_c + I_e|\alpha|^2 - 2I_e|\alpha|^2 \frac{\alpha_{dc}}{\alpha_0}\right]\Delta f\\ &= 2eI_{eq}\Delta f \qquad (8)\end{aligned}$$

since $4kTG_{eo}\Delta f = 4\ eI_e\Delta f(\alpha_{dc}/\alpha_0)$ according to (4a). Hence:

$$I_{eq} = I_c + I_e|\alpha|^2 - 2I_e|\alpha|^2 \frac{\alpha_{dc}}{\alpha_0} \qquad (9)$$

which reduces to the earlier equation:[1]

$$I_{eq} = I_c - |\alpha|^2 I_e \qquad (9a)$$

if $\alpha_{dc} = \alpha_0$. Substituting (4a), (9) becomes

$$I_{eq} = I_c + I_e|\alpha|^2 - \frac{2I_e|\alpha|^2}{m_e}. \qquad (10)$$

Eq. (9) expresses I_{eq} in terms of more accessible quantities; (10) is given for comparison with a corresponding relation deduced from Schneider and Strutt's (25) for the noise figure of a transistor:

$$I_{eq} = I_e = \frac{I_e|\alpha|^2}{m_e}. \qquad (10a)$$

This lies halfway between (9a) and (10); the discrepancy is caused by Schneider and Strutt's (23). Chenette's experimental data[4] agree better with (9) than with (9a) or (10a), indicating that the assumption of full shot noise in the current due to process 4) is allowed.

We finally calculate the high-frequency effect due to process 4). The elementary event consists of two pulses of random time delay τ each transferring a charge[2] $\frac{1}{2}e$. If all events had the same time delay τ:

$$\begin{aligned}\overline{i_4^2} &= 1/2eI_4\Delta f\,|1 - \exp(-j\omega\tau)|^2\\ &= eI_4\Delta f(1 + \cos\omega\tau) \qquad (11)\end{aligned}$$

where I_4 is the current associated with process 4). Assuming an average lifetime τ_0 of the trapped carriers, one would expect a probability $\exp(-\tau/\tau_0)d(\tau/\tau_0)$ of a time delay between τ and $\tau + d\tau$. Consequently:

$$\begin{aligned}\overline{i_4^2} &= \int_0^\infty eI_4\Delta f(1 + \cos\omega\tau)\exp(-\tau/\tau_0)d(\tau/\tau_0)\\ &= 2eI_4\Delta f - eI_4\Delta f\,\frac{\omega^2\tau_0^2}{1 + \omega^2\tau_0^2}. \qquad (11a)\end{aligned}$$

The effect can thus be taken into account by subtracting a term $-eI_4\Delta f\omega^2\tau_0^2/(1+\omega^2\tau_0^2)$ from (5) or by adding a term $\frac{1}{2}I_4\omega^2\tau_0^2/(1+\omega^2\tau_0^2)$ to (9). This effect may be hardly noticeable if τ_0 is sufficiently small.

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[4] E. R. Chenette, "Frequency dependence of the noise and the current amplification factor of silicon transistors," Proc. IRE, this issue, p. 111.

General Aspects of Beating-Wave Amplification*

This paper reports a suggestion as to a new way to achieve large amplification of microwave-frequency signals on an electron beam without involving the problem of survival of the helix or other traveling-wave circuits under high-power conditions. This problem of circuit survival has caused substantial difficulty in high-power continuous-wave, traveling-wave amplifiers.

It has been recognized for some time that it is possible under certain conditions to produce a growing wave by employing two beams (which might be concentric, one inside the other) of differing dc beam velocities. This is done in the Haeff growing-wave electron wave amplifier. The theory of this device shows that in order to produce a growing wave, there must be rather close coupling between the two beams. Also, even when there is adequate coupling, the growth rate may not be very substantial when thought of in terms of service as a power amplifier.

However, even when there is relatively weak coupling between the two beams, there may exist substantial deviations in phase velocities of the waves relative to the beam velocity, and the deviations are different for the several waves that are present. Thus, it does not require intimate coupling to produce two waves of substantially differing phase velocities, in addition to perhaps several other waves. The recent reporting, by Rowe, of the Crestatron type of tube,[1] which provides substantial traveling-wave amplifier gain as a result of the beating among waves that are not growing waves but have differing phase velocities, shows that it is possible to obtain substantial amplification in a working device without a growing wave. The gain results from power input at a node of a standing-wave pattern, and output at an antinode.

It should therefore be possible, by using the standing-wave pattern between uniform-amplitude space-charge waves of unlike phase velocities, to use the beating between these waves to produce substantial gain of the electron-wave type without growing waves. This suggests the possibility of substantial gain of the general broad-band traveling-wave amplifier type, without having the problem of maintaining a circuit which must be kept close to the beam, and that must correspondingly handle substantial amounts of RF power over a considerable length.

There remains, of course, the problem of introducing the signal onto the beam. On a narrow-band basis this is quite straightforward, in that an arrangement similar to that of a klystron gap can be employed to introduce the signal into the beam quite satisfactorily. This suggests then that it should be possible to build a klystron, having in the drift space two concentric beams, which will have substantial gain between the two gaps (conceivably either by means of a growing wave or by the beating-wave method), making the klystron gain cease to be primarily a function of the impedance of the output gap. The problem of utilizing the beating wave phenomena in a klystron-type device has recently been analyzed by one of the authors.[1]

However, quite apart from this klystron type of device, it should be possible to devise a broad-band type of drive for the space-charge wave combination of two beams without experiencing the problems of severe dissipation that occur in a long circuit. This results from the fact that the launching of the signal can occur in a relatively short axial length of circuit; whereas, if one wishes to use a traveling-wave amplifier with a long circuit, the coupling must be maintained over quite a long range.

There is also suggested the possibility of employing this two-wave principle in a crossed-field situation, employing two strip beams, again without a circuit over the major length of the device, and employing a difference of phase velocity and a standing wave pattern to produce the gain, without a growing wave, and essentially without a slow-wave circuit.

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* Received by the IRE, April 27, 1959.

[1] J. E. Rowe, "Theory of the Crestatron: a forward-wave amplifier," Proc. IRE, vol. 47, pp. 536–545; April, 1959.

A Strip-Line *L*-Band Compact Circulator*

The realization of a ferrite circulator at *L*-band presents some difficulties, both in obtaining suitable materials with a good figure of merit in this frequency region[1] and in reducing within reasonable limits the geometrical dimensions. The use of polycrystalline YIG or of magnesium manganese aluminates having a narrow resonance linewidth can offer a good approach for the first problem; but the engineering work still meets appreciable practical difficulties in obtaining a reasonably compact device if the classical types of circulators using Faraday rotators or nonreciprocal differential phase shifters are considered.

The utilization of a waveguide *Y* junction containing symmetrically located ferrite has been proved to give circulator action.[2] In order to solve the problem of compactness for the *L*-band case, we used a similar geometry for a stripline circulator which has been realized with successful results. This circulator utilizes a symmetrical Y-type junction of three flat-strip transmission lines using air dielectric and containing two disks

* Received by the IRE, May 11, 1959.

[1] B. Lax, "Frequency and loss characteristics of microwave ferrite devices," Proc. IRE, vol. 44, pp. 1368–1386; October, 1956.

[2] H. N. Chait and T. R. Curry, "*Y* circulator," *Proc. Fourth Symp. on Magnetism and Magnetic Materials;* also *J. Appl. Phys.*, suppl. to vol. 30, pp. 152S–153S; April, 1959.

of YIG located at the center of the junction as schematically indicated in Fig. 1. The magnetostatic polarizing field H_0 can be supplied by a small permanent magnet in such a way that the total volume does not exceed $3\frac{1}{2} \times 6 \times 7\frac{1}{2}$ inches. The polarizing field is of greater value than that required for the ferromagnetic resonance.

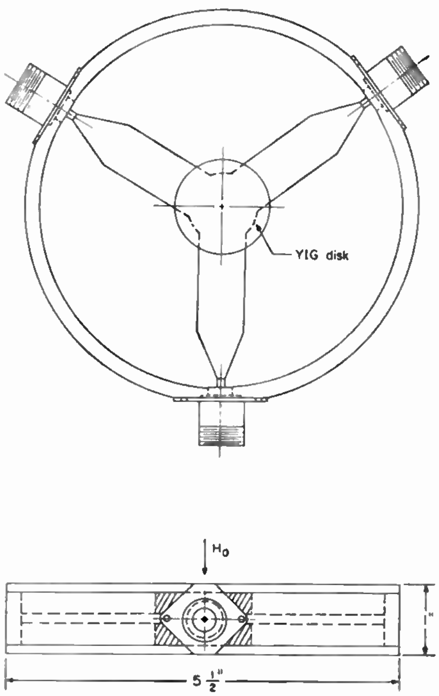

Fig. 1—Geometrical configuration of the Y-junction strip-line circulator.

This type of circulator has been tested at different frequencies in the range 950–1600 mc and, by choosing the appropriate dimensions of the YIG discs, satisfactory results have been obtained in all this frequency range. With a fixed H_0 field, the performance at the middle frequency can always give an isolation >25 db, insertion loss <0.4 db and input standing wave ratio <1.1. Curves showing the performance of the circulator vs the magnetostatic field H_0 at the frequency of 975 mc are plotted in Fig. 2. It appears from these curves that it is possible to combine at a single frequency an isolation >35 db with an insertion loss of 0.3 db and an input SWR ≤1.05.

The measurements of the insertion loss were taken with a tunable bolometer and are believed to have an accuracy of approximately ±0.05 db. The tunable bolometer was matched in each condition for an input SWR ≤1.02. The performance vs frequency for the same geometry and a fixed H_0 field is shown in Fig. 3. In this case, the isolation is ≥25 db over a bandwidth of 50 mc and in the same frequency region the insertion loss is approximately 0.3 db and the input SWR is ≤1.25.

The results obtained show that this configuration offers an attractive solution for the realization of low-power compact circulators in the low-frequency region of the microwave spectrum.

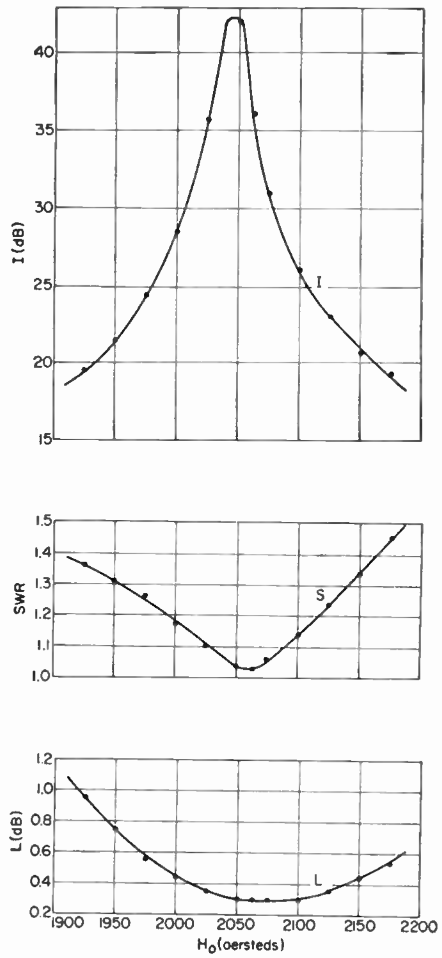

Fig. 2—Performance vs magnetostatic field of the strip-line circulator at 975 mc.

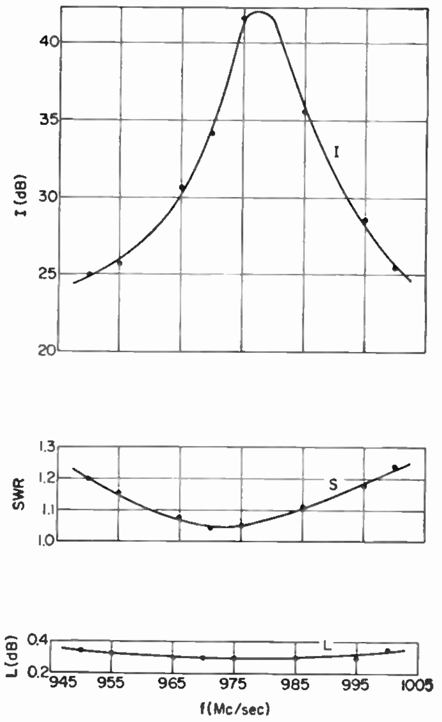

Fig. 3—Performance vs frequency of the strip-line circulator with a fixed polarizing magnetic field.

List of Symbols

I: Isolation
L: Insertion Loss
S: Input Voltage-Standing Wave Ratio.

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The Impulse Response of an All-Pass Network Having Infinite-Order Phase Distortion*

In an earlier paper,[1] Di Toro considered the separate effects of phase and amplitude distortion on the impulse and step function responses of linear networks. He showed that the phase distortion is principally responsible for the oscillatory character of the transient response, and presented curves showing the transient responses for third-, fifth-, and seventh-order phase distortion. Here, the effect of extremely high-order phase distortion is considered. It is shown that, as the order of the phase distortion tends to infinity, the impulse response tends to a sin x/x type of behavior, identical to the impulse response of a linear-phase system having an infinitely sharp amplitude cutoff.

Given a system function

$$y(p) = \frac{1 + g_1 p + g_2 p^2 + \cdots + g_s p^s}{1 + h_1 p + h_2 p^2 + \cdots + h_r p^r}, \quad (1)$$

where the g's and h's are all real, $r > s$, and $p = j\omega$; we then define a propagation factor $\Gamma(j\omega)$ such that

$$\Gamma(j\omega) = j\omega t_d + \Delta A + j\Delta B, \quad (2)$$

where t_d represents a pure time delay, while ΔA and ΔB represent amplitude and phase distortions in nepers and radians respectively and are expressed in terms of power series in ω, with coefficients that are functions of the g's and h's. We are concerned here with the special case in which

$$\Delta A = 0$$
$$\Delta B = (b\omega)^n, \quad (3)$$

where n is an odd integer representing the order of the phase distortion. Now the time delay of the system is the slope of the phase function, and is equal to

$$\frac{d\Delta B}{d\omega} = nb^n \omega^{n-1}. \quad (4)$$

As n tends to infinity, $d\Delta B/d\omega$ tends to zero for $\omega b < 1$, and to infinity for $\omega b > 1$. This means that all frequencies for which $\omega b < 1$ suffer no time delay (other than that caused by the $j\omega t_d$ term), while all frequencies for which $\omega b > 1$ suffer infinite time delay; that

* Received by the IRE, May 11, 1959.
[1] M. J. Di Toro, "Phase and amplitude distortion in linear networks," Proc. IRE, vol. 36, pp. 24–36; January, 1948.

is, they never appear in the output of the system. A moment's thought will show that this situation is analogous to a pure amplitude-distortion system in which all frequencies below an infinitely sharp cutoff appear undistorted in the output, while those above this cutoff frequency never appear. To an observer at the output, therefore, infinite time delay has the same effect as an infinite amplitude attenuation, and one might expect that the transient responses in these two cases would be similar. That this is so will now be demonstrated.

The impulse response of an all-pass network having a phase distortion $\Delta B = (b\omega)^n$ radians may be shown to be[2]

$$y = \frac{1}{\pi}\int_0^\infty \cos(\omega\tau - (b\omega)^n)d\omega. \quad (5)$$

The solution of this is[3]

$$y = \frac{1}{\pi n b}\sum_{s=0}^{\infty}\left(-\frac{\tau}{b}\right)^s \frac{1}{s!}\Gamma\left(\frac{s+1}{n}\right) \times \cos\frac{\pi}{2n}[1+s(1+n)]. \quad (6)$$

We begin by expanding (6):

$$\begin{aligned} y = \frac{1}{\pi b}\Big[\frac{1}{n}\Gamma\Big[\frac{1}{n}\Big]\cos\frac{\pi}{2n} &- \left(\frac{\tau}{b}\right)\frac{1}{n}\Gamma\left(\frac{2}{n}\right)\cos\frac{\pi}{2n}[2+n] \\ &+ \left(\frac{\tau}{b}\right)^2\frac{1}{2!n}\Gamma\left(\frac{3}{n}\right)\cos\frac{\pi}{2n}[3+2n] \\ &- \left(\frac{\tau}{b}\right)^3\frac{1}{3!n}\Gamma\left(\frac{4}{n}\right) \times \cos\frac{\pi}{2n}[4+3n]\cdots\Big]. \end{aligned} \quad (7)$$

Now we note that

$$\frac{1}{n}\Gamma\left(\frac{z}{n}\right) = \frac{1}{z}\Gamma\left(1+\frac{z}{n}\right). \quad (8)$$

In addition, the cosine terms may be simplified. The final result is

$$\begin{aligned} y = \frac{1}{\pi b}\Big[\Gamma\left(1+\frac{1}{n}\right)\cos\frac{\pi}{2n} &+ \left(\frac{\tau}{b}\right)\frac{\Gamma\left(1+\frac{2}{n}\right)}{2}\sin\frac{\pi}{n} \\ &- \left(\frac{\tau}{b}\right)^2\frac{\Gamma\left(1+\frac{3}{n}\right)}{3\cdot 2!}\cos\frac{3\pi}{2n} \\ &- \left(\frac{\tau}{b}\right)^3\frac{\Gamma\left(1+\frac{4}{n}\right)}{4\cdot 3!}\sin\frac{2\pi}{n} \\ &+ \left(\frac{\tau}{b}\right)^4\frac{\Gamma\left(1+\frac{5}{n}\right)}{5\cdot 4!}\cos\frac{5\pi}{2n}\cdots\Big]. \end{aligned} \quad (9)$$

When n tends to infinity, the sine terms drop out, and the cosine terms plus the gamma function factors tend to unity. We then have the result

$$y = \frac{1}{\pi b}\left[1 - \left(\frac{\tau}{b}\right)^2\frac{1}{3!} + \left(\frac{\tau}{b}\right)^4\frac{1}{5!} - \left(\frac{\tau}{b}\right)^6\frac{1}{7!}\cdots\right]. \quad (10)$$

Upon multiplying the series by τ/b, and the factor in front by b/τ, we get the final result

$$y = \frac{1}{\pi b}\,\frac{\sin\left(\frac{\tau}{b}\right)}{\left(\frac{\tau}{b}\right)}. \quad (11)$$

This is identical in form to the impulse response of the linear-phase system having infinitely sharp amplitude cutoff. Thus the physical argument presented earlier has been shown to be mathematically correct.

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[2] *Ibid.*, (25).
[3] *Ibid.*, (26).

The Production of Whistlers by Lightning*

An association between whistlers and other atmospherics (atmospheric clicks) was noted 30 years ago by Eckersley, and about the same time a causal relation between lightning strokes and whistlers was reported by Tremellen.[1] It has been generally accepted since that time that whistlers originate in the low-frequency radiation emitted by lightning strokes.[2] Recent observations by Morgan[3] have gone far to establish the circumstantial evidence relating whistlers to lightning strokes. Despite these facts there remain some difficult points in the explanation of whistler production by individual lightning strokes.

Present evidence on the rate of whistler incidence shows the following points: a) Not all lightning strokes produce whistlers. b) In some storms there may be a high correlation between lightning strokes and resulting whistlers. c) Atmospherics which produce whistlers often have recognizable characteristics (high intensity, special sound, etc.). d) Whistlers are associated frequently with atmospherics presenting no special characteristics. e) A given storm may produce whistlers during part of its life but not at other times.

The somewhat paradoxical nature of this evidence suggests that the whistler-producing capacity of individual lightning strokes is governed in part by conditions which depend on the nature of a whole storm area. It seems likely that these conditions may be related to the region above the storm area, but below the ionosphere.

The static electrical conductivity of the atmosphere increases rapidly with altitude and is already so large at about the 18 km level that this is known as the equalizing layer. This conductivity is unimportant in most considerations on radio propagation since it is produced by molecular ions rather than by electrons. However, in the case of the quasi-static and low-frequency fields from lightning strokes the situation is somewhat different.

The equalizing layer serves as an effective shield for the upper ionosphere against the slow electrical field changes associated with a thunderstorm. During the process of charge separation in the thundercloud there will appear an appreciable induced charge in the region above the cloud. This fact is emphasized further by the existence of electrical currents flowing upwards from a storm area to the equalizing layer.[4] When a lightning stroke occurs the emission of radiation is accompanied by a relaxation of the electric field in the region above the cloud as well as on the ground below it. Since the radiation field is emitted over a time interval of the order of 100–200 μsec, the currents produced in the equalizing layer below the D layer may have an appreciable influence on propagation conditions. Furthermore, the passage of this strong field might be associated with some release of electrons from negative ions. Lightning strokes from the clouds upwards into the higher atmosphere might also have some special influence on whistler production.

The evidence on whistler incidence becomes a little clearer if we infer from it that the region below the ionosphere represents an important coupling mechanism between the storm area and the ionosphere. Smith, Helliwell, and Yabroff[5] have suggested that the radiation energy from a lightning stroke may be transmitted through the ionosphere by means of electrical ducts reaching down to the E layer. No mechanism was proposed for the formation of these ducts. It seems possible that such an effect might arise through coupling conditions in the lower ionosphere in the manner suggested here.

If whistler production is related to a type of ionospheric control by the storm area, this connection may be reflected in meteorological conditions. An attempt to relate whistlers to meteorological factors has already been initiated independently by Mook.[6] It is of great interest that whistler studies may yield information about the lower atmosphere as well as the exosphere.

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* Received by the IRE, May 19, 1959.

[1] R. A. Helliwell and M. G. Morgan, "Atmospheric whistlers," Proc. IRE, vol. 47, pp. 200–208; February, 1959.
[2] L. R. O. Storey, "An investigation of whistling atmospherics," *Phil. Trans. Roy. Soc. (London) A*, vol. 246, pp. 113–141; July 9, 1953.
[3] M. G. Morgan, "Correlation of whistlers and lightning flashes by direct aural and visual observations," *Nature*, vol. 182, pp. 332–333; August 2, 1958.
[4] R. Gunn, "The electrification of precipitation and thunderstorms," Proc. IRE, vol. 45, pp. 1331–1358; October, 1957.
[5] R. L. Smith, R. A. Helliwell, and I. Yabroff, "The Trapping of Whistlers by Columnar Irregularities in the Outer Ionosphere," presented at the URSI-IRE Meeting, Washington, D. C.; May 4–7, 1959.
[6] C. P. Mook, "A preliminary meteorological study of the origin of whistlers," *J. Geophys. Res.*, July, 1959.

Reactance Transistor*

The equivalent circuit of reactance transistor is given in Fig. 1, where its operation frequency is lower, compared with its cutoff frequency.

The output admittance of reactance transistor is

$$Y = h_{22e} + \frac{h_{21e}}{Z_1 + h_{11e}} \cdot \frac{Z_1}{Z_2} \qquad (1)$$

where

$$h_{12e} \ll \frac{Z_1}{Z_1 + Z_2} \quad \text{and} \quad Z_2 \gg Z_1.$$

If

$Z_1 \gg h_{11e}$, (1) will be

$$Y = h_{22e} + \frac{h_{21e}}{Z_2}. \qquad (2)$$

and if

$Z_1 \ll h_{11e}$

$$Y = h_{22e} + \frac{h_{21e}}{h_{11e}} \cdot \frac{Z_1}{Z_2} \qquad (3)$$

Here, we check (2) and (3). The current amplification factor h_{21e} and the short circuit input impedance h_{11e} are functions of the emitter current I_e, as shown in Fig. 2, and the output admittance of reactance transistor varies in accordance with the value of the emitter current. The variation of h_{21e} is monotonous so far as the emitter current is from 0.1–3 ma; however, the variation of h_{11e} is approximately inversely proportional to emitter current. Then, the variation of output admittance Y is determined with h_{11e}, so that (3) is available. At low values of emitter current, (2) and (3) are available, since the variation of output admittance is determined with both h_{11e} and h_{21e}.

In (3) where $Z_2 = R_1$, $Z_1 = 1/j\omega C_1$, the output admittance Y becomes inductive and its equivalent inductance L_e is

$$Le = \frac{h_{11e}}{h_{21e}} C_1 R_1. \qquad (4)$$

and the frequency deviation $\Delta f/f$ is

$$\frac{\Delta f}{f} = \frac{L}{2C_1R_1}\,\delta\,\frac{h_{21e}}{h_{11e}} \qquad (5)$$

Where

$$Z_2 = \frac{1}{j\omega C_2}, \qquad Z_1 = R_2,$$

Y becomes capacitive and its equivalent capacitance C_e is

$$C_e = \frac{h_{21e}}{h_{11e}} C_2 R_2. \qquad (6)$$

and the frequency deviation $\Delta f/f$ is

$$\frac{\Delta f}{f} = \frac{C_2R_2}{2C}\,\delta\,\frac{h_{21e}}{h_{11e}} \qquad (7)$$

The frequency modulation circuit of an oscillator using this reactance transistor is illustrated in Fig. 3.

* Received by the IRE, May 14, 1959.

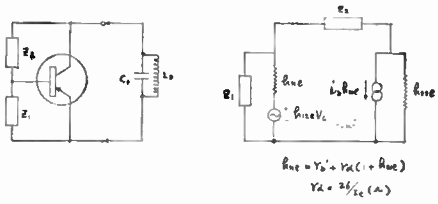

Fig. 1—Equivalent circuit of reactance transistor.

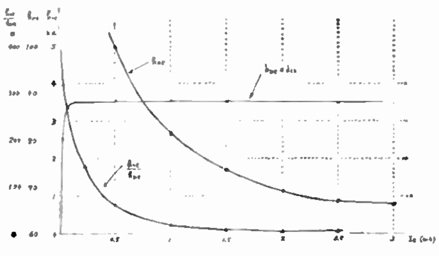

Fig. 2—h_{11e}, h_{21e}, h_{11}, h_{21}—I_e curves.

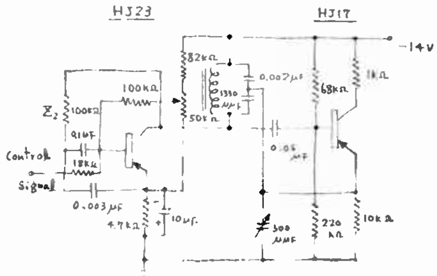

Fig. 3—Frequency modulated oscillator.

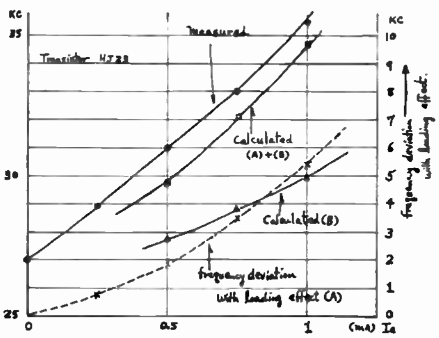

Fig. 4—Δf—I_e curve.

In this configuration, the frequency deviation curve is obtained in case a center frequency is 31.5 kc and a center biased emitter current is 0.6 ma, as shown in Fig. 4.

The difference between the measured curve and the calculated curve (A) with (4) and (5) arises from the frequency variation which is due to load the reactance transistor on oscillator.

The calculated curve (B) which takes this effect into account coincides with the measured one as shown in Fig. 4.

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A Method of Combining Two Frequencies*

Introduction

Recent interest in frequency synthesizers[1,2] has initiated studies of one of the most basic problems in electrical engineering, that is, means for combining two frequencies to produce a new frequency.

Of course, the most common method to accomplish this is to mix or heterodyne two waves in some type of nonlinear device so as to produce sum and difference components. One of these components is then selected by a bandpass filter. When such a mixing process is used in a synthesizer, either a large number of filters are required or some method of tuning filters must be provided in order to isolate the desired components. The phase shift technique of sideband generation may be used for isolating a desired component but this requires special wide-band phase shift devices. The purpose of this note is to introduce a new system for combining two frequencies.

New System

This new technique is based upon the fact that when two equal amplitude sine waves of difference frequencies f_1 and f_2 are combined additively and the resultant wave is passed through a perfect limiter the output is a phase-modulated sine wave. This phase-modulated wave can be viewed in a number of ways. One way of viewing it is as a sine wave of frequency F_1 phase-modulated by a sawtooth wave as shown in Fig. 1 (assuming that $F_1 < F_2$). Alternatively, it may be viewed as a sine wave of frequency F_2 modulated by the reversed sawtooth wave as shown in Fig. 2. A third possibility is to consider the limiter output as a sine wave of frequency $(F_1+F_2)/2$ phase-modulated by a square wave of fundamental frequency $(F_2-F_1)/2$ (Fig. 3).

It should be noted that the slope of the sawtooth phase modulated wave shown in Fig. 1 is such that the phase gains π radians per beat difference cycle. Thus, if $F_1 - F_2$ is equal to 1 mc, the phase would gain π radians every microsecond. However, the phase whip negates this gain of π radians. If we were to reverse the polarity of this wave at the time when the phase whip occurs, the π radian whip would be nullified and the wave would continue upward at a constant phase slope. A constant rate of phase change is indicative of a new frequency. This new frequency is π radians per beat cycle higher in frequency than F_1 and π radian per beat cycle less than F_2; thus, it is exactly half way between F_1 and F_2 or $(F_1+F_2)/2$.

Another way of looking at this would be to view the phase modulation component of the two tone wave at $(F_1+F_2)/2$ wherein the phase slope is nullified (Fig. 3). If this wave were reversed every time the phase switched, the resultant would be a wave having no

* Received by the IRE, May 14, 1959.

[1] B. Fisk and C. L. Spencer, "Synthesizer stabilized single-sideband systems," Proc. IRE, vol. 44, pp. 1680–1688; December, 1956.

[2] D. Makow, "Generations of oscillations with equally spaced frequencies in a given band," IRE Trans. on Communications Systems, vol. CS-5, pp. 13–20; September, 1957.

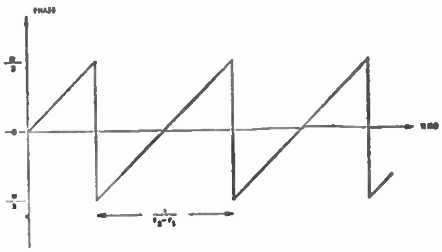

Fig. 1.

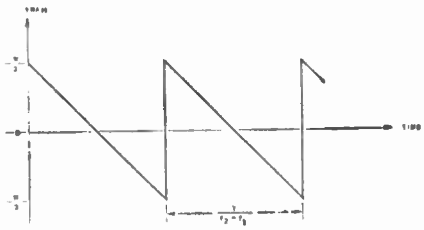

Fig. 2.

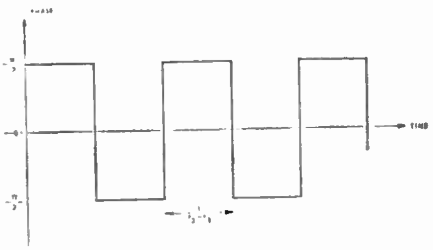

Fig. 3.

phase modulation relative to the $(F_1+F_2)/2$. Therefore, the wave is merely a sine wave which has a frequency equal to the arithmetic mean of F_1 and F_2.

Please refer to Fig. 4, which is a block diagram of one possible method for accomplishing this effect. The two equal amplitude tones are fed to a linear summation network which may consist merely of two resistors or two capacitors. The combined two equal amplitude tone wave is then fed to an amplitude limiter where the AM component of the wave is eliminated resulting in a pure phase modulated wave. The output of the limiter is fed to a phase inverter which provides two waves having opposite polarity. These two waves are then fed to a gate which either passes one polarity or the other, according to a control wave. The control wave may be derived from the two equal tone wave before the limiter by envelope detecting it with a diode detector. Whenever the amplitude of the two equal amplitude tone wave goes through zero, the gate switches polarity. This is exactly the proper time to reverse polarity in order to counteract the phase whip of the phase modulated wave.

It is important, when using this technique, to have the frequency of the two equal amplitude tones separated by a small percentage. As a rough rule, this should be $F_1 \geq 10(F_2-F_1)$. Otherwise, the envelope detector would be confused by the normal zero crossing of the RF wave and would not be able to distinguish between RF zero crossing and the envelope variations.

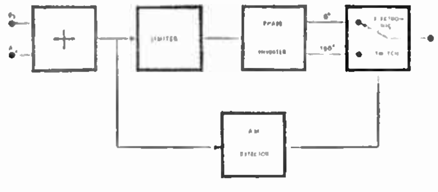

Fig. 4.

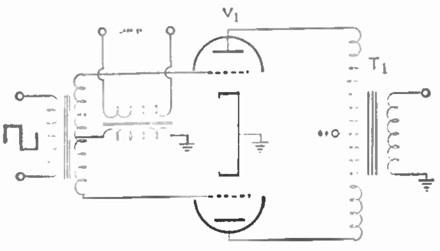

Fig. 5.

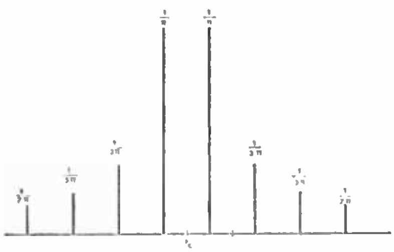

Fig. 6.

Analogy to a Double-Sideband Suppressed Carrier Wave

The correctness of this method of combining two frequencies may be demonstrated by use of the analogy of a square wave modulated double-sideband suppressed carrier wave. Fig. 5 shows a balanced modulator which may be used to produce this type wave.

Let us assume that a square wave is used to modulate the balanced modulator. Tube V1 produces most of the output when the square wave voltage is positive, and V2 produces most of the output when the square wave is negative. Since the plates of tubes V1 and V2 are connected to opposite ends of transformer T_1, the phase of the output RF wave reverses when the square wave input polarity reverses. If a perfect 50-50 square wave input is used, the positive and negative excursions of the square wave must be equal and, therefore, assuming perfect symmetry, the RF output amplitude will be constant.

The spectrum of a double-sideband suppressed carrier signal may be calculated by using the Fourier series representation of the input square wave and adding and subtracting each modulation component to the suppressed carrier frequency. (See Fig. 6.) It turns out that this symmetrical double-sideband suppressed carrier wave is identical to a limited two equal tone wave. This may be seen by comparing the square wave modulated DSBSC wave (Fig. 6) with a limited two equal tone wave shown in a prior publication.[3]

If the output of the balanced modulator is fed to a polarity reversing switch, which is keyed by the square wave, the effect of the balanced modulator will be nullified. Thus, the output of the system will be a pure carrier wave which will be exactly in the center of the double-sideband suppressed carrier wave. It is interesting to note that this technique may also be used to isolate the carrier wave for use in product demodulation of a double-sideband suppressed carrier wave.

Acknowledgment

I would like to acknowledge the helpful suggestions on this work of a number of my associates at Kahn Research Laboratories, Inc., including L. A. Ottenberg, Dr. L. O. Goldstone, H. L. Pull, and K. B. Boothe.

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[3] L. R. Kahn, "Comparison of linear single-sideband transmitters with envelope elimination and restoration single-sideband transmitters," Proc. IRE, vol. 44, pp. 1706–1712; December, 1956.

Temporary and Permanent Deterioration of Microwave Silicon Crystal Diodes*

Two recently declassified British reports[1,2] and an article in the Russian technical literature[3] have reported temporary changes in microwave diode performance under certain conditions of operation.

Quite independently we have accumulated many experimental data that complement these reported results. Temporary deterioration of microwave silicon diodes was observed by us in late 1957, during an investigation conducted at The Moore School of Electrical Engineering with the aim of establishing the burnout properties of crystal mixers exposed to RF pulses. In that investigation the performance of numerous microwave crystals subjected to X-band pulsed power of continuously increasing levels was studied by measuring the noise figure of a microwave crystal mixer followed by a very low noise IF amplifier. During these measurements it was noticed that

* Received by the IRE, April 24, 1959. The work reported here was done under Contract AF 30(602)-1615 with Rome Air Dev. Center, Griffiss Air Force Base, Rome, N. Y. and under Contract DA-36-039-sc-78055 with The U. S. Army Signal Supply Agcy., Ft. Monmouth, N. J.

[1] C. Ullyat, "Burnout Characteristics of CVX2154," TRE Memo. No. 556; May 29, 1952.

[2] W. P. Cole, S. Benjamin, and A. C. Leash, "Temporary Deterioration in the Noise Factor of X-band Crystal Mixers due to Diplexer Leakage Power," Rept. No. SL185, G. E. Ltd., Eng.; July 20, 1956.

[3] S. E. Temkin and K. M. Krolvets, "Temporary deterioration of the detecting properties of crystal diodes in high frequency operation," *Radiotekh. Elektrotekh. Z.*, vol. 2; pp. 1062–1070; 1957.

while the noise figure was not a sensitive parameter of the crystal deterioration process, the local oscillator (LO) rectified current was a very sensitive one. In other words, the LO rectified current showed that considerable changes occur in the crystal at RF power levels at which no deterioration is observed in noise figure. It was found that the deterioration of 3 db in noise figure corresponded to a relative decrease in LO rectified current of about 70 per cent.

Using LO rectified current as a parameter, hundreds of 1N23 *B* and *C*, and 1N21 B_a nd *C* crystals were tested under increasing pulsed *X*-band power for pulsewidths ranging from 5 mμsec to 1.8 μsec, and pulse repetition frequencies from 0.1 to 1000 cps. It was observed that, although considerable difference of behavior existed from sample to sample and between crystals of different types, all crystals followed a common pattern of deterioration of LO rectified current for continuously increasing RF pulse power. This pattern can be analyzed by considering four contiguous regions of deteriorating power, labeled *A*, *B*, *C*, and *D* in Fig. 1. In

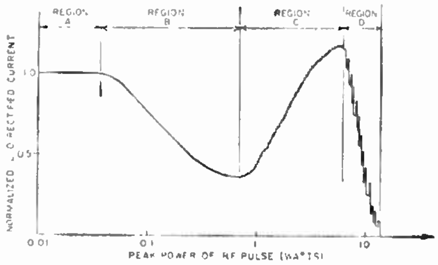

Fig. 1—Normalized rectified LO current vs peak RF power.

region *A* no deterioration occurs. In region *B* the rectified current decreases for increasing power. Discontinuing the application of pulse power, the current recovers to its initial value in a time of the order of minutes. In region *C* the rectified current slowly tends to regain its initial level. The recorded trace of the rectified current in this region exhibits barely detectable fluctuations. Region *D* is where permanent deterioration occurs. It starts with a sudden "jump" of current, followed by a succession of other jumps of various amplitudes, some downward, some upward. The trend is, however, definitely downward, and the process ends with the complete destruction of the rectifying properties of the crystal. Upon discontinuation of the pulse power in this region, the current remains at the level established by the last jump.

The temporary changes of rectified current associated with region *B* have been the object of special attention. It has been established that the *B* region begins when the voltage peak of the RF pulse reaches the reverse breakdown region of the characteristic of the crystal. The end of region *B* seems to be connected with a thermal process in the barrier.

In another type of test, the RF pulsed diode had applied to it a dc bias of either polarity, instead of a CW signal, and the resulting direct current was observed. Under forward bias the direct current remains essentially constant under increasing RF pulse power until the point where complete burnout occurs. Under reverse bias the current varies in analogy with the LO rectified current pattern; *i.e.*, in region *B* increases, in region *C* falls down, and in region *D* jumps upward and downward, but with an upward trend, eventually reaching the absolute value of the forward current.

Recently it has been found that temporary changes of both rectified and reverse current can be also obtained applying to silicon diodes dc pulses of short duration in the negative direction. There is sound reason to believe that the mechanism involved in the phenomenon is the same for RF or dc pulses, since an RF pulse can be considered as a long train of extremely short backward dc pulses, intermixed by an equal number of forward pulses which apparently have no major effect in the picture. When the pulse duration is short with respect to the thermal relaxation time of the barrier, the effect of many pulses can be accumulated to produce a large temporary change in the crystal characteristics without concomitant heating process.

All observations lead to the conclusion that the temporary deterioration is basically an ionization phenomenon taking place in the barrier and conditioned by the surface state of the silicon crystal wafer.

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Cathode-Ray Tube Triode Gun with Beam Former Electrode*

The absence of aberrations in the electron gun of a cathode-ray tube is important for obtaining high resolution on the phosphor screen. Astigmatism, one of the geometric aberrations commonly encountered, may result when an electron beam penetrates a lens region under a certain tilt. Such a beam tilt may be caused by eccentricities of apertures, tilted diaphragms, etc.[1] A gun has been developed which achieves a substantial reduction in astigmatic aberrations and thus promises to be less sensitive to mounting tolerances.

Such a gun is shown in Fig. 1. It consists of a cathode *C*, a modulating grid *G* with an aperture diameter 0.025 inch, and a bullet-shaped anode *A* with an aperture 0.080 inch. Both anode *A* and grid *G* are completely enclosed and shielded by an additional electrode,[2] the "beam former electrode" *B*. For flying-spot scanner CRT application, an external magnetic focusing unit of conventional design (not shown in Fig. 1) is employed together with magnetic deflection coils. No external centering device is required for proper operation.

While the performance of this gun has been evaluated in a 5-inch diameter flying-spot scanner CRT envelope (over-all length 14⅜ inches, distance yoke reference line to screen 4⅝ inches, neck length 9¾ inches), its application is not limited to this tube type. The screen of some of the tubes tested consisted of aluminized, high efficient yellow-green P-20 phosphor of small particle diameter (most probable particle size 2μ) to permit an accurate microscopic observation of the spot shape.

The tubes were operated with grid modulation at 30 kv anode voltage and +2.5 kv beam former potential. The grid voltage required for visual cutoff of the undeflected beam was −64 volts. The resolution was determined by raster compression. Aberrations were subjectively judged by microscopic observation of the "spot" (obtained by pulsing and deflecting the beam; pulse length, 0.1 μsec).

The *center resolution* in total number of TV lines (raster height 4½ inches) as a function of the *screen current* is depicted in Fig. 2. A

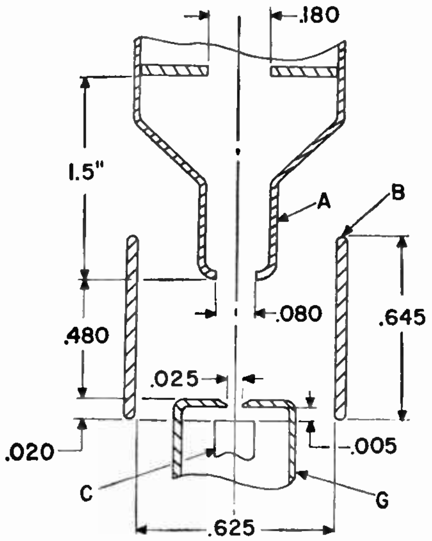

Fig. 1—Triode gun with beam former electrode.

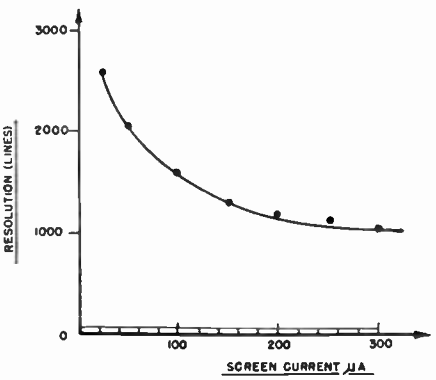

Fig. 2—Center resolution of a flying spot scanner CRT employing the beam former gun as function of the screen current.

* Received by the IRE, April 29, 1959.

[1] H. Moss, "The electron gun of the cathode-ray tube," *J. Brit. IRE*, vol. 6, pp. 99–124; June, 1946.

[2] A flying-spot scanner CRT employing a triode gun with an additional component has been described elsewhere, *i.e.*, A. Brill, J. deGier, and H. A. Klasens, "A cathode-ray tube for flying spot scanning," *Philips Tech. Rev.*, vol. 15/8–9, pp. 233–237; February/March, 1953. However, that additional component serves only as a spark trap and not as a gun electrode.

resolution of more than 1000 lines is obtainable at a screen current of 200 μa. For low screen current values, namely 15 μa or less, the resolution is in excess of 2500 lines.

The *current efficiency* of this gun or the ratio of screen current to total beam current has been determined to be higher than 85 per cent for screen current values below 500 μa.

Deflection distortions were judged by comparing edge and center resolution with the screen current as independent variable. The edge resolution averages approximately 30 per cent lower than the center resolution, and the ratio between edge and center resolution decreases slightly with increasing screen current.

Center resolution and *deflection distortions* depend strongly on the choice of the *beam former voltage* for any given anode potential. Both center resolution and deflection distortions decrease with decreasing beam former voltage. In this respect, the gun discussed here yields results similar to those achieved with a pentode gun described by Francken, deGier, and Nienhuis.[3]

All the guns were assembled by employing conventional jigging and beading methods. No special precautions, such as optical control methods, were used to achieve optimum alignment of gun components. Nevertheless, astigmatism could not be observed in any of the guns tested. While this result may be scientifically inconclusive, it seems to indicate that the structure chosen is less sensitive to the influence of unavoidable mounting tolerances than conventional guns, which, mounted in the same way, almost always show astigmatism.

The cutoff voltage of the beam former gun depends on both the anode potential and the beam former voltage. Thus, cutoff changes caused by anode voltage fluctuations are less pronounced than in the case of a simple triode gun. Further, the presence of the beam former electrode permits a relatively small cathode-grid spacing at high anode-potential and low cutoff voltage.

A triode gun employing a beam former electrode possesses a somewhat unusual current characteristic. The maximum obtainable beam current, $I_{\max}$, is expressed by $I_{\max} = K_E V_C^{-3/2}$, wherein V_C denotes the cutoff voltage and K_E the "effective perveance" or "drive factor." When the beam former voltage is increased, starting from ground potential, the effective perveance determined from the above relation, shows a relative maximum, as depicted in Fig. 3. The maximum effective perveance achievable with this gun at an anode potential of 27 kv is 4.3μAV$^{-3/2}$. The effective perveance of a conventional tetrode gun with cathode modulation is approximately 3.5μAV$^{-3/2}$, and the perveance of a so-called "low drive tube" is approximately 4.5 to 5.0μAV$^{-3/2}$.[4]

Summarizing, it may be stated that the triode gun with beam former electrode described here may represent an improvement in the performance quality of a CRT gun

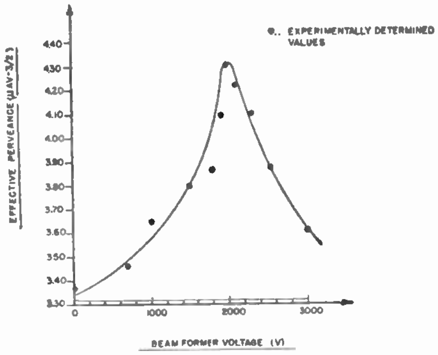

Fig. 3—Effective perveance as function of the beam former voltage.

without employing either a complicated mechanical construction or unconventional mounting methods and/or correcting means.

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[3] J. C. Francken, J. deGier, and W. F. Nienhuis, "A pentode gun for television picture tubes," *Philips Tech. Rev.*, vol. 18/3, pp. 73–81; September, 1956.

[4] W. F. Niklas, C. S. Szegho, and J. Wimpffen, "A television picture tube with increased effective perveance for cathode modulation," *J. Television Soc.*, vol. 8/9, pp. 1–8; April, 1958.

A Note on the Stability of Linear, Nonreciprocal n-Ports*

An n-port N is said to be stable if the port currents are zero under all passive 1-port terminations and weakly stable if the port currents are zero under all 1-port terminations whose real parts are simultaneously greater than zero.[1] Recently, the author succeeded in proving that a linear reciprocal n-port is weakly stable if and only if it is passive, thus generalizing Llewellyn's well-known result for $n=2$.[1,2] A compact set of stability criteria for nonreciprocal n-ports is still not available and the object of this paper is to make a start on this problem.

Let A be an arbitrary matrix. Then A', $\bar{A}$, A^* and det A stand for the transpose, the complex conjugate, the complex conjugate transpose, and the determinant of A, respectively. Column vectors are represented by $\boldsymbol{a}$, $\boldsymbol{b}$, etc. For a hermitian matrix $A = A^*$, $A \geq 0$ means that A is the matrix of a nonnegative quadratic form. A diagonal matrix A with diagonal elements $\mu_1, \mu_2, \cdots, \mu_n$, is written as $A = \text{diag}\,[\mu_1, \mu_2, \cdots, \mu_n]$.

Theorem:[3] Let $Z_1(j\omega)$ and $Z_2(j\omega)$ be the impedance matrices of two n-ports N_1 and N_2. Then if Z_1 and Z_2 possess identical principal minors of all orders, N_1 and N_2 are stable (weakly stable) or unstable (weakly unstable) together.

Proof: An n-port N with impedance matrix $Z(j\omega)$ is stable if and only if the linear system

$$(Z + Z_0)I = 0, \tag{1}$$

$$Z_0 = \text{diag}\,[z_1, z_2, \cdots, z_n], \tag{2}$$

possesses only the trivial solution $I = 0$ for every choice of n complex numbers $z_1, z_2, \cdots, z_n$ with nonnegative real parts. Thus N is stable if and only if

$$\det\,(Z + Z_0) \neq 0 \tag{3}$$

for any choice of n z's with nonnegative real parts. Now det $(Z+Z_0)$ can be expanded in terms of the elements of Z_0 as follows:

$$\det\,(Z + Z_0) = \det Z + \sum_{k=1}^{n} z_k B_k + \sum_{k<r}^{n} z_k z_r B_{k,r} + \sum_{k<r<l}^{n} z_k z_r z_l B_{k,r,l} + \cdots + z_1 z_2 \cdots z_n, \tag{4}$$

where B_k is the principal minor of Z obtained by striking out the kth row and column, $B_{k,r}$ is the principal minor obtained by deleting the kth and rth rows and the kth and rth columns, etc. Clearly, if Z_1 and Z_2 have identical principal minors of all orders,

$$\det\,(Z_1 + Z_0) = \det\,(Z_2 + Z_0) \tag{5}$$

and the vanishing or nonvanishing of either side of (5) implies that of the other. Hence N_1 and N_2 are stable or unstable together, Q.E.D.

To see how this theorem works consider a 2-port N with impedance matrix

$$Z = \begin{pmatrix} z_{11} & z_{12} \\ z_{21} & z_{22} \end{pmatrix}.$$

It is obvious by inspection that Z and

$$Z_s = \begin{pmatrix} z_{11} & \sqrt{z_{12}z_{21}} \\ \sqrt{z_{12}z_{21}} & z_{22} \end{pmatrix} \tag{6}$$

have the same principal minors of orders one and two. By the above theorem, N is stable if and only if the 2-port represented by Z_s is stable. But this latter 2-port being reciprocal (Z_s is symmetric) is stable (weakly stable) if and only if

$$\text{Re}\, Z_s = \text{Re}\begin{pmatrix} z_{11} & \sqrt{z_{12}z_{21}} \\ \sqrt{z_{12}z_{21}} & z_{22} \end{pmatrix} > 0(\geq 0). \tag{7}$$

In other words, N is stable (weakly stable) if and only if Re $Z_s > 0(\geq 0)$. As far as the author can determine the very succinct formulation (7) for the stability of a general 2-port has not been previously reported in the literature. A set of conditions equivalent to (7) is

$$\begin{aligned} \text{Re}\, z_{11} &= r_{11} > 0(\geq 0), \\ \text{Re}\, z_{22} &= r_{22} > 0(\geq 0), \text{ and} \\ |\delta| &+ \text{Re}\,\delta < 2(\leq 2), \end{aligned} \tag{7}$$

where

$$\delta = \frac{z_{12}z_{21}}{r_{11}r_{22}} \tag{8}$$

and are the conditions usually quoted.[4] The inequalities in (7) merely express the fact that the three principal minors of Z_s must be positive (nonnegative).

Unfortunately it is not possible, in general, to create a symmetric $n \times n$ matrix Z_s whose principal minors of all orders are

* Received by the IRE, May 1, 1959. This work was sponsored by the Air Force Cambridge Res. Center under Contract AF-19(604)-4143.

[1] D. Youla, "A stability characterization of the reciprocal linear passive N port," PROC. IRE, vol. 47, pp. 1150–1151; June, 1959.

[2] F. B. Llewellyn, "Some fundamental properties of transmission systems," PROC. IRE, vol. 40, pp. 271–283; March, 1952.

[3] Due to Dr. H. Kurss of the Microwave Research Institute.

[4] E. F. Bolinder, "Survey of some properties of linear networks," IRE TRANS. ON CIRCUIT THEORY, vol. CT-4, pp. 70–78; September, 1957.

identical with those of a prescribed nonsymmetric $n\times n$ matrix Z. A little thought should convince the reader that the element in the rth row and kth column of Z_s must have the form

$$(Z_s)_{r,k} = z_{r,r}, \qquad r = k,$$
$$= \epsilon_{r,k}\sqrt{z_{r,k}z_{k,r}}, \qquad r \neq k, \qquad (9)$$

where $\epsilon_{r,k}=\epsilon_{k,r}=\pm 1$. Thus the principal minors of Z_s of orders greater than two will not necessarily equal the corresponding ones of Z so that the theorem is inapplicable.

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High-Power Effects in Ferrite Devices*

It has been known for a long time that the performance of ferrite devices deteriorates at power levels much lower than one would expect from analogy with saturation effects in paramagnetic resonance. Suhl[1] has shown that these premature saturation effects are due to the unstable growth of spinwaves which extract energy from the uniform procession. The Suhl threshold for the onset of the decrease in susceptibility under conditions of resonance is given by

$$h_c = \Delta H \sqrt{\frac{\Delta H_k}{4\pi M_s}}. \qquad (1)$$

Here h_c is the RF field across the sample, ΔH is the linewidth of the sample, $4\pi M_s$ is the saturation magnetization of the sample, and ΔH_k is the linewidth of the spinwave that goes unstable. In applying this relation it has been common to assume that $\Delta H=\Delta H_k$ in the absence of any other knowledge of the relation between these two quantities. However, this assumption has always given power levels that are too high. Other assumptions have been made that perhaps the "intrinsic linewidth" of the material should be used, and that this linewidth is smaller by a factor of anywhere from 5 to 20 than the actual measured linewidth of the sample. Through artifices such as these it has been possible to obtain good agreement between experiment and theory.

We have measured the profile linewidth and Suhl threshold for a number of samples of ferrite[2] and have calculated the spinwave linewidth from (1). Representative values are given in Table I, and it is seen that ΔH_k

TABLE I
FERRITE PROFILE AND SPINWAVE LINEWIDTHS

Sample	ΔH (oersteds)	ΔH_k (oersteds)
Yttrium Iron Garnet	46	1.2
R-1 (Mg-Mn ferrite)	463	0.55
Mn ferrite	310	0.30
Mn single crystal	37	3.9

falls[3] between 0.3 and 4 oersteds for all samples. If we consider a sample of R-1 and calculate the critical field from Suhl's relation, we find that h_c should equal 204 oersteds. This is under the assumption that $\Delta H=\Delta H_k$ and ΔH is given by the observed linewidth on a polished sphere of the material. If, however, we use the value given in Table I, we find that the critical field is 7 oersteds. This is a factor of 29, or a factor of 840 in power. If we use the expression for the fields in ordinary X-band waveguide, the threshold for Suhl instability will be 5 megw according to the assumption $\Delta H=\Delta H_k$. If one uses the measured values, then the threshold for Suhl instability is reduced to 6 kw.

Some specific examples may now be considered. Schulz-Dubois, Wheeler, and Sirvetz[4] have made some measurements of a high-power L-band resonance isolator. One version of the isolator used a nickel cobalt ferrite with the saturation magnetization 3000 gauss, and a linewidth of 200 oersteds. In calculating the threshold power level they assumed a so-called "intrinsic linewidth" equal to $\frac{1}{5}$ of the measured linewidth. They then calculated that nonlinearity should set in at about 1 megw. If we assume a linewidth of 1 oersted for ΔH_k and use the linewidth given for ΔH we find that the instability should occur at about 600 kw. Measurements were performed by Schulz-Dubois, Wheeler, and Sirvetz at powers in the range of 300 to 500 kw, and they found that the reverse attenuation was substantially independent of power in this range. A second version was made with a nickel aluminate with a $4\pi M_s$ of 300 gauss, and a linewidth of 825 oersteds. The Suhl power threshold of this isolator is then 10 megw. Measurements were performed up to 2 megw and no deterioration in performance was noted.

It should be pointed out that in the case of the main resonance the decline in susceptibility after Suhl threshold is very slow, that is, the threshold is not very sharp. For example, from the onset of Suhl instability one must increase the power by 8 db to get to the point at which the susceptibility is $\frac{1}{2}$ its small-signal value. Therefore, it is conceivable that in devices one may use powers of up to perhaps 4 times that required for Suhl threshold without serious decline of device performance.

Soohoo[5] has made some calculations of threshold fields in power limiters, and has compared these calculations with experiments on a device operating at subsidiary resonance with a sample whose saturation magnetization is 1800 gauss and whose linewidth is 600 oersteds. With the assumption $\Delta H=\Delta H_k$, Soohoo calculates a critical power of 30.6 megw from the Suhl[1] threshold equation for subsidiary resonance,

$$h_c = \frac{\Delta H_k}{4\pi M_s} g(\theta)\sqrt{(H-H_0)^2-(\Delta H)^2} \qquad (2)$$

where H is the dc magnetic field, H_0 is the resonant dc magnetic field, and $g(\theta)$ is a function of order one at the minimum of h_c.

He finds experimentally that the threshold power level is only about 300 watts. If we now recalculate from (2) using $\Delta H_k=1$ oersted, we reduce h_c by a factor of 600 and the critical power by $(600)^2$, resulting in a threshold of 100 watts.

All the above calculations can only be considered valid to an order of magnitude, since when a ferrite is placed in a waveguide the modes in field patterns are distorted, and, therefore, the RF field across the ferrite is not that which the sample would experience in an empty waveguide. These calculations have used empty waveguide expressions. In addition, this last calculation is not strictly correct, since the ΔH_k's have been measured at the main resonance and the power limiter operates at the subsidiary resonance. However, some measurements of Spencer, LeCraw, and Porter[6] of ΔH_k under conditions of main resonance degenerate with subsidiary resonance (the spinwaves excited here are similar in character to those excited for the subsidiary resonance[1]) shows that ΔH_k is of the same order of magnitude as the main resonance and is independent of inhomogeneity broadening.

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[1] H. Suhl, "The nonlinear behavior of ferrites at high microwave signal levels," PROC. IRE, vol. 44, pp. 1270–1284; October, 1956. Also, "Theory of ferromagnetic resonance at high signal powers," *J. Phys. Chem. Solids*, vol. 1, pp. 209–227; April, 1957.

[2] The manganese ferrite and yttrium iron garnet samples were supplied by the Solid State Physics and Tech. Ceramics Secs. of Lockheed Missiles and Space Division. The manganese single crystal was supplied by Linde Air Products.

[3] ΔH_k is calculated, from (1), where, for the present purpose, we take h_c from the first observable decline of susceptibility.

[4] E. O. Schulz-Dubois, G. J. Wheeler, and M. H. Sirvetz, "Development of a high-power L-band resonance isolator," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-6, pp. 423–428; October, 1958.

[5] R. F. Soohoo, "Power limiting using ferrites," 1958 IRE NATIONAL CONVENTION RECORD, pt. 1, pp. 36–44.

[6] E. G. Spencer, R. C. LeCraw, and C. S. Porter, "Microwave Properties of Yttrium Iron Garnet," Diamond Ord. Fuze Labs., Washington, D. C., Tech. Rept. TR-633; June 30, 1958.

Method and Means of Rotating Large Satellites*

Document No. 1354 in the files of my patent attorney, dated January 9, 1958, herewith disclosed in lieu of a Lunar Patent:

It is the purpose of my invention to make large satellites in interstellar space

* Received by the IRE, June 1, 1959.

rotate with any desired speed and around any desired axis through their center of gravity. To this end, I propose to anchor to the surface of such a satellite a rocket or other source of directed blast, preferably on or near the intended equator, and with the direction of its blast opposite to and parallel to the tangent of the intended rotation, and to let it act with the power and for the duration corresponding to the rotational moment I of inertia of the satellite and of the intended speed of rotation.

For instance, I may desire to advance the gaiety of nations by making the (old) moon of the earth rotate around a suitably chosen axis through its center and at a pleasing speed. This should relieve the monotony of the presently unvarying aspect and permit everybody to see what is on the side now averted.

The mass of the moon is known to be $m=7.4\times10^{25}$ grams, and its radius (if it is a perfect sphere) $r=1.74\times10^{8}$ cm; hence, its moment of inertia (around any axis) $I=9\times10^{41}$ grams cm², and the power and duration of the tangential blast is to be computed accordingly.

The moon may be considered a sphere for the computation of its moment of inertia; but it is perhaps not a perfect sphere. In that case, with the distance of its center of gravity finite and given, it will tend for minimum potential energy to turn one particular aspect towards the earth. A modest rotational disturbance will then merely make the moon oscillate around this preferred orientation. Further, since the moon is not to be considered perfectly rigid, soft or sliding parts will suffer friction each time the oscillatory acceleration reverses, and this will finally damp the oscillation. In order to achieve sustained rotation, this tangential push must be strong enough to turn the moon at least 180°, *i.e.*, to the orientation opposite to the preferred one, and to make it turn through this orientation with finite speed.

If it is inconvenient to meet this requirement with a single blast, I may set off a sequence of such blasts spaced to aid one another, as nearly as practicable, by an integral multiple of the period of oscillation around the preferred orientation, thus increasing each time the amplitude of oscillation, at least until the amplitude exceeds the required minimum of 180°.

Thus far document No. 1354. Comments:

1) I have considered the case of a satellite that is not a perfect sphere, for instance shaped like an egg, by assuming, in familiar manner, two unequal point masses connected by a rigid weightless rod, at large but finite distance from the earth. The forces of mutual attraction then have two maxima, both when the axis through both masses points to the earth. Of these, the maximum with the smaller mass nearer the earth is a little higher. One must assume that the moon now has one of these attitudes. The larger the push needed to turn the moon past the "dead point" after a quarter turn, the faster would be the slowest possible sustained rotation.

2) I have been reminded that the value given above for the moment of inertia I is unrealistic since it assumes a sphere of homogenous density ρ. More probably, the density of the moon—as that of the earth—is greatest at the center and falls off towards the surface. I have computed a simple model of a sphere of radius r_0 consisting of a spherical core of radius $0.2r_0$ and of four concentric shells each $0.2r_0$ thick, and with a density $2\rho_0$ at the core falling in steps of $0.25\rho_0$ to ρ_0 at the last shell. Such a model has the same mass as a sphere of radius r_0 and of homogenous density $1.2\,\rho_0$. Comparing their moments of inertia, I find that of the graded model 8 per cent smaller. The difference is small since $I\propto r^5$, thus a heavy core contributes little.

3) Assuming again the true homogenous sphere with $I=9\times10^{41}$ grams cm², we may estimate the push needed to make it turn, let us say once in five years, thus $\omega\approx4\times10^{-8}$ rad/sec. Also arbitrarily, we choose a push of constant force lasting 10^6 seconds (*i.e.*, some 11 days). About 2×10^{20} dynes are needed. (For comparison, to accelerate a car of 1500 kg in 10 seconds from stop to 120 km/hr takes 5×10^{8} dynes.) It would be wiser to apply two equal tangential forces on opposite points of the equator to avoid all but rotational disturbances.

4) The first step would be to apply a small known tangential push and to telemeter the resulting minute oscillation. Accurate knowledge of its amplitude and period would permit more reliable computations.

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The Computation of Single-Sideband Peak Power*

Comparisons of amplitude-modulation techniques such as AM, DSB, and SSB are most frequently made on the basis of sinusoidal modulation; however, the modulating signals of the most interest are far from sinusoidal and often are nearly rectangular waveforms such as encountered with clipped speech and pulse-coded data. One of the important parameters of these techniques is the transmitter average-to-peak power ratio as a function of the modulating waveform. While the computation of this parameter is moderately straightforward for AM and DSB, the calculation of SSB peak power is somewhat more involved. This paper describes an approach which permits simple and direct computation of the average-to-peak SSB power for a continuous range of modulating signals from sinusoidal to square waveforms.

The modulating signal will be represented by the periodic function $\sin^\nu x$, $0\leq x\leq\pi$, and $-\sin^\nu x$, $\pi\leq x\leq2\pi$, where x is a linear function of time and ν lies between zero and unity. When plotted over the interval $(0,\pi)$, this appears as shown in Fig. 1 for various values of ν; it can be seen that when $\nu=1$, the waveform is sinusoidal and as $\nu\to0$ the waveform becomes square.

* Received by the IRE, May 22, 1959.

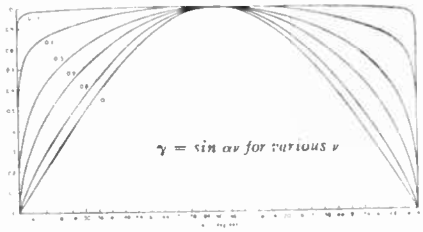

Fig. 1—Sin να from 0 to 180°.

The SSB *average* power is simply equal to the modulating signal average power. Since the SSB modulation process shifts the signal in the frequency domain without altering its total energy content, the average power must be identical before and after modulation. (The power gain of the modulator or transmitter is assumed here to be unity, for convenience.) The average power of the modulating signal is[1]

$$P_{\text{ave}}=\frac{1}{\pi}\int_0^\pi\sin^{2\nu}x\,dx=\frac{1}{\sqrt{\pi}}\,\frac{\Gamma(\nu+\frac{1}{2})}{\Gamma(\nu+1)}\qquad 0<\nu\leqq1. \quad (1)$$

The SSB *peak* power is computed as follows. Before modulation, the peak power of the modulating signal is unity for all values of ν $(0<\nu\leq1)$, the harmonic frequencies composing the modulating signal are commensurable, and all add in such phase as to produce the unity peak amplitude. The modulation process shifts all the harmonics in the frequency domain by an arbitrary amount equal to the (virtual) carrier frequency. The frequencies are now incommensurable and, with probability unity, at some time all add "in phase." Therefore, if the modulating signal is written as a Fourier series expansion, the peak *amplitude* after modulation will be equal to the summation of the Fourier coefficients, or

$$[P_{\text{SSB}}]^{1/2}=A=\sum_{n=1}^{\infty}|a_n(\nu)| \quad (2a)$$

where, in the interval $(0,\pi)$,

$$a_n=\frac{2}{\pi}\int_0^\pi\sin^\nu x\sin nx\,dx. \quad (2b)$$

If the integration and summation are carried out in the order indicated in (2), an unwieldy expression results which is unsatisfactory for computation since it is not obvious that the resulting infinite series can either be expressed in closed form or that it converges rapidly enough for efficient numerical computation. Instead, the order of summation and integration are interchanged, yielding the result directly in closed form. Since, from symmetry, the even terms vanish and the odd terms are positive, then

$$A=\sum_{n=1}^{\infty}|a_n|=\sum_{n=1}^{\infty}a_{2n-1}$$
$$=\sum_{n=1}^{\infty}\frac{2}{\pi}\int_0^\pi\sin^\nu x\sin(2n-1)x\,dx \quad (3)$$

which can be rewritten as

[1] W. Gröbner and N. Hofreiter, "Integraltafel, Zweiter Teil, Bestimmte Integrale," Springer-Verlag, Vienna, Austria, no. 332.9a, p. 108; 1958.

$$A = \lim_{N\to\infty} \sum_{n=1}^{N} \frac{2}{\pi}\int_0^\pi \sin^\nu x \sin(2n-1)\,x dx$$
$$= \lim_{N\to\infty} \frac{2}{\pi}\int_0^\pi \sin^\nu x \sum_{n=1}^{N} \sin(2n-1)\,x dx \quad (4)$$

since the interchange is justified for a finite series. Now,[2]

$$\sin\alpha + \sin(\alpha+\delta) + \sin(\alpha+2\delta) \cdots + \sin[\alpha+(n-1)\delta] = \frac{\sin\left(\alpha + \frac{n-1}{2}\delta\right)\sin\frac{n\delta}{2}}{\sin\frac{\delta}{2}}.$$

Letting $\delta = 2\alpha$,

$$\sum_{n=1}^{N}\sin(2n-1)\alpha = \frac{\sin^2 N\alpha}{\sin\alpha} = \frac{1-\cos 2N\alpha}{2\sin\alpha}$$

which when substituted in (4) gives

$$A = \frac{1}{\pi}\int_0^\pi \sin^{\nu-1} x dx - \frac{1}{\pi}\lim_{N\to\infty}\int_0^\pi \sin^{\nu-1} x \cos 2Nx dx. \quad (5)$$

According to the Riemann-Lebesque theorem,[3] if $f(x)$ is integrable over (a, b), then as $\lambda \to \infty$

$$\int_a^b f(x)\cos\lambda x dx \to 0.$$

Since $\sin^{\nu-1} x$ is integrable in $(0, \pi)$ for $0<\nu\leq 1$, the right-hand integral in (5) vanishes in the limit. The left-hand integral is evaluated,[4] giving

$$A = \frac{1}{\sqrt{\pi}}\frac{\Gamma\left(\frac{\nu}{2}\right)}{\Gamma\left(\frac{\nu+1}{2}\right)} \qquad 0 < \nu \leq 1. \quad (6)$$

When P_{ave} is divided by A^2, the SSB average-to-peak power ratio as a function of ν is obtained. This is plotted in Fig. 2, with the corresponding ratios for AM and DSB. The curves can be read in PEP by inserting a factor of two, if desired.

It can be seen that for small values of ν (waveforms approaching square) the SSB ratio compares poorly with either AM or DSB; in fact, the SSB ratio can be approximated in this region by ν^2. Clearly, with perfectly square waves the SSB average power output is zero for any finite peak-power limitation. For applications involving even moderately rectangular waveforms, it can be concluded that DSB is superior to AM, which is considerably better than SSB. Although bandwidth limitations prevent attainment of perfectly square waves, it would be expected that SSB would exhibit poor effective signal-to-noise ratios when modulated by clipped speech or pulse-coded data, as compared to alternative techniques.

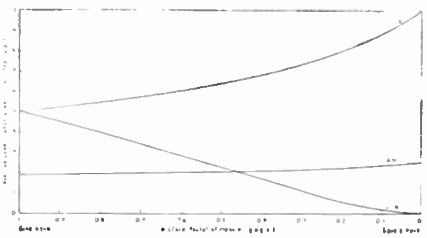

Fig. 2—Average-to-peak power relations as a function of modulating signal with no output bandwidth limitations.

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[2] H. Dwight, "Tables of Integrals," The Macmillan Co., New York, N. Y., rev. ed., no. 420.3, p. 86; 1947.
[3] E. Titchmarsh, "The Theory of Functions," Oxford University Press, New York, N. Y., 2nd ed., p. 403; 1939.
[4] Gröbner and Hofreiter, *op. cit.*, no. 331.28d, p. 97.

Limitations of Microwave Computer*

The trend toward higher and higher computing speeds on digital computers[1] has often been compared with the successful trend toward higher frequencies for broadcast, communication, radar, and microwave purposes. While the desire for higher speeds is understandable, I believe there are basic differences in the development of the two fields which do not favor a continued increase in clock frequency for digital computers.

Digital computers have had their great upswing by the coding of information into the energy of the electromagnetic field around the conductor and by the ease and speed with which this energy can be channeled into the proper place and cut off from the unwanted place. As the frequencies are increased beyond the present commonly used frequencies, the electromagnetic energy around the conductor is more and more represented by a radiating field, rather than one that "clings" to the conductor. While this phenomena is desirable for communications, it is diametrically opposed to the very idea of a digital computer, namely, the ability to channel detectable energy (carrying the information) to a specific point and *not* to a point immediately adjacent to it.

The latter features can be safeguarded only to a limited extent as the frequencies are increased by more complex construction and by an increase in distances between adjacent information carrying "lines."

The same purpose of fast computation can also be achieved, in general, by more conventional means, by paralleling in connection with miniaturization. The majority of information-handling problems are not of an entirely serial nature. A high computing speed in complex systems can almost always be achieved by doing several computations simultaneously, provided, naturally, that the size, cost, and complexity remain commensurate.

Whereas I do not see any limitations as yet to the possible miniaturization of computing circuitry, I do see basic limitations to very high clock frequencies. It appears to me that the former method will meet with more success in the foreseeable future, or, specifically as long as we will be using electrical energy to carry information.

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* Received by the IRE, June 1, 1959.
[1] R. L. Wigington, "A new concept in computing," Proc. IRE, vol. 47, pp. 516–523; April, 1959.

Passive Duplexer Design as a Function of Noise Temperature*

In a previous paper,[1] the writer gave criteria for optimum design of a directional coupler or bridge for passive duplexing service (Fig. 1). This showed that the traditional 6 db transmission-reception loss need not apply on a noisy channel.

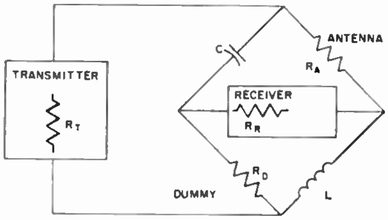

Fig. 1—Model passive duplexer used in the derivation.

Additionally considering the use of a refrigerated dummy resistance and analyzing signal degradation with the variables of dummy noise temperature and antenna combined noise plus interference temperature, the criteria for optimum design are:

1) $R_T = R_R = R_A = R_D = R$ (assumption)
2) $L = CR^2$ (consequent balance condition)
3) $\omega^2 LC(T_A/T_D)^{1/2} = 1$ (derived optimum), where
4) T_A = Antenna noise + interference temperature, °K
5) T_D = Dummy noise temperature, °K.

Under the stated conditions, the transmitter power must be multiplied by the factor $[1+(T_D/T_A)^{1/2}]^2$ to give equal signal/noise performance compared to a system with an antenna ideally switched between transmitter and receiver.

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* Received by the IRE, June 5, 1959.
[1] D. T. Geiser, "Degradation of Signal-to-Noise Ratio when Using a Directional Coupler for Duplexing," presented at Tenth Annual Southwestern IRE Convention; April 10–12, 1958.

Contributors

Charles R. Cahn (S'51–A'52–M'57) was born on October 7, 1929, in Syracuse, N. Y. He received the B.E.E. degree in 1949, the M.E.E. degree in 1951, and the Ph.D. degree in electrical engineering in 1955, all from Syracuse University.


C. R. Cahn

From 1949 to 1956, he served as instructor and later as assistant professor in the Electrical Engineering Department of Syracuse University, where he engaged in research work in information theory and microwave antennas and in studies on systems engineering. From 1952 to 1953, on a leave of absence, he was with the System Planning Department of the Niagara Mohawk Power Corporation, Buffalo, N. Y., where he was concerned with system planning and economic operation of a large integrated power system.

At present, he is a member of the senior staff of Ramo-Wooldridge, a division of Thompson Ramo Wooldridge Inc., Los Angeles, Calif., where he is working on systems analysis and synthesis, with emphasis on applications of information theory in the field of digital communications. He has also investigated techniques for electronic countermeasures and methods of achieving reliable transmission over fluctuating circuits.

He is a member of Sigma Xi and AIEE.

❖

David B. Geselowitz (S'51–A'54) was born in Philadelphia, Pa., on May 18, 1930. He attended the University of Pennsylvania, Philadelphia, Pa., from which he received the B.S. degree in electrical engineering in 1951, the M.S. degree in 1954, and the Ph.D. degree in 1958.


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Since 1951, he has been with The Moore School of Electrical Engineering of the University of Pennsylvania, working in the fields of radio noise measurement, shielding studies, operations research, and electrocardiography. As an associate, he teaches undergraduate courses in electrical engineering.

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❖

J. W. Herbstreit (A'40–SM'45–F'58) was born in Cincinnati, Ohio, on September 7, 1917. He received the M.S.E.E. in 1939 from the University of Cincinnati. In 1939 and 1940, he was associated with the Crosley Corporation in Cincinnati, as transmitter engineer. He then joined the FCC National Defense Operations in Marietta, Ga., later acting as radio inspector at the Atlanta office. In 1941, he was transferred to the FCC Engineering Department in Washington, D. C., as radio engineer in the Safety and Special Services Division.


J. W. Herbstreit

The following year, he joined the Operational Research Staff in the Office of the Chief Signal Officer, Dept. of the Army. While with this group, he made numerous operational radio systems studies, including measurements of atmospheric noise levels and the attenuation of radio signals by jungles in Panama and the southwest Pacific, measurements and analyses of experimental low-frequency loran in the Western Hemisphere, and frequency requirements for low-power radio communications and navigation equipment.

In 1946, Mr. Herbstreit joined the Central Radio Propagation Laboratory of the National Bureau of Standards, conducting research on cosmic radio noise and VHF and UHF propagation. He was responsible for the preparation of radio propagation when serving as technical advisor to the International High Frequency Broadcast Conference, in Mexico City in 1948, and in Florence and Rapallo, Italy in 1950. Since 1949, he has been in charge of the tropospheric propagation research work being conducted at the NBS, now centered at Boulder, Colo. He was a delegate to the VIIIth and IXth Plenary Assemblies of the CCIR, held in Warsaw, in 1956, and in Los Angeles in 1959. He received the 1959 Harry Diamond Award.

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❖

C. T. Kohn was born in Ostrzeszow, Poland, in 1908. He received the Dipl.-Ing. degree in electrical engineering in 1932 from the Institute of Technology in Lwow, Poland.


C. T. Kohn

From 1934 to 1939 he was employed by the National Establishment for Tele- and Radio-communications in Warsaw, Poland. After World War II, he was associated with the Signals Research and Development Establishment in Christchurch, England.

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J. N. Pierce

Since 1952, he has been employed at the U. S. Air Force Cambridge Research Center, Bedford, Mass., where he has worked on ultra-high frequency data transmissions systems, long-range anti-jamming systems, and high-reliability scatter transmission.

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❖

Seymour Stein (SM'57) was born in Brooklyn, N. Y., in April, 1928. He received the B.E.E. from the College of the City of New York, N. Y., in 1949, and the M.S. 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 1953–1956, he was an engineering specialist with the Waltham Laboratories of Sylvania Electric Products, Inc. From 1956 to September, 1959, he was with the technical staff at Hermes Electronics Company (formerly Hycon Eastern, Inc.), where he worked on a variety of problems in antennas, propagation, and communications system theory and was group leader of the Antenna Systems Group. He has just rejoined Sylvania as head of the Communications Research Department within the Applied Research Laboratory at Waltham, Mass.

Dr. Stein is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

Books

Circuit Theory of Linear Noisy Networks, by Hermann A. Haus and Richard B. Adler

Published (1959) by the Technology Press, MIT, and John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 75 pages+3 index pages+xii pages. Illus. 6×9. $4.50.

In this little book the authors have given a very generalized discussion of the spot noise figure of cascade-connected linear noisy networks and their optimization. This leads first of all to a full discussion of the noise performance of an n-terminal pair noisy network, in which stage variables other than the source impedance play a part and in which stage interconnections other than the simple cascade are important. This is, for example, the case if feedback from output to input is applied to the amplifier stage, if feedback is applied from one of the other terminal pairs, or if signals are received through different amplifier channels.

The noise of such a stage can be represented by a noise power density matrix. The discussion then leads to an investigation of the invariants of the system under lossless transformations. The search for a noise criterion that is independent of the gain of the stage leads to the concept of *noise measure* and to its optimization by means of lossless transformers. By putting the discussion in matrix notation and by associating it with the general circuit theory of active networks, the authors have been able to wrap the solution of the problem into a nice package, well suited for presentation in monograph form.

Some of the results obtained in this booklet are not fully new. The earlier results were derived for particular cases so that the question of how general the results were remained. It is the merit of this monograph that this earlier work is now augmented and that it is presented in such a generalized form. By doing this, the authors have indebted all those who are interested in noise problems in general, and those interested in minimum noise figure design in particular.

The authors make full use of matrix algebra, but do not explain their notation. Some readers will find the reading of several chapters rather difficult for that reason. Their difficulties could have been eliminated to a considerable extent by adding a short appendix on the matrix notation and the matrix algebra theorems used in this presentation.

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Recent Books

Fano, U. and L. Fano, *Basic Physics of Atoms and Molecules*. John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $10.00.

Jacobowitz, H., *Fundamentals of Nuclear Energy and Power Reactors*. John F. Rider Publisher, Inc., 116 W 14 St., N. Y. 11, N. Y. $3.95.

Kingery, W. D., *Property Measurements at High Temperatures*. John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $16.50. Concentrates on factors affecting the properties of materials and their measurement at temperatures above 1400°C.

Kullbach, Solomon, *Information Theory and Statistics*. John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $12.50.

Scanning the Transactions

The application of microwave techniques to computers is bringing together heretofore strange bedfellows in one of the most interesting inter-field unions to be found in electronics. For one thing, it represents the first time that microwaves have found important engineering use outside the radar and communications field. But more important, it offers a method of greatly increasing the operating speed of computers. The speed of present computer circuits is fast approaching the maximum theoretical limit which is imposed by the inductances and capacitances associated with conventional "lumped-constant" components. If we want faster circuits we must either develop lumped-constant components that are much smaller in size or else turn to distributed-constant techniques. Cryogenic computer components offer one possibility for extremely small-size circuit elements. Microwave techniques, on the other hand, offer the possibility of the distributed-constant approach. Because of the great interest that has recently evolved in microwave computers, the Office of Naval Research sponsored a symposium of invited papers on the subject last March. Six of these papers have now been published by the IRE TRANSACTIONS ON ELECTRONIC COMPUTERS. They describe systems in which binary information is represented not only by pulses (RF as well as dc) but also by the phase of an RF wave. The latter is a relatively new method which employs parametric subharmonic oscillators what will lock into either of two phases 180° apart, so that one phase represents the binary digit "one" and the other, "zero." As a result of this recent flurry of activity, computer engineers are now finding themselves very much involved with microwave diodes, parametric amplifiers and oscillators, traveling-wave tubes and strip-line printed circuits. They are also finding themselves concerned with time intervals they have never had to worry about before. In dealing with frequencies in the microwave range, the loss of as little as a millimicrosecond in switching a device from one binary state to the other can mean a two to one reduction in the rate at which information can be processed in the computer. And if the information has to be sent to another part of the computer that is a foot away, this will take another millimicrosecond and cause a further substantial reduction in computer speed. This latter limitation of distance makes the question of size of components still an important one, even in microwave computers. Although there are many practical difficulties to be overcome, the results to date, even at this early stage of development, are most impressive. The fastest information rate that has been attained with a conventional computer is about 50 megacycles, whereas rates of over 500 mc have already been achieved with experimental microwave models, and 1000 mc seems to be certain in due time. (Papers from Symposium on Microwave Techniques for Computing Systems, IRE TRANS. ON ELECTRONIC COMPUTERS, September, 1959.)

An experimental high school science program in Westbury, L. I., involving the participation of local industries has met with such success that it might well serve as a model for other communities across the nation. The experiment was spearheaded by a group of IRE members who were convinced that engineers, especially electronics engineers, are in a key position to influence the attitude of their communities on matters of science education. The plan was to hold an industrial Science Fair which would be directed towards the schools and would show where science education ultimately leads. The exhibits were to be supplied by local scientific and industrial organizations and be manned by high school students. A citizens committee of engineers and science teachers was formed to contact the appropriate organizations in the area and help them to select and plan their exhibits. Each exhibitor was then assigned a team composed of a high school science teacher and a group of selected students who were later to man the exhibit at the Fair. Each team visited their exhibitor's plant in advance of the Fair to familiarize themselves with the exhibit. For many students, these field trips were the high point of the whole project. All in all, 10 high schools, 40 exhibitors, 40 science teachers and 400 students were involved in this phase of the project. The success of the experiment is attested to by the fact that 25,000 people, both young and old, came to the Westbury High School to see the three-day Science Fair. (R. K. Hellmann, "An engineering experiment in industry-community-school relations," IRE TRANS. ON EDUCATION, September, 1959.)

Super-pressure research activities are being watched with growing interest by electronics people, especially those in the components field, as a potentially important source of new materials. During the past decade controllable pressures of 2 million p.s.i. have been achieved and exploited to produce synthetic diamonds on a commercial basis. The Fort Monmouth Signal Corps Laboratory is now working on a high-pressure chamber which will develop 3 million p.s.i. Meanwhile, research physicists have set as their next goal controllable pressures as high as 6 million p.s.i. It is expected that super pressures of this magnitude will shift electrons out of their normal orbits and create new dense forms of some of our known materials. It is theoretically possible, for example, that such gases as ammonia or hydrogen will be made available in a metallic state by these super-pressure techniques. What the fruits of these efforts will be, no one can tell for certain, except to say that once the problems of generating and controlling super pressures are solved, new materials as yet unknown to man can be created for exploitation by components engineers. (L. J. D. Rouge, "Army electronics research: theory to reality," IRE TRANS. ON COMPONENT PARTS, September, 1959.)

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*
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Audio

VOL. AU-7, NO. 5, SEPTEMBER, 1959

The Editor's Corner (p. 1)

PGA News (p. 110)

The "Null Method" of Azimuth Alignment in Multitrack Magnetic Tape Recording—A. G. Evans (p. 116)

A number of methods for azimuth alignment were investigated. A technique for alignment which compares the output from two tracks of a multitrack tape provided a substantial improvement in alignment accuracy as compared to the methods which had been in use up to this time. A method for adjusting the the lateral position of the head across the width of the tape was also developed which made use of the same basic principles as the "null method" of azimuth alignment.

Phase Shift in Loudspeakers—W. R. Stroh (p. 120)

A simple method of measuring the phase characteristic of a loudspeaker is described and typical phase curves are given for moving-coil and electrostatic loudspeakers. Distortion in correlation functions measured with an electrostatic loudspeaker is described and related to the phase characteristic of the speaker.

A Two-Watt Transistor Audio Amplifier—W. D. Roehr (p. 125)

For low distortion, power transistors should be driven from a low impedance source. Thus an emitter-follower driver has definite appeal. A further advantage is that the driver transistor may be direct coupled to the output transistor.

This paper describes the basic design method for this circuit and illustrates its performance. Important considerations such as stability, transistor interchangeability, frequency response, and distortion are discussed and typical measurements show.

The features of this circuit are: good stability and the facts that transistor parameters are noncritical, no bias adjustments are required, frequency response is flat over the audio range, and distortion is low.

Nonlinear Distortion Reduction by Complementary Distortion—J. R. Macdonald (p. 128)

Nonlinear distortion produced in a given circuit can be reduced by pre- or postdistorting the signal applied to or from the circuit. Such complementary distortion cannot reduce the original distortion to zero in practice because of distortion of distortion, but it can result in greatly reduced output distortion over a limited amplitude range. General results for the design of pre- or postdistortion circuits are given, and the mathematical results are illustrated by comparing the total harmonic distortions obtained with pre- and postdistortion corrections of increasing complexity applied to a simple nonlinear circuit.

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Component Parts

VOL. CP-6, NO. 3, SEPTEMBER, 1959

Information for Authors (p. 125)

Who's Who in PGCP (p. 126)

A Message from the Chairman (p. 127)

Progress Report on Ad Hoc Study on Parts Specifications Management for Reliability—E. J. Nucci (p. 128)

Trends of Things to Come—C. H. Lewis (p. 144)

Capacitors—1959—L. Kahn (p. 150)

Electronic Materials—an Industry-Wide Problem—A. M. Hadley (p. 175)

After a review of the reasons for the current emphasis on electronic materials, both in civilian industry and in the military, the government-sponsored research and development program in this area is reviewed and its basic objectives outlined. A brief discussion of probable long-term investigations in the area of electronic materials is followed by a listing of immediate materials requirements based on specific recommendations of qualified representatives of the Army, Navy, and Air Force.

Emphasis throughout the paper is directed at the need for close cooperation between industry and the military, and two procedures—one conventional and obvious, and the second of a less conventional nature—are proposed as avenues leading toward an improvement in the current materials situation.

Uniform Cooling Air Flow During Computer Maintenance and Operation—A. Perlmutter (p. 180)

Many designs have recently evolved for cooling digital equipments. The Sylvania Waltham Laboratories, in building the Universal Digital Operational Flight Trainer, UDOFT, has developed a cooling method applicable to many systems. Every designer is faced with the dilemma of obtaining an optimum balance between ease of testing finished equipment and ease of construction testing of the first engineering system. In the cooling plan used in UDOFT, both types of test work as well as system operation maintenance can be accomplished without disturbing the air-flow path over the many small plug-in packages which contain the computer circuitry. Any package, or all packages, can be removed from their respective slots and all remaining packages will still receive the same constant air supply. This is accomplished without any moving flow-control devices and without gasketing seals. The hot-spot temperature of the electron tubes and other heat sensitive parts is controlled by maintaining a known, fixed air flow through each individual package. This arrangement provides optimum thermal control of the system, thereby improving both component and total system reliability.

A Practical, Comprehensive Component Application Program—C. G. Walance (p. 190)

Army Electronics Research: Theory to Reality—L. J. D. Rouge (p. 193)

A Review of the Influence of Recent Material and Technique Development on Transformer Design—H. M. Nordenberg (p. 201)

The requirements for greater miniaturization, higher operating temperature capability, and better moisture protection have resulted in the development of new constructional techniques for electronic transformers. The developments in application of new materials and techniques are detailed. Among the new developments discussed are the application of fluorochemicals to transformers designed for maximum coil temperatures of 200°C.; the philosophy of "controlled environment" for the insulation system through the application of inert gases for transformers designed for maximum coil temperatures of 350°C., and the application of epoxies to the molded coil encapsulated transformer.

Improvements Made in Electronic Parts During the Past Ten Years—H. V. Noble (p. 210)

An Analysis of Printed Wire Connectors—D. R. Sheriff (p. 223)

Education

Vol. E-2, No. 4, September, 1959

An Engineering Experiment in Industry-Community-School Relations—R. K. Hellmann (p. 113)

The engineering experiment described here was spearheaded by a group of IRE Members who were convinced that the attitude of their communities on matters of science education could be influenced. In this project, ten high schools and over 40 industrial and scientific organizations joined in a series of supervised field trips by 400 selected high school students. The program culminated in a three-day industrial exhibit which was viewed by approximately 25,000 visitors of all ages. Details on how this was organized as a community project are given.

Physics Courses in the New Arizona Engineering Curriculum—A. B. Weaver (p. 117)

To facilitate the adoption of a common core curriculum in the Schools of Engineering and of Mining at the University of Arizona, the general physics course for engineers has been dropped. The physical foundation for engineering is taught in a new series of courses in the Civil, Mechanical, Electrical Engineering, and Physics Departments as part of the common core. The Physics Department teaches atomic, molecular, nuclear, and solid-state physics in a seven-unit junior-level course. There are grounds for a re-examination of the role of physics in engineering education and perhaps for extensive changes. It is suggested, however, that the role is, and should remain, vital, in the interests of science, of engineering, and of education.

Laboratory: Its Scope and Philosophy—R. F. Schwartz (p. 120)

Some of the problems, methods, and philosophies of teaching laboratory are reviewed. Laboratory furnishes one channel by which knowledge can be acquired, supplements other teaching methods, and is particularly useful where mathematical descriptions are imperfect. In addition, laboratory deals with precision measurement, data recording, development of mechanical aptitude, and learning of specific techniques. Coordination of laboratory with other work usually follows one of two conflicting principles. A laboratory course may be conducted in a rigid manner or in a highly individualistic, flexible way. Each has its advantages. Laboratory provides good practice for report writing. By requiring at least one formal paper per student on laboratory work performed, a more realistic introduction to report writing is secured than in any other course work.

A Course in Engineering Analysis for Superior Students—G. B. Hoadley (p. 122)

As a part of the expanding honors group program for superior students, a course in Engineering Analysis using the assistance of engineers from cooperating industry has been evolved. The selection of students, the orientation part of the course, the selection of problems from industry, and a discussion of the results are presented.

Elementary Introduction to Electrodynamics—V. Bevc (p. 124)

Objectionable examples of time-varying electromagnetic fields occurring in elementary textbooks are pointed out and the difficulties arising in connection with them are discussed. The interdependence of spatial and time variations of the electromagnetic fields is recalled. As an illustration, two simple examples are given to show how the variation of fields with time is automatically determined by the wave equation and initial spatial distribution. It is suggested that only fields which satisfy the wave equation be used in textbooks.

A Combined Machinery and Control Systems Laboratory—W. A. Blackwell and H. E. Koenig (p. 128)

This paper shows how, in the interest of emphasizing the systems concept, rotating machines have been integrated into a systems laboratory. Small machines provide the mechanical flexibility necessary in the emphasis of the systems concept, yet essential machine characteristics are retained. The paper also demonstrates how machine characteristics are interpreted to provide information for the analysis for an electromechanical system which includes the machine as a component.

A Realistic Program in Energy Processing—R. D. Chenoweth (p. 134)

A general revision of the undergraduate curriculum in electrical engineering at Case Institute of Technology served as the stimulus and foundation for an updating of the course content in the area previously occupied by rotating machinery. The series of three courses which resulted has now been tested in the classroom for appropriateness, unity, depth of coverage possible, and palatability to the student. This paper reports the content and achievement of this initial trial.

Results of the past year indicate that the material presented is pedagogically sound and that the program is realistic for the undergraduate student.

Energy Conversion and Control at Berkeley—H. C. Bourne, *et al.* (p. 138)

The recent developments in this area in terms of curricular changes in the Electrical Engineering program have been described and evaluated. A previous required course, which covered electric machinery in some detail, has been replaced by a more general course which studies the principles of electromechanical energy conversion and the control of the flow of electrical energy by both static and motional devices. In addition, a new experimental two-semester course has been developed which integrates the study of energy conversion and control devices into a study of systems and the role of these devices in meeting system specifications.

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Electronic Computers

Vol. EC-8, No. 3, September, 1959

The Chairman's Column—R. O. Endres (p. 261)

The ONR Symposium on Microwave Techniques for Computing Systems—M. C. Yovits (p. 262)

History and Introduction—Microwave Techniques for Computers—R. E. Meagher (p. 263)

Nanosecond Logic by Amplitude Modulation at X-Band—W. C. G. Ortel (p. 265)

A basic circuit, consisting of a diode modulator controlled by the signal from a diode detector, may perform logical AND, EXCLUSIVE-OR and OR functions upon pulsed microwave signals. Pulse rates up to 500 mc have been used at a carrier frequency of 11,000 mc. To demonstrate that microwave circuits may be used for the regeneration and circulating storage of pulses, as well as for logic, a digital arithmetic unit has been built which multiplies two 8-digit binary numbers. Various forms of the basic circuit have been studied in operation.

A Logic Design for a Microwave Computer—S. Frankel (p. 271)

The properties of presently available components place special emphasis on two desiderata of logic design for use in a microwave digital computer: 1) Smallness of the number of active elements; 2) elimination of information-cycling paths having delay times comparable or short compared with the bit period, as in the conventional flip-flop. A logic design developed in response to these pressures is described in substantially complete detail. Property 1) is obtained by the use throughout of a multiplexing procedure such that the computer functionally (although not physically) resembles a number of nearly identical, and correspondingly slower, computers which are able to operate either independently or in concert.

Parametric Phase-Locked Oscillator—Characteristics and Applications to Digital Systems—L. S. Onyshkevych, *et al.* (p. 277)

The ability of the Parametric Phase-Locked Oscillator (PLO) to detect, amplify, and store binary digital signals, in the form of two distinct phases of a carrier, makes it possible to use the device as the sole component in a digital computer system. The variable-capacitance version of the device operates readily at kilomegacycle frequencies, thus forming the basis of a digital computer at a kilomegapulse clock rate.

In the present paper the results of an investigation of the behavior and possible applications of the variable-capacitance PLO are presented. The investigation was supported by experimental work with lumped-component variable-capacitance PLO's at 5 mc, and microwave variable-capacitance PLO's at 4 kmc.

The steady-state behavior of the device is described; variations of the output voltage with pump voltage, loading, tuning and frequency variations are presented in the form of characteristic curves. Results indicate that the device is rather insensitive to reasonable changes in operating conditions and parameter values.

The transient behavior of the PLO shows that the device can be switched in a number of different ways. Five such modes of operation are discussed; these are phase initiation, forced switching, burst generation, tri-stable operation and unconditional switching. Each of these modes has particular advantages for various applications. Switching times of the order of 3 to 10 cycles of the signal frequency are readily obtainable.

The various modes of operation of the device suggest a number of applications both in logic and in memory. To illustrate the versatility of the device, a random access memory is described as an example.

Semiconductor Parametric Diodes in Microwave Computers—J. Hilibrand, *et al.* (p. 287)

The parametric subharmonic oscillator operates by energy transfer from the pump frequency to the oscillator frequency through a nonlinear energy storage element—in the present case, the nonlinear capacitance of a semiconductor diode. This paper examines both the requirements on the diode for satisfactory performance in this circuit and the limitations on oscillator performance which arise from the nature of the semiconductor diode.

The analysis shows that abrupt junction diodes must have a Q of at least four at the oscillation frequency if there is to be any useable energy transfer, and that graded junction diodes must have a Q of six. The time constant governing the rise of the envelope of the subharmonic waveform is a marked function of the stray capacitances; this function is examined in detail. The choice of bias voltage to obtain the fastest possible rise time involves consideration of the stray capacitance, the Q of the available diode, and limitations imposed by excessive pump power requirements. For negligible stray capacitance, it is shown that the subharmonic waveform can rise by a factor e in 1.3 cycles of the subharmonic frequency for an abrupt junction diode, or in 1.9 cycles for a graded junction diode.

The principles involved in the design of the semiconductor diode are examined and the choice of materials, impurity distributions, and fabrication techniques are discussed. A new diode encapsulation intended for direct mounting in microstrip transmission line is described.

Measurement of appropriate diode parameters is vital to diode research. An equivalent circuit characterization in which the parameters may be directly related to the diode structure is used. Several techniques for the measurement of these parameters are discussed.

Fast Microwave Logic Circuits—D. J. Blattner and F. Sterzer (p. 297)

In a carrier-type digital computer system, binary information can be represented by the presence or absence of an RF pulse in a given time interval. Using strip-line printed circuit techniques and point-contact diodes, passive AND and NOT gates were constructed which operate with RF pulses of less than 2 mμsec duration (*i.e.*, an effective pulse repetition rate of 500 mc), at a carrier frequency of 3000 mc. The basic gates were combined to form half-adders. Unlike other carrier approaches, these circuits keep the information in RF form through all steps of the logic operations; *i.e.*, both inputs and outputs of all elements are RF pulses.

Microwave Logic Circuits Using Diodes—W. Sauter and P. J. Isaacs (p. 302)

It is possible to control the transmission of microwave power in a waveguide via external control of the dc bias on a semiconductor diode mounted across the waveguide in a direction parallel to the E field. The combination of a microwave detector with such a modulator affords a means whereby RF power is one waveguide can be made to control RF power in a second waveguide. In order to test the applicability of this circuit to binary logic functions, a regenerative memory loop has been constructed. Traveling-wave tubes were employed to raise the level of a controlled signal to that required by the detector. Using an X-band-carrier, binary pulse stability was observed at pulse repetition rates of 685 mc.

Properties of Propagating Structures with Variable Parameter Elements (Abstract)—N. Kroll (p. 307)

The Parametron Digital Computer MUSASINO-1—S. Muroga and K. Takashima (p. 308)

Features of a large-scale digital computer with novel logical elements, the parametrons, are described. The machine, which is located at Musashino City, Tokyo, was named the MUSASINO-1, and has been in almost continuous operation since its completion in the spring of 1957. Primarily for scientific uses, it does arithmetic operations in parallel, and has a fast access memory of ferrite cores with nonrectangular hysteresis curve. Maintenance experience has indicated its extreme stability and low incidence of faults.

A Glow Counting Tube Read-Out Technique and Its Application—S. K. Chao (p. 317)

The cold cathode counting tube, also called the decade glow transfer tube, is used principally in preset counter, timing and gating circuits. The number stored in the tube is often required to be recorded. This paper describes a technique whereby the content of the tube is recognized and read out through a carrier signal applied to the anode and 10 detectors connected to the 10 cathodes. The readout is of the nondestructive type since it does not alter the content of the tube.

A large number of glow tubes can be conveniently read out in this manner simply by connecting all corresponding cathodes together. The carrier signal is then successively distributed to their anodes. An example of such an application is given where 19 channels of four glow tubes each are read into an IBM card punch.

An Idealized Over-All Error-Correcting Digital Computer Having Only an Error-Detecting Combinational Part—W. L. Kilmer (p. 321)

The block diagram of an idealized over-all error-correcting digital computer is presented. This computer has the property that during each unit time interval, it can correct the effects of a specific maximum number of transient-type component failures which might occur anywhere within it. Yet, all its combinational logic circuitry is only of the error-detecting type. The corresponding reduction in equipment that this design feature makes possible is achieved at the expense of the computer's having to sit idle during a large percentage of those time intervals in which component failures occur. In a sense, therefore, the computer utilizes a great deal of time-domain redundancy as well as equipment-domain redundancy. This paper discusses some of the design requirements that are involved in using this type of redundancy structure

System Organization of a Multiple-Cockpit Digital Operational Flight Trainer—H. J. Gray, Jr., *et al.* (p. 326)

This paper describes the system organization of a digital computer whose purpose is to activate simultaneously more than one cockpit of an operational flight trainer. The simulated aircraft are assumed to be all of the same type, but each is simulated independently. The computer is drum-sequenced and represents an application of the theory of multiple computers, since there are several different kinds of memories and more than one arithmetic unit in the system.

The CORDIC Trigonometric Computing Technique—J. E. Volder (p. 330)

The *CO*ordinate *R*otation *DI*gital *C*omputer (CORDIC) is a special-purpose digital computer for real-time airborne computation. In this computer, a unique computing technique is employed which is especially suitable for solving the trigonometric relationships involved in plane coordinate rotation and conversion from rectangular to polar coordinates. CORDIC is an entire-transfer computer; it contains a special serial arithmetic unit consisting of three shift registers, three adder-subtractors, and special interconnections. By use of a prescribed sequence of conditional additions or subtractions, the CORDIC arithmetic unit can be controlled to solve either set of the following equations:

$$Y' = K(Y \cos \lambda + X \sin \lambda)$$

$$X' = K(X \cos \lambda - Y \sin \lambda)$$

or

$$R = K\sqrt{X^2 + Y^2}$$

$$\theta = \tan^{-1} Y/X,$$

where K is an invariable constant.

This special arithmetic unit is also suitable for other computations such as multiplication, division, and the conversion between binary and mixed radix number systems. However, only the trigonometric algorithms used in this computer and the instrumentation of these algorithms are discussed in this paper.

Decimal-Binary Conversions in CORDIC—D. H. Daggett (p. 335)

A special-purpose, binary computer called CORDIC (*CO*ordinate *R*otation *DI*gital *C*omputer) contains a unique arithmetic unit composed of three shift registers, three adder-subtractors, and suitable interconnections for efficiently performing calculations involving trigonometric functions. A technique is formulated for using the CORDIC arithmetic unit to convert between angles expressed in degrees and minutes in the 8, 4, 2, 1 code and angles expressed in binary fractions of a half revolution. Decimal-to-binary conversion is accomplished through the generation of an intermediate binary code in which the variable values are +1 and −1. Each of these intermediate code variables controls the addition or subtraction of a particular binary constant in the formation of accumulated sum which represents the angle. Examples are presented to illustrate the technique. Binary-to-decimal conversion is accomplished by applying essentially the same conversion steps in reverse order, but this feature is not discussed fully. Fundamental principles of the conversion technique, rather than details of implementation, are emphasized. The CORDIC conversion technique is sufficiently general to be applied to decimal-binary conversion problems involving other mixed radix systems and other decimal codes.

Minimal Sequential Machines—D. B. Netherwood (p. 339)

The general class of sequential machines defined by Mealy is investigated. It is found that any such machine can be identified with a set of machines of equivalent minimality. A procedure for developing the aggregate of all sets of gates for such minimal machines is evolved, and the problem of selecting components for constructing machines is discussed.

A Technique for the Reduction of a Given Machine to a Minimal-State Machine—S. Ginsburg (p. 346)

A technique is presented for reducing an arbitrary machine S as much as possible to a machine T which can do everything (from the input-output point of view) that S can do. Since the technique is always applicable, it is more powerful (although more cumbersome) than the well-known merging technique. Several examples are given.

Minimizing the Number of States in Incompletely Specified Sequential Switching Functions—M. C. Paull and S. H. Unger (p. 356)

Given a sequential switching function in the form of a flow table in which some of the entries are unspecified, the problem of reducing the number of rows in that flow table is extremely complex, and cannot, in general, be solved by any simple extension of the methods used for completely specified functions. An analysis of the problem is presented, and a partially enumerative solution is evolved. A rough indication of the efficiency of the given procedures may be obtained from the fact that these techniques have been successfully applied to approximately two dozen tables ranging up to about 15 rows. No solution required more than two hours.

Logical Machine Design II: A Selected Bibliography—D. B. Netherwood (p. 367)

The bibliography which appeared in the June, 1958 issue of these TRANSACTIONS is extended to a total of 777 titles. The original format is retained, but in this supplement the scope of material is restricted to technical publications pertaining to the logical design of machines.

Operational Analog Simulation of the Vibration of a Beam and a Rectangular Multicellular Structure—A. B. Clymer (p. 381)

A feasibility study of the use of an operational analog computer for solution of structural problems was undertaken. A beam problem and a rectangular multicellular structure problem were run to test the method. In this paper, which is a progress report, it is shown that the method is highly competitive with digital computer and passive-element computer methods for solution of any structural problem.

The Design of Position and Velocity Servos for Multiplying and Function Generation—E. O. Gilbert (p. 391)

The design of position and velocity servos used in analog computation and simulation for multiplying and function generation is considered. The important characteristics of potentiometers, gear train, motor, amplifier, and tachometer are defined and discussed. Nonlinear performance requirement, such as velocity and acceleration limits, overshoot for large step inputs, and static resolution, are defined in terms of component parameters. A minimum gear reduction ratio is determined on the basis of acceleration, frictional torque ratio, overshoot for large step inputs, or static resolution. Linear system analysis is made and related to system components and nonlinear performance; in particular, it is shown that static resolution is limited by servo amplifier bandwidth for given motor, potentiometers, and gear train. The selection of damping methods and the reduction of steady-state errors is described. An example design is considered.

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Vehicular Communications

PGVC-13, September, 1959

Meeting the Demands for Vehicular Communications—C. B. Plummer (p. 1)

Interference—A Look at the Ounce of Prevention—N. H. Shepherd and A. C. Giesselman (p. 4)

One of the major interference problems in the field of vehicular communications is brought about by the concentration of a number of base station installations in a relatively small area, or even at the same antenna site. Most cases of interference resulting from such multiple installations are difficult to analyze after the fact. A method is described for predicting such interference before it occurs. Furthermore, as a method for alleviating radio interference problems due to frequency and geographic congestion, a technique is suggested for processing new applications for frequency assignment.

Operation of Close-Spaced Antennas in Radio Relay Systems—W. F. Biggerstaff (p. 11)

The required attenuation between transmitter output and receiver input for any given pair of equipments may be measured by a method that takes into account both receiver desensitization caused by front end overloading and the interference effects of transmitter noise sidebands. These are the two principal causes of degradation in receiver performance where simultaneous transmission and reception is to be accomplished from antennas to be placed in close proximity. With the required isolation known, curves may be consulted to determine the necessary antenna physical spacing, either colinear or in the same plane. The actual attenuation between two antennas including their associated transmission lines may be measured and, lastly, the presence of any receiver desensitization may be determined under practical operating conditions. A method for increasing the isolation between close-spaced vertical antennas for a given transmitter operating frequency is described.

A Report on Interference Caused by Intermodulation Products Generated In or Near Land Mobile Transmitters—N. H. Shepherd (p. 16)

This report is a condensation of published papers combined with unpublished information on the subject of intermodulation products generation and interference created in or near land mobile transmitters. The following papers have been used in preparation of this report:

Radio Report No. 50, "Investigation of Intermodulation Interference in the 100–108 Mc/S Mobile Radio-Telephone Service." Report prepared by C. H. Sturton assisted by G. J. Campbell and B. T. Harnett of the New Zealand Post Office.

"Interference Due to Intermodulation Products UHF Maritime Service," Special Committee No. 36 of the RTCM.

N. H. Shepherd. "Bi-Dimensional Interference Analysis," Proceedings of the Third ference on Radio Interference Reduction.

Each of these papers provide excellent material that is useful as a continual reminder of the effects of interference and practical techniques for tracking down and reducing interference. A particular type of interference, intermodulation, is troublesome to locate since it can be generated in a variety of devices.

A VHF Pocket Receiver—J. F. Mitchell (p. 20)

Motor Carrier Mobile Radio Systems—C. Johnston (p. 24)

An Analysis of Radio Flutter in Future Communications—N. W. Feldman (p. 27)

It is expected that future military operations will be extremely mobile. It is planned to provide for the Army full duplex communications while its subscribers are in motion. This will require the use of radio for serving the subscribers.

Reception from or in a vehicle in motion is plagued by a malady known as flutter. It results in cyclical impairment of reception. It is caused by the varying field strength pattern generated on multiple path reception of the same radio signal. The audio signal level can be maintained by an AVC. The increase of noise level during operation of AVC can adversely affect the operation of system signaling.

All sorts of schemes are being studied to overcome the effect of the flutter. In this paper the flutter effects on the communication system will be analyzed. The improvements resulting from certain remedies will be predicted. Propagation data at 150 mc will be presented. Emphasis will be placed on the switched antenna type of space diversity reception. The value of this technique will be established by a brief discussion of the benefits that may be expected.

Recent advances in the Solid State Field make feasible the design and construction of an applique unit which will assure us of benefits of space diversity reception at a relatively low cost. The design of antenna switching circuitry which operates on a programmed basis will be described. Problems generated by these techniques will be aired.

FM Interference and Noise-Suppression Properties of the Oscillating Limiter—E. J. Baghdady (p. 37)

An amplitude limiter with regenerative feedback can, under appropriate conditions, provide an effective tool for suppressing interference from undesired signals, as well as interstation noise. The amount of positive narrowband feedback that is prescribed for marked improvement in the stronger-signal capture performance causes the limiter to oscillate in the absence of an input signal. Although oscillation results in automatic noise squelch, it also imposes limitations on the frequencies and minimum amplitudes of receivable signals, in the form of a locking frequency range and a locking threshold. The alleviation of these limitations requires careful design of the feedback phase characteristic. The theoretical discussion is followed by a summary of experimental data which brings out several important aspects of oscillating-limiter operation.

A New Approach to Compactness in Mobile Radio-Telephone Design—W. Ornstein (p. 64)

A mobile radiotelephone having a power rating of 30 watts is described. The unit incorporates transmitter, receiver, operating controls and loudspeaker in a single compact aluminum alloy case. Total weight is less than twelve pounds. The unit is small enough for underdash mounting in a modern passenger car. The transistor power supply is in a separate case designed to be mounted on the vehicle firewall. Heater voltage regulation is employed to maximize vacuum tube reliability. Careful attention to thermal design has minimized the development of hot spots in the interior of the equipment.

A New Manual Mobile Telephone System—A. F. Culbertson (p. 73)

A new manual mobile telephone system has been designed around the requirements of operation by "landline" telephone companies in the domestic public land mobile service. A specific objective has been a significant reduction in first cost and annual charges of the system to the operating companies. Achievement of this objective has resulted in a system which is expected to bring mobile telephone service to many areas which have previously been without it. This system makes mobile service feasible in many smaller exchanges which are typical of those served by the Independent Telephone Industry.

Performance of "Low-Plate-Potential" Tube Types at Mobile-Communications Frequencies (Abstract)—R. J. Nelson and C. Gonzales (p. 83)

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

The Index to the Abstracts and References published in the PROC. IRE from February, 1958 through January, 1959 is published by the PROC. IRE, May, 1959, Part II. It is also published by *Electronic and Radio Engineer*, incorporating *Wireless Engineer*, and included in the March, 1959 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.21:534.6 3900
Experimental Study of the Propagation of Sound over Ground—F. M. Wiener and D. N. Keast. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 724–733; June, 1959.) The propagation of sound over open level ground, through dense evergreen forests and between hilltops has been investigated in the frequency range 300–5000 cps. The instrumentation used to measure relevant micrometeorological parameters has been described by Wiener *et al.* (2081 of July). Charts based on the measurements are given for estimating the mean excess attenuation as a function of frequency and distance for a given set of micrometeorological conditions.

534.213–14:534.88 3901
The Field of a Pulse Radiator in an Underwater Sound Channel—V. A. Polyanskaya. (*Akust. Z.*, vol. 5, no. 1, pp. 91–100; 1959.) Investigation of sound propagation in an underwater channel by the method of normal waves in the WKB approximation for various pulse shapes.

534.213–14:534.88 3902
Scattering of Sound by Nonuniformities in a Layer of Discontinuity in the Sea—G. D. Malyuzhinets. (*Akust. Z.*, vol. 5, no. 1, pp. 70–76; 1959.) An approximate calculation of reverberation intensity and scattering coefficients for a sound wave propagated near a transition layer between water of different temperatures in the presence of internal gravitation waves.

534.22–14:534.88 3903
On the Calculation of the Velocity of Sound in Sea-Water—C. Froese. (*Canad. J. Phys.*, vol. 37, pp. 775–779; June, 1959.) Kuwahara's tables are checked using the same basic formulas but a different method of calculation.

534.232 3904
Electrical Equivalent Circuits for Transient Vibrations of Electromechanical Transducers of Active Material (Piezoelectric, Electrostrictive or Magnetostrictive)—F. A. Fischer. (*Akust. Beihefte*, no. 1, pp. 215–220; 1959.)

534.232 3905
Calculating the Concentration Coefficient of some Directional Acoustic Systems—N. F. Vollerner and M. I. Karnovskii. (*Akust. Z.*, vol. 5, no. 1, pp. 25–30; 1959). An application of the reciprocity theorem to establish relations between amplification factor and axial concentration coefficient for the case of cylindrical focusing systems.

534.232.089.6 3906
On the Possibility of Investigating the Frequency Sensitivity Characteristics of Transducers by a Spectral Analysis of their Thermal Noise—K. V. Goncharov. (*Akust. Z.*, vol. 5, no. 1, pp. 120–122; 1959.) Outline of a method for calibrating hydrophones using a thermal-noise analyzer. The method is based on a relation between the spectral density of thermal noise and an impedance/frequency characteristic $ReZ(\nu)$ derived earlier for piezoelectric crystals [2770 of 1956 (Goncharov and Krasil'nikov)].

534.414 3907
On the Treatment of Excitation and Perturbation Problems in Acoustic Resonators—W. Pechhold. (*Acustica*, vol. 9, no. 1, pp. 48–56; 1959. In German.) An application of an integral theorem of continuum mechanics to the determination of resonance conditions. The method is illustrated for the case of simple vibrations of a resonator in the form of a circular cylinder, and equivalent-circuit data for excitation by different types of transducer are tabulated.

534.44 3908
A Formation Rule for the Determination of the Frequency Spectrum of Power-Law Double Tones—W. Grahnert. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 4–18; July, 1958. Correction, vol. 67, p. 67; September, 1958.) The equation derived facilitates the evaluation of all the harmonics arising in a double-tone signal of the form $\hat{u}_a \cos \omega_a t + \hat{u}_b \cos \omega_b t$ when raised to the nth power, $\hat{u}_a$, $\hat{u}_b$ and ω_a, ω_b respectively being the amplitudes and frequencies of the constituent tones.

534.75 3909
Relation between Hearing Threshold and Duration for Tone Pulses—R. Plomp and M. A. Bouman. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 749–758; June, 1959.)

534.75 3910
Interference and Coherence of Acoustic Signals—V. V. Furduev. (*Akust. Z.*, vol. 5, no. 1, pp. 111–116; 1959.) The variation of the interference effect in respect of both magnitude and sign observed with reiterated speech or musical signals is considered with reference to the averaging time of the receiver. This variation is expressed in terms of running autocorrelation function from which the coherence interval can be determined.

534.75 3911
The Sensation of Loudness of Rhythmic Sounds of Very Low Periodicity—H. Niese. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 26–34; July, 1958.) Subjective tests on 1000-cps tones and white noise pulse-modulated at repetition frequencies up to 100 cps are analyzed, and proposals for the design of an objective loudness meter are based on the results. See also 1426 and 1427 of May.

534.78:534.846.4 3912
The Intelligibility of Reinforced Speech—J. P. A. Lochner and J. F. Burger. (*Acustica*, vol. 9, no. 1, pp. 31–38; 1959.) Discussion of factors affecting intelligibility with reference to results of experiments conducted in the South African House of Assembly and Bloemfontein University Great Hall. See also 5 of January.

534.78:621.39 3913
Automatic Speech Recognition—K. Steinbuch. (*Nachrichtentech. Z.*, vol. 11, pp. 446–454; September, 1958.) Review of progress in the technique on the basis of published information. 45 references.

534.839 3914
On the Frequency Evaluation of Noise Spectra—E. Lübcke. (*Akust. Beihefte*, no. 1, pp. 243–246; 1959.) Sound-pressure/frequency curves used in evaluating noise measurements are discussed and a method of grouping noises into a definite scale is suggested. See also 2453 of August.

534.84 3915
Fluctuations of the Sound Pressure Level in Rooms when the Receiver Position is Varied—P. E. Doak. (*Acustica*, vol. 9, no. 1, pp. 1–9; 1959.) Theoretical treatment of "space irregularity" in rooms with reference to experimental data. Fluctuations are insensitive to wall shape. Far from the source and for frequencies satisfying Schröder's "large-room" condition (618

of 1955) the rms deviation of pressure level is the same for variation in the source or receiver position as it is for variation of the source frequency. Near the source or at lower frequencies fluctuations depend upon source radiation characteristic, reverberation time, room volume and the predominant type of eigenfunctions.

534.84 3916

The Effect of the Room Shape on the Sound Field in Rooms (Studies on the Measurement of Absorption Coefficient by the Reverberation-Chamber Method: Part 1)—K. Sato and M. Koyasu. (*J. Phys. Soc. Japan*, vol. 14, pp. 365–373; March, 1959.) Model room experiments are reported, relating the effect of the shape and size of a room on the steady-state and transient characteristics of low-frequency sound waves. Results illustrate the advantage of a reverberation chamber of irregular shape for obtaining a desired sound field.

534.84 3917

New Wall Design for High Transmission Loss or High Damping—G. Kurtze and B. G. Watters. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 739–748; June, 1959.) A wall design is described in which the ratio of the static to the dynamic stiffness can be in excess of 1000:1 and the acoustic behavior is that of a perfectly limp wall.

534.84 3918

Measurements of Diffusion and Application to Room Acoustics—R. Lamoral. (*Acustica*, vol. 9, no. 1, pp. 57–60; 1959. In French.) Irregularities in sound level are measured at a fixed frequency using a rotating microphone and a special electronic counter [3619 of 1956 (Lamoral and Trembasky)]. The number of irregularities recorded during one rotation of the microphone is plotted as a function of frequency. Curves obtained from measurements in rooms fitted with sound diffusers show discontinuities, in agreement with results of model tests, at ultrasonic frequencies.

534.844.1 3919

Methods for the Measurement of Diffusivity in Reverberation Chambers—M. R. Schroeder. (*Akust. Beihefte*, no. 1, pp. 256–264; 1959.) Diffusivity is defined as the angular distribution of sound energy flux, and the problem of measuring this accurately without disturbing the sound field is discussed. A method based on a spectral analysis of the sound field at the measuring wall is described.

534.845.1 3920

Calculation of the Statistical Absorption Coefficient from Acoustic Impedance Tube Measurements—P. Dubont and W. Davern. (*Acustica*, vol. 9, no. 1, pp. 15–16; 1959.) A chart is given enabling the coefficient to be deduced directly from measured quantities.

534.845.1 3921

A Semi-empirical Method of Calculating Reverberation Chamber Coefficients from Acoustic Impedance Values—B. S. Atal. (*Acustica*, vol. 9, no. 1, pp. 27–30; 1959.) In the method proposed, the value obtained from impedance-tube measurements is multiplied by a suitably chosen complex quantity dependent only on frequency. The absorption coefficient in a reverberant field is derived from this modified impedance value by an averaging procedure. Coefficients have been calculated for nine commercial materials at five standard frequencies.

621.395.623.7:534.6 3922

Evaluation of a Modulated Air-Flow Loudspeaker—J. C. Webster, R. G. Klumpp and A. L. Witchey. (*J. Acoust. Soc. Amer.*, vol. 31, pp. 795–799; June, 1959.) Speech intelligibility and alarm detection tests showed the "air speaker" to be superior to conventional loudspeakers under noise conditions created by jet aircraft idling at 30 per cent full power; neither system was adequate at full power.

621.395.625.3 3923

Magnetic Tape Recording—L. H. Bedford. (*Electronic Radio Eng.*, vol. 36, pp. 320–322; September, 1959.) The basic theory of the recording process is discussed, ignoring fringing effects.

621.395.625.3:534.844 3924

Delay Equipment for Unattended Operation—H. Hepper. (*Telefunken Ztg.*, vol. 31, pp. 200–202; September, 1958. English summary, pp. 206–207.) Equipment for delaying reproduced sound to improve intelligibility in public-address systems is described. A magnetic layer on the rim of a rotating disk is scanned by magnetic tape-recorder heads at a distance of 30 μ from the rim to eliminate wear. The delay time is continuously adjustable between 30 and 975 ms.

ANTENNAS AND TRANSMISSION LINES

621.315.212 3925

Coaxial-Cable Performance—W. A. Cameron. (*Electronic Radio Eng.*, vol. 36, pp. 349–352; September, 1959.) Methods are described for the calculation and measurement of the response of a length of cable to a transient waveform of given shape. A curve is derived for the attenuation of the peak amplitude of a 0.1-μs $\sin^2$ pulse propagating through a 0.375 inch-diameter-coaxial cable.

621.315.212 3926

Compensated Support Disks for 50-Ω Coaxial Lines—G. W. Epprecht. (*Tech. Mitt. PTT*, vol. 36, pp. 373–376; October 1, 1958.) Dimensions of perforated polystyrene and p.t.f.e. supporting and retaining disks are given with curves of residual reflection.

621.315.212.062 3927

Rapid Insertion Device for Coaxial Attenuators—A. Y. Rumfelt and R. J. Como. (*Rev. Sci. Instr.*, vol. 30, pp. 687–688; August, 1959.) An alignment aid during insertion.

621.372.2:621.315.213.12.011.21 3928

The Characteristic Impedance of the Balanced Two-Wire Line in a Rectangular Shield—R. J. F. Guertler. [Proc. IRE (*Australia*), vol. 20, pp. 262–271; May, 1959.] An expression for the characteristic impedance is obtained in a closed form, and design charts are given.

621.372.8 3929

Propagation of Fundamental Modes in Circular or Square Curved Waveguides of Constant Curvature—M. G. Andreasen. (*Arch. elekt. Übertragung*, vol. 12, pp. 414–418; September, 1958.) The undesirable energy conversion of the two cross-polarized fundamental modes in curved waveguides is analyzed. The fundamental modes polarized in parallel with the plane of curvature are those most strongly affected by the curvature of the waveguide.

621.372.8:537.226 3930

Some Characteristics of Dielectric Image Lines at Millimetre Wavelengths—J. C. Wiltse. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 65–69; January, 1959. Abstract, Proc. IRE, vol. 47, p. 498; March, 1959.)

621.372.821:621.372.831.6 3931

A Wide-Band Strip-Line Balun—E. M. T. Jones and J. K. Shimizu. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 128–134; January, 1959. Abstract, Proc. IRE, vol. 47, p. 499; March, 1959.)

621.372.823:621.372.83 3932

Mode Conversion at the Junction of Helix Waveguide and Copper Pipe—J. W. Lechleider. (*Bell Sys. Tech. J.*, vol. 38, pp. 1317–1329; September, 1959.) Scattering coefficients for the junction are calculated, taking into account reflected modes but neglecting fields exterior to the helix in the helix waveguide.

621.372.823:621.372.853.2 3933

A Gyrotropic Elliptical Waveguide—E. S. Kovalenko. (*Dokl. Akad. Nauk SSSR*, vol. 128, pp. 276–279; September 11, 1959.) Mathematical analysis of a waveguide with elliptical cross-section and loaded with ferrite magnetized parallel to the waveguide axis. The normal waves are elliptically polarized except at the walls of the waveguide where the polarization is linear.

621.372.829 3934

Periodic Structures in Trough Waveguide—A. A. Oliner and W. Rotman. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 134–142; January, 1959. Abstract, Proc. IRE, vol. 47, p. 499; March, 1959.)

621.372.832.4 3935

Mode Couplers and Multimode Measurement Techniques—D. J. Lewis. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 110–116; January, 1959. Abstract, Proc. IRE, vol. 47, p. 499; March, 1959.)

621.372.832.43 3936

Microwave Directional Couplers—G. L. Allerton. (*Electronics*, vol. 32, pp. 40–41; September 18, 1959.) Four types of coupler are briefly compared.

621.372.85:621.318.134 3937

Ferrite High-Power Effects in Waveguides—E. Stern and R. S. Mangiaracina. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 11–15; January, 1959. Abstract, Proc. IRE, vol. 47, p. 497; March, 1959.)

621.372.85:621.318.134 3938

Temperature Effects in Microwave Ferrite Devices—J. L. Melchor and P. H. Vartanian. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 15–18; January, 1959. Abstract, Proc. IRE, vol. 47, p. 497; March, 1959.)

621.372.852.1 3939

The Design of Band-Pass Filters in Waveguides—S. S. Forte. (*Marconi Rev.*, vol. 22, pp. 99–116; 2nd Quarter 1959.) Theoretical analysis is given for the "maximally flat" type starting from the low-pass lumped-element prototype. This form of analysis is then applied to the "Tchebycheff equal-ripple response" case and relations are given for a complete design of such filters. The practical realization of the resulting network of either type is described.

621.372.852.21:621.317.328 3940

Measurement of Electric-Field Distributions in a Waveguide containing a Dielectric Slab—K. W. H. Foulds and P. M. J. C. da S. Sampaio. (Proc. IRE, vol. 47, pp. 1663–1664; September, 1959.) The sharp increase in field strength at the dielectric interface (see Strauss, 1958 IRE Wescon Convention Record, vol. 2, pt 1, pp. 135–146) is shown to be associated with the technique of measurement.

621.372.852.22 3941

Ferrite Phase Shifter for the U.H.F. Region—C. M. Johnson, (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 27–31; January, 1959. Abstract, Proc. IRE, vol. 47, p. 498; March, 1959.)

621.372.852.22:621.372.632 3942
A Ferrite Serrodyne for Microwave Frequency Translation—F. J. O'Hara and H. Scharfman. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 32-37; January, 1959. Abstract, PROC. IRE, vol. 47, p. 498; March, 1959.)

621.372.853.2 3943
Characteristics of Ferrite Microwave Limiters—G. S. Uebele. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 18-23; January, 1959. Abstract, PROC. IRE, vol. 47, pp. 497-498; March, 1959.)

621.396.67 3944
High-Frequency Power Transmission at High Efficiency—A. Ditl. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 59-64; September, 1958.) The highest possible efficiency of hf power transmission between antennas a short distance apart is calculated from Maxwell's equations. The efficiency is lower than that given by Fränz's formula (2898 of 1944), the correctness of which is confirmed for large distances. Dipoles and beam antennas are considered as examples.

621.396.67 3945
A Contribution to the Theory of Cylindrical Antennas—Radiation between Parallel Plates—L. Lewin. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 162-168; April, 1959. Abstract, PROC. IRE, vol. 47, p. 1282; July, 1959.)

621.396.67 3946
Bandwidth and Efficiency of Short Antenna Systems ($h<\lambda/10$)—H. E. Fröling. (*Frequenz*, vol. 12, pp. 273-284 and 324-330; September and October, 1958.) Linear, folded and frame antennas of various diameters, without or with capacitive loading, used as omnidirectional single antennas or as directional arrays are investigated. Formulas are derived for two methods of calculating bandwidth, and are applied to practical examples. The bandwidth can be increased by enlarging the diameter of the vertical section of the antenna and at the same time increasing the loading capacitance up to a limiting value. In comparison with the other types, frame-type antennas with arms fed in antiphase have the lowest bandwidth and efficiency.

621.396.67 3947
Earth Currents Near a Monopole Antenna with Symmetrical Top Loading—J. R. Wait. (*J. Res. Nat. Bur. Stand.*, vol. 62, pp. 247-255; June, 1959.) Expressions for the fields are developed for a vertical ground-based monopole with a cone or disk located at the top of the antenna to simulate umbrella top loading. The current distribution on the structure is assumed. Using spherical-wave functions, the magnetic-field distribution on the ground plane near the base of the antenna is computed and illustrated by graphs. For the case where the antenna is electrically small, the currents flowing on the cone or disk are shown to contribute only slightly to the total field.

621.396.67 3948
Earth Currents Near a Top-Loaded Monopole Antenna with Special Regard to Electrically Small L- and T-Antennas—H. L. Knudsen. (*J. Res. Nat. Bur. Stand.*, vol. 62, pp. 283-287; June, 1959.) A study of ground currents near a top-loaded monopole antenna with nonazimuthal symmetry as a sequel to the study by Wait of a similar antenna with symmetrical loading, (*see* 3947 above).

621.396.674-428 3949
The Equiangular Spiral Antenna—J. D. Dyson. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 181-187; April, 1959. Abstract, PROC. IRE, vol. 47, p. 1283; July, 1959.)

621.396.674.095 3950
On the Electromagnetic Radiation from a Vertical Dipole over the Surface of Arbitrary Surface Impedance—K. Furutsu. (*J. Radio Res. Labs., Japan*, vol. 6, pp. 269-291; April, 1959.) A mathematical study which includes both flat and spherical surfaces, and propagation across discontinuities in the surface impedance.

621.396.674.3:621.396.61:627.95 3951
Ferromagnetic Transmitter Antenna for Cases of Distress at Sea—Baur and Ziehm. (*See* 4223.)

621.396.677 3952
Super-Gain Antenna Beam—R. F. Kyle. (*Electronic Radio Eng.*, vol. 36, pp. 338-340; September, 1959.) A method is described for deducing the distribution of field on a cylinder of any prescribed radius from a given distribution on a cylinder of infinite radius.

621.396.677 3953
Multiplicative Receiving Arrays—V. G. Welsby and D. G. Tucker. (*J. Brit. IRE*, vol. 19, pp. 369-382; June, 1959.) The performance of arrays in which the outputs from two groups of elements or sections are multiplied together is studied from the points of view of directional patterns, directional discrimination and signal/noise performance. Whereas, in ordinary linear arrays, these three considerations are usually all determined by the directional pattern, yet in multiplicative arrays they are all quite distinct and often unreconcilable with one another. A comparison with super-directive arrays is also made.

621.396.677.029.64:621.372.83 3954
A Network for Combining Radio Systems at 4, 6 and 11 kMc/s—E. T. Harkless. (*Bell. Sys. Tech. J.*, vol. 38, pp. 1253-1267; September, 1959.) Development of an experimental coupling network described earlier by Dawson (2032 of 1957), with return loss 30 db, insertion loss 0.5 db and inter-port coupling 18-50 db.

621.396.677.31 3955
Directivity of a Broadside Array of Isotropic Radiators—H. E. King. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 197-198; April, 1959. Abstract, PROC. IRE, vol. 47, p. 1283; July, 1959.)

621.396.677.43 3956
The Influence of Gain and Current Attenuation on the Design of the Rhombic Antenna—R. P. Decker. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 188-196; April, 1959. Abstract, PROC. IRE, vol. 47, p. 1283; July, 1959.)

621.396.677.5 3957
Radiation from a Small Loop Immersed in a Semi-infinite Conducting Medium—J. R. Wait. (*Canad. J. Phys.*, vol. 37, pp. 672-674; May, 1959.) The radiation fields are given by a generalization of Sommerfield's solution for dipoles at the surface of the medium.

621.396.677.7 3958
Diffraction of Electromagnetic Waves at the Open End of a Helix Waveguide—P. S. Mikazan. (*Dokl. Akad. Nauk. SSSR*, vol. 128, pp. 502-505; September 21, 1959.) Formulas are derived for calculating the optimum dimensions of a helix waveguide for application as a radiator.

621.396.677.7 3959
Investigation of the Radiation Characteristics of Open Waveguides—K. E. Müller. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 35-42; September, 1958.) The aperture field is considered to be composed of the reflected wave due to mismatch added to the incident wave. The radiated field is calculated at $\lambda=3.2$ cm for the H_{10} mode in 23, 32, and 60 mm$\times$10mm waveguide and for the H_{30} mode in 60 mm$\times$10 mm waveguide. Calculations are also made for circular waveguide.

621.396.677.7 3960
Asymmetrical Trough-Waveguide Antennas—W. Rotman and A. A. Oliner. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 153-162; April, 1959. Abstract, PROC. IRE, vol. 47, p. 1282; July, 1959.)

621.396.677.75 3961
Microwave Antenna Saves Space—J. O. Pullman. (*Electronics*, vol. 32, pp. 54, 56; September 18, 1959.) Plexiglas plates in front of a small X-band horn reduce the beamwidth from about 70° to 40° and increase the gain by 3 db.

621.396.677.8:523.164 3962
The Ohio State University 360-ft Radio Telescope—J. D. Kraus. (*Nature*, vol. 184, pp. 669-672; August 29, 1959.) The telescope described consists of a fixed standing parabola and a flat reflector which may be tilted to deflect radiation into the parabola.

621.396.677.81+621.372.852.1 3963
Transmission of Electromagnetic Waves through Wire Gratings (Theory)—J. K. Skwirzynski and J. C. Thackray. (*Marconi Rev.*, vol. 22, pp. 77-90; 2nd Quarter, 1959.) Theory is provided for a general angle of incidence of a plane polarized em wave on a grating of cylindrical wires. Numerical results are given for normal incidence and possible application to design of waveguide filters is discussed.

621.396.677.81 3964
Transmission of Electromagnetic Waves through Wire Gratings (Experimental)—E. G. A. Goodall and J. A. C. Jackson. (*Marconi Rev.*, vol. 22, pp. 91-98; 2nd Quarter, 1959.) Comparison is made between experimental and theoretical values of power transmitted at normal incidence and at angles of incidence between 5° and 50°, through gratings of cylindrical wires of various diameters and spacings. For theory see 3963 above.

621.396.677.833 3965
Radiation from Ring Sources in the Presence of a Semi-infinite Cone—L. B. Felsen. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 168-180; April, 1959. Abstract, PROC. IRE, vol. 47, pp. 1282-1283; July, 1959.)

AUTOMATIC COMPUTERS

681.142 3966
The Characteristics of the Final Computer Model of the C.S.C.E. [Electronic Computer Study Centre] of Pisa from the Point of View of Mathematical Logic—A. Caracciolo and L. Guerri. (*Nuovo Cim.*, vol. 12, pp. 111-115; April 16, 1959.) A note of operating characteristics with special reference to the facility for modifying instructions by means of parametric cells. Other aspects of design and operation are dealt with in three short papers: *ibid.*, pp. 116-122, 123-125, 126-129.

681.142 3967
Using Inductive Control in Computer Circuits—W. M. Carey. (*Electronics*, vol. 32, pp. 31-33; September 18, 1959.) Various switching circuits are described in which transistors are used as the amplifying element and inductance as the timing element.

681.142:535.376.07 3968
Character Display System for Use as Digital

Computer Output—P. V. S. Rao. (*Rev. Sci. Instr.*, vol. 30, pp. 749–750; August. 1959.) Short description of a cr tube unit displaying numerals or letters and operating as an auxiliary device for a digital computer.

681.142:621.317.727 3969

How to Reduce Errors in Loaded Potentiometers—M. Kanner. (*Electronics*, vol. 32, pp. 34–35; August 21, 1959.) Compensation technique can reduce error by a factor of 100.

681.142:621.373.431.2:621.314.7 3970

Transistor Blocking Oscillator for Use in Digital Systems—A. A. Kaposi. (*Electronic Engrg.*, vol. 31, pp. 480–484; August, 1959.) A saturable transformer is used and pulse lengths from 1.5 μs upwards have been obtained. It is shown that the pulse length is independent of transistor parameters, temperature and load current.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.029.6 3971

New Microwave Systems using Low-Noise Devices—T. Maguire. (*Electronics*, vol. 32, pp. 27–30; August 21, 1959.) Review of a number of microwave techniques including maser and parametric amplifier operation and application, resonator bandwidth measurement, reduction of waveguide microphonics, ferrite phase shifters and the suppression of spurious signals in waveguides.

621.3.049.7 3972

Network Design of Microcircuits—C. K. Hager. (*Electronics*, vol. 32, pp. 44–49; September 4, 1959.) A detailed discussion of distr buted resistance and capacitance networks leading to new types of linear transfer functions.

621.319.43 3973

Gamma-Derived Capacitors—B. M. Oliver. (PROC. IRE, vol. 47, pp. 1654–1655; September, 1959.) Mathematical derivation of the rotor profile for a parallel-plate capacitor giving a linear frequency variation.

621.319.43:621.314.63 3974

Circuit Design using Silicon Capacitors—J. Hammerslag. (*Electronics*, vol. 32, pp. 48–50; September 18, 1959.) The capacitor is basically a semiconductor diode whose capacitance can be varied by the applied bias voltage. Operating characteristics and applications are described.

621.372.412 3975

The Coupling between Thickness-Shear and Flexural Vibrations of Circular AT-Cut Crystal Plates—R. Stark. (*Telefunken Ztg.*, vol. 31, pp. 179–187 and 232–239; September and December, 1958. English summary, pp 205–206 and 273.) Calculations are based on the theory of Mindlin and Deresiewicz (657 of 1955). Vibration patterns are derived which indicate a linear dependance of the resonance frequency on the sum of nodel semicircles and diameters. Results are verified experimentally using a multiple-beam interferometer.

621.372.5 3976

Simultaneous Approximation of the Modulus and Argument of a Transmission Factor by a Polynomial—E. Schuon. (*Frequenz*, vol. 12, pp. 318–323; October, 1958.) A polynomial is derived whose locus curve approximates, over a given range, the locus cur ve of the quadripole transfer function. An example is given.

621.372.543.2.029.4:621.375.232 3977

A Multirange, Low-Frequency Band-Pass Filter—J. B. Bratt. (*Electronic Engrg.*, vol. 31, pp. 458–462; August, 1959.) A frequency range of 16 cps to 175 cps is covered in nine bands with the response flat within ±5 per cent in the pass band.

621.372.63 3978

Synthesis of Active RC Networks—B. K. Kinariwala. (*Bell Sys. Tech. J.*, vol. 38, pp. 1269–1316; September, 1959.) Basic theory is developed to show that no more than one active element embedded in a passive RC network is needed to realize any driving-point function. A synthesis technique for n-port passive RC networks is developed to establish the sufficiency proof of the basic theorem. A more practical synthesis technique is also presented and applied to the design of a tenth-order Tchebycheff parameter filter.

621.373:621.316.729 3979

Transfer Function of an Oscillating System, Synchronised by an External Sinusoidal Signal—R. Beretze. (*Rev. HF, Brussels*, vol. 4, no. 6, pp. 125–136; 1959.) The amplitude and phase of forced oscillations within the range of synchronization are determined.

621.373.012:681.142 3980

Study of an Oscillator with Two Degrees of Freedom by a Differential Analyser—B. R. Nag. (*Indian J. Phys.*, vol. 33, pp. 57–73; February, 1959.)

621.373.029.6:621.372.44:621.314.63 3981

Semiconductor Diodes in Parametric Subharmonic Oscillators—J. Hilibrand and W. R. Beam. (*RCA. Rev.*, vol. 20, pp. 229–253; June, 1959.) The effects of stray capacitance, resistance, junction conductance and capacitance-voltage sensitivity are considered in the analysis of theoretical oscillator performance.

621.373.029.65 3982

Present State of the Millimetre-Wave Generation and Technique Art—1958—P. D. Coleman and R. C. Becker. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 42–61; January, 1959. Abstract, PROC. IRE, vol. 47, p. 498; March, 1959.)

621.373.029.65:538.221 3983

Millimetre-Wave Generation Experiment utilizing Ferrites—W. P. Ayres. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 62–65; January, 1959. Abstract, PROC. IRE, vol. 47, p. 498; March, 1959.)

621.373.4.029.63:621.372.822 3984

The Use of Open Rectangular Waveguides in R.F. Oscillator Circuits—J. J. M. Warringa and H. Piepers. (*Electronic Applic.*, vol. 19, pp. 84–92; May, 1959.) The theory of rectangular-type waveguide resonators is discussed and a 500-mc oscillator is described using two forced-air-cooled triodes and transmission-line tuning; radiation is suppressed by an open rectangular waveguide of small dimensions.

621.373.42 3985

Investigations of Two-Valve Oscillators in π-Network Form—E. Frisch. (*Arch. elekt. Übertragung*, vol. 12, pp. 401–406; September, 1958.) Two types of circuit are discussed in which feedback is applied via a π-network. One is independent of load input impedance and covers a wide range of frequencies; the other type has characteristics similar to those of differential-bridge oscillators but does not require a transformer.

621.373.42:621.317.74 3986

Oscillator Design using Voltage-Variable Capacitors—M. M. Brady. (*Electronics*, vol. 32, pp. 38–40; August 21, 1959.) A reverse biased p-n-junction diode is used as a variable capacitor in the design of a linear frequency-sweep generator. The output waveform is a sinusoid of high purity.

621.373.421.14.029.64 3987

The Generation of Stable Carrier Frequencies in the Range 3800–4200 Mc/s—R. J. D. Scarbrough, K. H. Ferguson, and A. W. Searls, (*P.O. elec. Engr. J.*, vol. 52, pp. 125–128; July, 1959.) Describes a microwave precision-cavity-controlled triode oscillator and a quartz-controlled carrier source suitable for incorporation in microwave communication systems.

621.373.431.1 3988

A Multivibrator with One Triode and with No Filament Current—M. R. E. Bichara. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2740–2742; May 11, 1959.) A circuit capable of producing waves rich in harmonics and similar to those of a multivibrator is described. A single triode is used and there is no heating current. The effect was observed by chance during tests of a super-regenerative receiver and is explained on the basis of a temperature cycle initiated by cutting off the filament current.

621.373.431.2 3989

Core-Saturation Blocking Oscillator Control—R. Spinrad. (*Rev. Sci. Instr.*, vol. 30, pp. 647–653; August, 1959.) The control of pulse duration by saturation of tube current or by saturation of the transformer core is discussed. The latter is shown to be preferable. An externally applied magnetic bias can control pulse width over a range >10:1. A particular circuit is described.

621.374.32 3990

A New Type of Differential Discriminator for Fast Pulses—C. Rubbia and G. Torelli. (*Nuovo Cim.*, vol. 12, pp. 144–147; April 16, 1959.) Description of a pulse-height analyzer based on a beam-deflection tube type E 80 T.

621.374.4:621.372.44:621.314.63 3991

Solid-State Generator for Microwave Power—M. M. Fortini and J. Vilms. (*Electronics*, vol. 32, pp. 42–43; September 4, 1959.) 10 mw at 2 kmc is produced by a capacitive-mode diode harmonic generator from a transistor oscillator and power amplifier delivering 153 mw at 250 mc.

621.375:621.376.2 3992

The Influence of Unsymmetric R.F. Stages on the Signal Band with Amplitude Modulation—R. Hofer. (*Arch. elekt. Übertragung*, vol. 12, pp. 381–393; September, 1958.) With the method of calculation given, sideband currents can be determined in nonlinear rf amplifier stages with amplitude-dependent source impedance even if the amplitude/frequency function of the load is unsymmetric.

621.375.018.75 3993

Linear Amplifier for Decimicrosecond Pulses—J. F. Golding and L. G. White. (*Electronic Radio Eng.*, vol. 36, pp. 323–327; September, 1959.) A system is described in which the bandwidth is increased by increasing the output power of the amplifier for the duration of the leading and trailing edges of the pulse.

621.375.1.024 3994

Transient Response of Direct-Current Amplifier Systems—R. D. Stuart. (*Brit. J. Appl. Phys.*, vol. 10, pp. 326–328; July, 1959.) A mathematical analysis shows that the signal/noise ratio of an amplifier with feedback may be improved by a filter, the design of which is independent of the feedback characteristics. See also 1594 of 1955 (Sanders).

621.375.121 3995

The Theory of Broadband Coupled Circuits—W. Dougharty. (*Electronic Engrg.*, vol. 31, pp. 488–494; August, 1959.) The exact transfer admittance and the phase response for various types of parallel-tuned coupled circuits are given. Various design formulas are derived which are sufficiently accurate for wide bandwidths.

621.375.2.024 3996

A Voltage-Operated Logarithmic Amplifier—R. F. Mathams. (*Electronic Engrg.*, vol. 31, pp. 463–465; August, 1959.) Input voltages from 0.1 to 100 v are chopped by a vibrating relay and differentiated. The time for the resulting exponential pulses to decay to a reference voltage is proportional to the logarithm of the input voltage.

621.375.3 3997

Shunt-Coupled Magnetic-Amplifier Circuits—R. M. Hubbard. (*Commun. and Electronics*, pp. 124–131; May, 1959.) Advantages can be obtained by using the average gate voltage directly as the output quantity.

621.375.3 3998

On the Feedback in Magnetic Amplifiers: Part 2—Combined Magnetic and Electric Feedbacks—L. A. Finzi and J. J. Suozzi. (*Commun. and Electronics*, pp. 136–141; May, 1959.) Consideration is given to a certain combination of positive magnetic and negative electric feedback. Part 1: 2172 of July.

621.375.4 3999

The Design Principles of Low-Frequency Small-Signal Amplifiers with Junction Transistors—J. Meinhardt. (*NachrTech.*, vol. 8, pp. 415–422; September, 1958.) Detailed exposition of design procedure based on fundamental transistor parameters.

621.375.4 4000

Stabilities of Common-Emitter and Emitter-Follower Transistor Amplifiers—B. D. Wedlock. (Proc. IRE, vol. 47, pp. 1657–1658; September, 1959.) Design equations for stability with respect to common-emitter current gain are developed and discussed.

621.375.4 4001

Design of Low-Noise Input Stages of L.F. Amplifiers using the Transistor Type OC 603—J. Schubert. (*Frequenz*, vol. 12, pp. 285–293 and 330–331; September and October, 1958.) The transistor noise/frequency characteristics are plotted with collector current as parameter, and formulas are given for circuit designs giving the minimum noise figure in the white-noise and $1/f$-noise regions of the frequency spectrum.

621.375.9:538.569.4.029.6 4002

A U.H.F. Solid-State Maser—R. H. Kingston. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 92–94; January, 1959. Abstract, Proc. IRE, vol. 47, p. 498; March, 1959.)

621.375.9:621.385.029.6:537.552 4003

Parametric and Pseudo-parametric Amplifiers—Clavier. (*See* 4217.)

621.375.9.029.6 4004

Solid-State Microwave Amplifiers—H. Heffner. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 83–91; January, 1959. Abstract, Proc. IRE, vol. 47, p. 498; March, 1959.)

621.375.9.029.62:621.3.011.23:621.314.63 4005

Analysis of a Four-Terminal Parametric Amplifier—K. K. N. Chang. (*RCA Rev.*, vol. 20, pp. 205–221; June, 1959.) See 1148 of April.

621.375.9.029.64:621.314.632 4006

An X-Band Parametric Amplifier—C. B. De Loach and W. M. Sharpless. (Proc. IRE, vol. 47, pp. 1664–1665; September, 1959.) Preliminary performance figures are given for an amplifier operating in the degenerate mode about a center frequency of 11.55 kmc, using point-contact GaAs diodes [1710 of May (Sharpless)].

GENERAL PHYSICS

530.162:519:061.3 4007

Fluctuation Phenomena and Stochastic Processes—C. Domb. (*Nature*, vol. 184, pp. 509–512; August 15, 1959.) A report of the Physical Society conference held in London, March 19 and 20, 1959.

534.2+538.566]:517.941.9 4008

On the Analogy of Acoustic and Electromagnetic Boundary-Value Problems—H. Severin. (*Akust. Beihefte*, no. 1, pp. 270–274; 1959.) Conditions for mathematical analogy are formulated and examples of complete and partial analogy are given.

534.2:537.3 4009

Experiments on the Acousto-electric Effect—P. Smith and D. O. Sproule. (*Nature*, vol. 184, suppl. no. 5, p. 264; 1959.) Experiments with copper wire at 25 kc produced a potential difference very much less than that predicted by theory [2189 of July (Parmenter)]. No effect was observed in aluminum.

535–15 4010

Infrared Physics and Technology—(Proc. IRE, vol. 47, pp. 1415–1649; September, 1959.) This issue is devoted to a review of the infrared field. The material has been prepared under the auspices of the Infrared Information Symposia (IRIS) and is grouped in six sections. Titles of some of the papers are given below; certain others are abstracted individually.

a) **Selective Radiators**—G. N. Plass (pp. 1442–1447).

b) **The Transmission of the Atmosphere in the Infrared**—J. N. Howard (pp. 1451–1457).

c) **Fundamentals of Infrared Detectors**—R. L. Petritz. (pp. 1458–1467).

d) **Infrared Photoemission**—G. A. Morton (pp. 1467–1469).

e) **Range Equation for Passive Infrared Devices**—L. Larmore (pp. 1489–1490).

f) **Range Equation for Active Devices**—K. V. Knight (pp. 1490–1492).

g) **Description and Properties of Various Thermal Detectors**—R. De Waard and E. Wormser (pp. 1508–1513).

h) **Cooling Techniques for Infrared Detectors**—J. G. Goodenough (pp. 1514–1515).

i) **Lumped-Parameter Behaviour of the Single-Stage Thermoelectric Microrefrigerator**—M. B. Grier (pp. 1515–1518).

j) **Preamplifiers for Non-image-forming Infrared Systems**—J. A. Jamieson (pp. 1522–1523).

k) **Optical-Mechanical Scanning Techniques**—M. R. Holter and W. L. Wolfe (pp. 1546–1550).

l) **Combined Optical-Microwave System Considerations**—S. H. Cushner (pp. 1553–1554).

m) **Methods of Background Description and their Utility**—D. Z. Robinson (pp. 1554–1561).

n) **Special Electronic Circuits for Non-image-forming Infrared System**—J. A. Jamieson (pp. 1570–1572).

o) **Infrared Colour Translation**—M. B. Grier (pp. 1574–1576).

537.122:539.2:537.228 4011

Nuclear Polarization via "Hot" Conduction Electrons—G. Feher. (*Phys. Rev. Lett.*, vol. 3, pp. 135–137; August 1, 1959.) Describes a polarization scheme based on the difference between the electron spin temperature and the temperature corresponding to the mean kinetic energy of the electrons accelerated in a dc electric field.

537.2 4012

The Calculation of the Electric Potential and the Capacity of a Tore by means of Toroidal Functions—S. C. Loh. (*Canad. J. Phys.*, vol. 37, pp. 698–702; June, 1959.) See also 2193 of July (Bouwkamp).

537.221 4013

The Generation of Static Charge on High Polymer—S. Kittaka. (*J. Phys. Soc. Japan*, vol. 14, pp. 532–538; April, 1959.) The generation of static electricity has been studied by the contact and gradual separation of three high-polymer substances from a platinum rod. Results show how the static charges generated vary with the oxygen content of the air and with relative humidity.

537.311.1 4014

The General Theory of Transport: the Difference between Electric Field and Density Gradient—T. Kasuya. (*J. Phys. Soc. Japan*, vol. 14, pp. 410–415; April, 1959.) The difference between the electric currents due to an external electric field and to the chemical potential gradient (see 1223 of April) is discussed for the limiting cases of weak and strong applied magnetic fields.

537.311.1 4015

Electrical and Thermal Conductivity of Monovalent Metals; The Influences of Coulomb Interaction—T. Kasuya and K. Yamada. (*J. Phys. Soc. Japan*, vol. 14, pp. 416–435; April, 1959.)

537.5 4016

Propagation of Non-ionizing Shock Waves Produced by a Capacitor Discharge in a Tube of Gas—R. der Agobian and L. Lifschitz. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2734–2736; May 11, 1959.) The method of detection of shock waves described earlier [3447 of October (der Agobian)] is applied.

537.533:621.384.6 4017

Behaviour of Intense Relativistic Electron Beams—R. H. Capps. (*Phys. Rev.*, vol. 114, pp. 1203–1217; June 1, 1959.) A theoretical investigation.

537.533.7:538.561 4018

Optical Transition Radiation from Protons Entering Metal Surfaces—P. Goldsmith and J. V. Jelley. (*Phil. Mag.*, vol. 4, pp. 836–844; July, 1959.) Experiments to establish the existence of the transition radiation effect [3763 of 1958 (Garibyan)] are described. The results obtained from studies of the polarization of the radiation, its excitation function and its absolute yield confirm the predictions of theory.

537.54:621.396.822 4019

Noise Spectra of a Probe in a Hot-Cathode Discharge—C. Singh. (*Proc. Phys. Soc.* (*London*), vol. 74, pp. 42–47; July 1, 1959.) Measurements with floating and biased probes of low-frequency noise in low-pressure mercury-vapor discharge tubes reveal a continuous and smooth frequency spectrum similar to that of the tube noise, and indicate that the noise source does not lie in the primary electron emission.

537.56 4020

On the Coefficients of Irreversible Processes in a Highly Ionized Gas—T. Kihara. (*J. Phys. Soc. Japan*, vol. 14, pp. 402–410; April, 1959.)

537.56:538.566 4021

Improvements in the Microwave Propagation Method of Studying Decaying Gas Plasmas—S. Takeda and E. H. Holt. (*Rev. Sci. Instr.*, vol. 30, pp. 722–725; August, 1959.)

537.56:538.569 4022
Traveling-Wave Focusing for Plasma Containment—C. K. Birdsall and A. J. Lichtenberg. (*Phys. Rev. Lett.*, vol. 3, pp. 163–164; August 15, 1959.) The bunches of charged particles accompanying a traveling wave can produce focusing of relatively stationary charged particles.

538.12:621.385.833 4023
Calculation of Fields. Elementary Solutions associated with the Flux Function across a Circle on an Axis *Oz*. Analytical Expressions for the Induction and its Derivatives on *Oz*—P. Gautier and M. Laudet. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2737–2739; May 11, 1959.) Analytical expressions that may be used in numerical calculations are obtained.

538.23:538.221 4024
Combined Action of Random Creep and Thermal-Fluctuation Fields—L. Néel. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2676–2681; May 11, 1959.) The different constants characteristic of creep may be calculated from measurements of the differential susceptibility made after some creep cycles have been described. See 3109 and 3110 of 1957.

538.56 4025
Electromagnetic Wave Problems—J. D. Lawson. (*Electronic Radio Eng.*, vol. 36, pp. 332–338; September, 1959.) Common features of wave and radiation problems encountered in physics and electrical engineering are discussed in terms of physical ideas obtained from the study of the two forms of the simple em plane wave in a loss-free medium.

538.561.029.6:538.221 4026
Microwave Radiation from Ferrimagnetically Coupled Electrons in Transient Magnetic Fields—F. R. Morgenthaler. (IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 6–11; January, 1959. Abstract, PROC. IRE, vol. 47, p. 497; March, 1949.)

538.566:535.42]+534.26 4027
Field Fluctuations near the Focus of a Lens—M. N. Krom. (*Akust. Zh.*, vol. 5, no. 1, pp. 45–50; 1959.) Mathematical analysis of the distribution of fluctuations near the focus of a large lens and the dependence of the fluctuations on the dimensions of the objective.

538.566:535.42 4028
Diffraction by an Imperfectly Conducting Wedge—T. B. A. Senior. (*Commun. Pure Appl. Math.*, vol. 12, pp. 337–372; May, 1959.) The problem of an em wave incident upon a wedge whose conductivity is large but not infinite is considered.

538.566:535.43 4029
On Scattering by Large Conducting Bodies—R. F. Harrington. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 150–153; April, 1959. Abstract, PROC. IRE, vol. 47, p. 1282; July, 1959.)

538.566:537.56 4030
Growth of Electron Space-Charge and Radio Waves in Moving Ion Streams—R. Q. Twiss. (*Phil. Mag.*, vol. 4, pp. 868–875; July, 1959.) A criticism of a recent paper by Piddington (3267 of October). His belief that the growing waves in a two-stream electron wave tube are really evanescent is claimed to be untenable. Piddington's theory is stated to be effectively equivalent to the conventional one in which these growing waves are identified as being true amplified waves. Editorial note, *ibid.*, p. 876.

538.566.2 4031
Dependence of the Pulsation Frequency of a Field at the Focus of a Lens on the Dimensions of the Diaphragm—E. A. Blyakhman and L. A. Chernov. (*Akust. Zh.*, vol. 5, no. 1, pp. 21–24; 1959.) Theoretical investigation of the influence of a focusing system, such as a lens-type receiving antenna, on the spectral characteristics of the field pulsations at the focus, for the case of incident waves which have traversed a medium containing random inhomogeneities in chaotic motion.

538.569.4.029.64 4032
Microwave Zeeman Spectrum of Atomic Oxygen—H. E. Radford and V. W. Hughes. (*Phys. Rev.*, vol. 114, pp. 1274–1279; June 1, 1959.) Measurements using the technique of paramagnetic resonance absorption.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.164:621.396.677.8 4033
The Ohio State University 360-ft Radio Telescope—Kraus. (*See* 3962.)

523.164.3 4034
On a Feature of Galactic Radio Emission—H. M. Johnson. (*Phil. Mag.*, vol. 4, p. 877; July, 1959.) Comment on a paper by Tunmer (1176 of April).

523.164.32 4035
The Detection of Coherent Harmonics in certain Solar Outbursts—R. C. Jennison. (*Observatory*, vol. 79, pp. 111–113; June, 1959.) An interferometer system is described which was used at 127 and 254 mc to investigate the origin of harmonics in solar radio outbursts [see 704 of 1955 (Wild *et al.*)]. The outbursts producing coherent harmonics appeared to be of type II.

523.165 4036
Change of Cosmic Rays in Space—H. V. Neher. (*Nature*, vol. 184, pp. 423–425; August 8, 1959.) A review of data on the variation of absolute intensity with time during the period 1954–1958, derived from balloon and rocket measurements.

523.165 4037
Transient Decreases in Cosmic-Ray Intensity during the Period October 1956 to January 1958—A. G. Fenton, K. G. McCracken, D. C. Rose and B. G. Wilson. (*Canad. J. Phys.*, vol. 37, pp. 569–578; May, 1959.)

523.165 4038
Cause of the Minimum in the Earth's Radiation Belt—S. F. Singer. (*Phys. Rev. Lett.*, vol. 3, pp. 188–190; August 15, 1959.) The inner Van Allen belt is attributed to secondary cosmic rays formed in the earth's atmosphere. The calculated density as a function of distance from the earth agrees with the observed minimum between this belt and the outer "solar particle" belt.

523.165 4039
The Upper Boundary of the Van Allen Radiation Belts—C. W. Snyder. (*Nature*, vol. 184, suppl. no. 7, pp. 439–440; August 8, 1959.) A re-examination of telemetry records for Pioneer III indicates that the counting rate dropped to 1.4 counts per second at a range of 65,000 km from the earth's center. These and other radiation intensity measurements suggest that the upper boundary of the radiation belt is at a range of 9–10 earth radii.

523.165:523.75 4040
Protons from the Sun on May 12, 1949—E. P. Ney, J. R. Winckler and P. S. Freier. (*Phys. Rev. Lett.*, vol. 3, pp. 183–185; August 15, 1959.) The integral flux of particles at the top of the atmosphere measured above Minneapolis, Minn., increased by a factor of 1000 above that of cosmic rays.

523.165:550.389.2:629.19 4041
Radiation Measurements to 658300 km with Pioneer IV—J. A. Van Allen and L. A. Frank. (*Nature*, vol. 184, pp. 219–224; July 25, 1959.) Radiation observations with Pioneer IV are compared with those obtained with Pioneer III (2553 of August) and the Soviet cosmic rocket, and the results are discussed.

523.165:550.389.2:629.19 4042
Radiation Observations with Satellite 1958 δ over Australia—A. J. Herz, K. W. Ogilvie, J. Olley and R. B. White. (*Nature*, vol. 184, pp. 391–395; August 8, 1959.) An analysis of data obtained from observations made during transits in July and August, 1958 indicates a minimum of radiation intensity at a latitude of 35°S. The intensity appears to increase by an order of magnitude for latitude changes of ±10° about this minimum [see also 2238 of July (Van Allen *et al.*)]. There is evidence to suggest a time variation in the radiation intensity. The characteristics of the signal and the method of analysis of records are described.

523.75 4043
Magnetic Field associated with a Great Solar Flare—R. Howard, T. Cragg and H. W. Babcock. (*Nature*, vol. 184, suppl. no. 6, p. 351; August 1, 1959.) Magnetograms obtained at Mount Wilson, Calif., during an unusually large solar flare on July 16, 1959, show no definite change in the magnetic-field pattern of the photosphere.

550.37:550.385 4044
Correlation between Earth-Current and Geomagnetic Disturbance—V. P. Hessler and E. M. Wescott. (*Nature*, vol. 184, suppl. no. 9, p. 627; August 22, 1959.) The high correlation coefficients obtained from an analysis of data for recent years suggest that earth currents may be used interchangeably with magnetic disturbances on an indicator of ionospheric activity.

550.385.37 4045
Micropulsation Measurements in California and Alaska—W. H. Campbell and B. Nebel. (*Nature*, vol. 184, suppl. no. 9, p. 628; August 22, 1959.) Large night-time storms in Alaska gave oscillations ten or fifteen times larger than in California. Great micropulsation activity in Alaska was accompanied by sw blackouts.

550.389.2 4046
Preliminary Results of the National Bureau of Standards Radio and Ionospheric Observations during the International Geophysical Year—D. M. Gates. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 1–14; July/August, 1959.)

550.389.2:629.19 4047
Determination of the Orbit of an Artificial Satellite—J. T. Anderson. (PROC. IRE, vol. 47, pp. 1658–1659; September, 1959.) A critical discussion of various Doppler tracking systems.

550.389.2:629.19 4048
Determination of the Earth's Gravitational Potential from Observations on Sputnik II (1957 β)—A. H. Cook. (*Geophys. J. R. Astr. Soc.*, vol. 1, pp. 341–345; December, 1958.)

550.389.2:629.19 4049
Signals from Satellite 1958 δ2 (Sputnik III)—B. G. Pressey. (*Nature*, vol. 184, suppl. no. 5, p. 261; July 25, 1959.) A note on lapses in the reception of the cw signal at Slough and at Singapore, corresponding to the lapse of modulation noted by Munro (2954 of September) when the satellite was not in sunlight.

550.389.2:629.19 4050
Solar Effects in the Motion of *Vanguard*—

S. P. Wyatt. (*Nature*, vol. 184, suppl. no. 6, pp. 351–352; August 1, 1959.) An analysis of period changes of satellite 1958 β2 shows a correlation with three solar effects: a) the hour angle of the Sun as reckoned from the perigee point of the orbit; b) the 27-day variations in solar activity; c) the total daily solar insolation at the latitude of perigee.

550.389.2:629.19 4051
Observations of the Russian Moon Rocket Lunik II—J. G. Davies and A. C. B. Lovell. (*Nature*, vol. 184, pp. 501–502; August 15, 1959.) A report of observations made at Jodrell Bank on the evenings of September 12 and 13, 1959, on frequencies of 183.6 mc and 19.992 mc. The signals on both frequencies ceased at 21h 02m 23s U.T. on September 13. A plot of the variations in frequency during the last hour of the rocket's flight is given and deductions are made from it.

550.389.2:629.19:551.594.5 4052
A New Method for Studying the Auroral Ionosphere using Earth Satellites—R. Parthasarathy, R. P. Basler and R. N. De Witt. (Proc. IRE, vol. 47, p. 1660; September, 1959.) Simultaneous recordings are taken at two stations, 19 km apart, of the field strength of the 20-mc transmissions from Sputnik III as it passes directly overhead. The results are used to study the structure of the absorption region.

551.510.52:621.396.11 4053
Climatology of Ground-Based Radio Ducts—B. R. Bean. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 29–34; July/August, 1959.) Several years of radiosonde observations at stations typical of arctic, temperate and tropical climates are used to examine the variations with climate of the frequency of occurrence, heights, thickness and refractivity gradients of atmospheric ducts.

551.510.535 4054
Size of Irregularities in the E Region of the Ionosphere—M. S. Rao. (*Canad. J. Phys.*, vol. 37, pp. 557–568; May, 1959.) An empirical relation $V=KN\lambda$, is deduced statistically to relate the velocity V, fading frequency N, and wavelength λ in observations of E-region drifts using the three-station fading record method. The constant K is compared for three stations in the northern hemisphere and used to derive estimates of the irregularity sizes. The variation of these sizes with latitude is discussed.

551.510.535 4055
On the Seasonal and Nonseasonal Annual Variations and the Semi-annual Variation in the Noon and Midnight Electron Densities of the F_2 Layer in Middle Latitudes—T. Yonezawa and Y. Arima. (*J. Radio Res. Labs, Japan*, vol. 6, pp. 293–309; April, 1959.) The basic data from which the various components are derived are the critical frequencies at nine northern and eight southern latitude stations. The results are shown graphically for low and high solar activity.

551.510.535:539.16 4056
Geophysical Effects of High-Altitude Nuclear Explosions—W. A. Feibelman. (*Nature*, vol. 184, suppl. no. 7, p. 442; August 8, 1959.) Report of an enhancement recorded at Pittsburgh, Pa., on a frequency of 27 kc, between 1200 and 1300 U.T. on August 12, 1958, an hour later than a similar enhancement recorded in Japan by Obayashi *et al.* (2969 of September).

551.510.535:550.385 4057
Upper Atmosphere Density Variations due to Hydromagnetic Heating—A. J. Dessler. (*Nature*, vol. 184, suppl. no. 5, pp. 261–262; July 25, 1959.) A mechanism of ionospheric heating by hydromagnetic waves generated by interactions between the solar wind and the geomagnetic field is proposed to explain a) irregular orbital accelerations of satellites, b) the sudden disappearance of trapped radiation from the Argus nuclear explosion coincident with a geomagnetic storm and c) X-ray flux at balloon altitudes during magnetic storms. See 2585 of August.

551.510.535:551.55:523.5 4058
Measurements of Turbulence in the Upper Atmosphere—J. S. Greenhow and E. L. Neufeld. (*Proc. Phys. Soc.*, vol. 74, pp. 1–10; July 1, 1959. Plate.) An account of investigations of eddy size at heights of 80–100 km based on the correlation between simultaneous measurements of wind velocities at two points along a meteor trail.

551.510.535:621.396.11 4059
A Theory of Ionospheric Radio Wave Scattering under the Influences of Ion Production and Recombination—K. Maeda, S. Kato and T. Tsuda. (*J. Geomag. Geoelect.*, vol. 10, no. 3, pp. 91–98; 1959.) The basis of the analysis is the pressure theory developed by Villars and Weisskopf (2747 of 1954) but the additional influences of electron production and recombination have been taken into account. The analysis is limited to the E layer and the results obtained account for the dependence of scattered signal intensity on both frequency and scattering angle.

551.510.535:621.396.11 4060
Ionospheric Radio Wave Scattering in the Electrodynamically Controlled Turbulence—K. Maeda, S. Kato and T. Tsuda. (*J. Geomag. Geolect.*, vol. 10, pp. 126–130; 1959.) An analytical study of turbulence in the ionosphere due to em forces rather than hydrodynamic ones. Numerical examples are given for the E layer at 90 km.

551.510.535:621.396.11.029.63 4061
Vertical-Incidence Scatter from the Ionosphere—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 43, p. 157; August, 1959.) Preliminary experiments using radar techniques at 41 mc show that measurements of temperature and electron-density at heights up to 400 miles should be possible.

551.594.5 4062
Auroral Frequency Lines—J. M. Stagg and B. Hultqvist. (*Nature*, vol. 184, suppl. no. 5, pp. 262–263; July 25, 1959.) Comment on 2975 of September and author's reply.

551.594.5:523.165 4063
The Aurora, the Radiation Belt and the Solar Wind: a Unifying Hypothesis—M. H. Rees and G. C. Reid. (*Nature*, vol. 184, suppl. no. 8, pp. 539–540; August 15, 1959.)

551.594.5:621.396.96 4064
Polarization of Radar Echoes from Aurora—A. Kavadas and D. G. Glass. (*Canad. J. Phys.*, vol. 37, pp. 690–697; June, 1959.) A description of the equipment and experimental procedure for investigating the influence of the earth's magnetic field on the polarization of a 48.2-mc signal reflected from aurora. Results indicate that the received wave contains in addition to an unpolarized component, a linear component of polarization tilted in the general direction of the earth's magnetic field.

551.594.5:621.396.96 4065
Preliminary Results of 400-Mc/s Radar Investigations of Auroral Echoes at College, Alaska—R. L. Leadabrand, L. Dolphin and A. M. Peterson. (IRE Trans. on Antennas and Propagation, vol. AP-7, pp. 127–136; April, 1959. Abstract, Proc. IRE, vol. 47, p. 1282; July, 1959.) See also 1430 of 1958 (Fricker *et al.*).

551.594.6 4066
Origin of "Very-Low-Frequency Emissions"—R. M. Gallet and R. A. Helliwell. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 21–27; July/August, 1959.) A theory is outlined, based on selective traveling-wave amplification in the outer ionosphere, to account for vlf emissions such as hiss, quasi-constant tones, transients, and very long trains of whistler echoes.

551.594.6:539.16 4067
Effect of Atomic Tests on Radio Noise—C. A. Samson. (*Nature*, vol. 184, suppl. no. 8, pp. 538–539; August 15, 1959.) Preliminary report of noise measurements made on six spot frequencies between 13 kc and 5 mc at a station 700 miles from the site of two atomic explosions. Results suggest that high-altitude explosions may have a persistent effect on radio communications at certain frequencies.

LOCATIONS AND AIDS TO NAVIGATION

621.396.663 4068
True-Phase Capacitive Goniometers—G. Ziehm. (*Frequenz*, vol. 12, pp. 293–299; September, 1958.) Methods are discussed for ensuring that the coupling capacitance between rotor and stator in goniometers for vhf direction-finders is a sinusoidal function of the angle of rotation. Design data are tabulated relating the number of rotor elements to the order of harmonics. The compensation of harmonics by the use of a second, displaced rotor is also described.

621.396.933.25:681.188 4069
An Electronic Clock Coder for Coded Radio Beacons—J. W. Nichols, A. C. MacKellar and A. J. B. Baty. (*Electronic Engrg*, vol. 31, pp. 466–474; August, 1959.) Marine radio beacons operating on a time-sharing basis use a crystal-controlled clock and electronic coder.

621.396.96:535–15 4070
Infrared Search-System Range Performance—R. H. Genoud. (Proc. IRE, vol. 47, pp. 1581–1586; September, 1959.)

621.396.96:535–15 4071
Simulation of Infrared Systems—H. F. Meissinger. (Proc. IRE, vol. 47, pp. 1586–1592; September, 1959.)

621.396.968 4072
Effect of Surface Reflections on Rain Cancellation of Circularly Polarized Radars—R. McFee and T. M. Maher. (IRE Trans. on Antennas and Propagation, vol. AP-7, pp. 199–201; April, 1959.) Calculations based on the reflection coefficients of sea water at 500 and 3000 mc indicate that only a moderate degree of cancellation of returns from rain can be achieved in search radars overlooking the sea or flat land.

621.396.969.36 4073
The Nose-On Radar Cross-Sections of Finite Cones—H. Brysk, R. E. Hiatt, V. H. Weston and K. M. Siegel. (*Canad. J. Phys.*, vol. 37, pp. 675–679; May, 1959.) An attempt to resolve theoretical difficulties between Keys and Primich (3320 of October) and Siegel (*Appl. Sci. Res.*, vol. B7, No. 4, pp. 293–328; 1958) for both the Rayleigh and resonance regions. Some measurements are described which confirm the theoretical discussion of Siegel.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.583:621.385.032.14 4074
Titanium as a Gettering Material—R. L. Stow. (*Nature*, vol. 184, suppl. no. 8, pp. 542–

543; August 15, 1959.) The ultimate low pressure obtainable with Ti is at least 40 times lower than that obtained with Ba.

535.215 4075

The Selective Photoelectric Effect—W. T. Doyle. (*Proc. Phys. Soc.*, vol. 74, pp. 27–32; July 1, 1959.) Theoretically derived wavelengths for peak photoemission from alkali metals are compared with observed values.

535.215–15 4076

Film-Type Infrared Photoconductors—R. J. Cashman. (Proc. IRE, vol. 47, pp. 1471–1475; September, 1959.) A review of the properties of film-type photoconductors.

535.215:546.482.21 4077

High-Pressure, High-Temperature Growth of Cadmium Sulphide Crystals—W. E. Medcalf and R. H. Fahring. (*J. Electrochem. Soc.*, vol. 105, pp. 719–723; December, 1958.)

535.215:537.311.33:621.317.3 4078

Test Equipment for Photosensitive Semiconductors—R. Gereth and H. A. Müser. (*Z. angew. Phys.*, vol. 10, pp. 419–424; September, 1958.) Sensitivity and noise characteristics of photoconductive devices can be determined with the equipment described. The illuminating beam is mechanically modulated in the frequency range 50 cps–8 kc and a noise-cancelling circuit is used. Results are given of tests on a PbS photocell.

535.215:539.23:546.817.221 4079

The Optical Properties of Photoelectric Thin Films. Case of Microcrystalline Films of Lead Sulphide—V. Schwetzoff. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2859–2861; May 20, 1959.)

535.37:546.472.21 4080

Time-Dependent Spectra of Electroluminescent ZnS:Cu, Pb—J. Weiszburg, J. Schanda and Z. Bodó. (*Phil. Mag.*, vol. 4, pp. 830–832; July, 1959.)

535.37:546.472.21 4081

Energy Transfer in ZnS:Cu:In Phosphors—N. T. Melamed. (*J. Phys. Chem. Solids*, vol. 9, pp. 149–152; February, 1959.) Results of observations on the transport of energy between luminescence centers support a photoconductive transfer mechanism. For a report of investigations of energy transfer in ZnS:Cu:Cl phosphors and a description of the apparatus used to obtain emission and excitation spectra in both experiments see *J. Phys. Chem. Solids*, vol. 7, pp. 146–152; November, 1958.

535.376 4082

Electroluminescence of AlN—G. A. Wolff, I. Adams and J. W. Mellichamp. (*Phys. Rev.*, vol. 114, pp. 1262–1264; June 1, 1959.) Experimental results are described and an explanation is given of the mechanism involved.

537.226/.227:546.431.824–31:539.23 4083

Some Properties of Thin Single-Crystal Films of Barium-Strontium Titanate Vapour-Deposited in an Electric Field—A. Moll. (*Z. angew. Phys.*, vol. 10, pp. 410–416; September, 1958.) Two methods of obtaining thin single-crystal films of (Ba, Sr) TiO_3 from the vapor phase of a mixture or solution of titanates are described. The dielectric constant of films is plotted as a function of thickness in the frequency range of 1–2 kmc.

537.227 4084

$LiH_3(SeO_3)_2$: New Room-Temperature Ferroelectric—R. Pepinsky and K. Vedam. (*Phys. Rev.*, vol. 114, pp. 1217–1218; June 1, 1959.) Results of X-ray crystallography and dielectric and thermal measurements are reported.

537.227 4085

Powder-Pattern Techniques for Delineating Ferroelectric Domain Structures—G. L. Pearson and W. L. Feldmann. (*J. Phys. Chem. Solids*, vol. 9, pp. 28–30; January, 1959.) Colloidal suspensions in insulating organic liquids have been found which produce powder patterns on polarized domains in ferroelectric crystals.

537.228.1 4086

Piezoelectric Ceramics—H. Jaffe. (*J. Amer. Ceram. Soc.*, vol. 41, pp. 494–498; November, 1958.) A survey, with data on lead metaniobate and lead titanate zirconate compositions.

537.311.33:536.2:530.17 4087

An Electrical Analogue for Heat Flow Problems in Semiconductors—N. L. Potter. (*Electronic Engrg.*, vol. 31, pp. 454–457; August, 1959.) Current flow in a RC network is used as an analog for the study of junction temperature.

537.311.33:538.614 4088

The Faraday Effect in Semiconductors—M. J. Stephens and A. B. Lidard. (*J. Phys. Chem. Solids*, vol. 9, pp. 43–47; January, 1959.) A general expression for the Faraday rotation, θ, in a semiconductor is derived and applied to the case of free carriers. Explicit expressions are given for particular cases, and it is shown how measurements of θ at high frequencies lead to information about energy surfaces.

537.311.33:539.12.04 4089

Conference on Radiation Effects in Semiconductors—(*J. Appl. Phys.*, vol. 30, pp. 1117–1322; August, 1959.) The text is given of the following papers which were included among those presented at the Conference sponsored by the Oak Ridge National Laboratory and held in Gatlinburg, Tennessee, May 6–9, 1959. Certain other papers are abstracted separately.

a) **Some Consequences of Thermal Neutron Capture in Silicon and Germanium**—H. C. Schweinler (pp. 1125–1126).

b) **Infrared Absorption and Photoconductivity in Irradiated Silicon**—H. Y. Fan and A. K. Ramdas (pp. 1127–1134.)

c) **Mechanism and Defect Responsible for Edge Emission in CdS**—R. J. Collins (pp. 1135–1140).

d) **Diffusion-Controlled Reactions in Solids**—H. Reiss (pp. 1141–1152).

e) **Radiation Effects in Semiconductors: Thermal Conductivity and Thermoelectric Power**—T. H. Geballe (pp. 1153–1157).

f) **Transport Properties in Silicon and Gallium Arsenide**—R. K. Willardson (pp. 1158–1165).

g) **Paramagnetic Resonance in Electron-Irradiated Silicon**—G. Bemski (pp. 1195–1198).

h) **Spin Resonance in Electron-Irradiated Silicon**—G. D. Watkins, J. W. Corbett and R. M. Walker (pp. 1198–1203).

i) **Disordered Regions in Semiconductors Bombarded by Fast Neutrons**—B. R. Gossick (pp. 1214–1218).

j) **Radiation-Induced Energy Levels in Silicon**—C. A. Klein (pp. 1222–1231).

k) **Electron-Bombardment Damage in Oxygen-Free Silicon**—G. K. Wertheim and D. N. E. Buchanan (pp. 1232–1234).

l) **High-Energy Electron Irradiation of Germanium and Tellurium**—V. A. J. van Lint and H. Roth (pp. 1235–1238).

m) **Radiation-Produced Energy Levels in Compound Semiconductors**—L. W. Aukerman (pp. 1239–1243).

n) **Precipitation in Semiconductors**—A. G. Tweet (pp. 1244–1248).

o) **Annealing of Radiation Defects in Semiconductors**—W. L. Brown, W. M. Augustyniak and T. R. Waite (pp. 1258–1268).

p) **Low-Temperature Annealing Studies in Ge**—J. W. Mackay and F. E. Klontz (pp. 1269–1274).

q) **Electron-Microscope Studies on the Etching of Irradiated Germanium**—T. S. Noggle and J. O. Stiegler (pp. 1279–1288).

r) **Irradiation Damage in Germanium and Silicon due to Electrons and Gamma Rays**—J. H. Cahn (pp. 1310–1316).

Discussion (pp. 1317–1322).

537.311.33:539.12.04 4090

Recombination Properties of Bombardment Defects in Semiconductors—G. K. Wortheim. (*J. Appl. Phys.*, vol. 30, pp. 1166–1174; August, 1959.) Emphasizes those aspects which complicate interpretation of lifetime data, such as the inherent difference between steady-state and transient measurements, large-signal behavior, competing recombination mechanisms, trapping, the possible existence of strongly temperature-dependent cross-sections, and the properties of multilevel defects.

537.311.33:539.12.04 4091

Nature of Bombardment Damage and Energy Levels in Semiconductors—J. H. Crawford, Jr. and J. W. Cleland. (*J. Appl. Phys.*, vol. 30, pp. 1204–1213; August, 1959.) The different effects of Co^{60} gamma-ray and fast-neutron bombardment on the electrical behavior of Ge are discussed in terms of different local distributions of lattice defects expected for these two types of radiation.

537.311.33:539.23:537.533.7 4092

Electron Characteristic Energy Losses in some Intermetallic Compounds—B. Gauthé. (*Phys. Rev.* vol. 114, pp. 1265–1268; June 1, 1959.) Experimental results and their interpretation.

537.311.33:541.135 4093

Saturation Currents in Germanium and Silicon Electrodes—J. B. Flynn. (*J. Electrochem. Soc.*, vol. 105, pp. 715–718; December, 1958.) Report of an experimental investigation of the contribution of surface generation to the saturation current. With n-type Ge anodes in 1.35 N KOH solution a current multiplication factor of about 4 was obtained.

537.311.33:[546.28+546.289 4094

On the Temperature Dependence of the Distribution Coefficient: the Solid Solubilities of Tin in Silicon and Germanium—F. A. Trumbore, C. R. Isenberg and E. M. Porbansky. (*J. Phys. Chem. Solids*, vol. 9, pp. 60–69; January, 1959.)

537.311.33:546.28 4095

Statistical Theory of Avalanche Breakdown in Silicon—R. E. Burgess. (*Canad. J. Phys.*, vol. 37, pp. 730–738; June, 1959.) A consideration of the properties of avalanche diodes that effectively contain a random switch in series with a resistive component which may be linear or nonlinear but which always has a positive differential resistance.

537.311.33:546.28 4096

Diffusion Control in Silicon by Carrier Gas Composition—C. J. Frosch and L. Derick. (*J. Electrochem. Soc.*, vol. 105, pp. 695–699; December, 1958.) Description of a single heating process for producing controlled single and double diffusion layers using a modified form of the apparatus described earlier (1182 of 1958).

537.311.33:546.28 4097

Effects of certain Chemical Treatments and Ambient Atmospheres on Surface Properties of Silicon—T. M. Buck and F. S. McKim. (*J. Electrochem. Soc.*, vol. 105, pp. 709–714; December, 1958.)

537.311.33:546.28:538.569.4 4098
Electron Spin Resonance Experiments on Donors in Silicon: Part 1—Electronic Structure of Donors by the Electron Nuclear Double Resonance Technique—G. Feher. (*Phys. Rev.*, vol. 114, pp. 1219–1244; June 1, 1959.) The ground-state wave function of the Sb, P and As impurities has been investigated. Details are given of the technique, and results are compared with theory.

537.311.33:546.28:538.569.4 4099
Electron Spin Resonance Experiments on Donors in Silicon: Part 2—Electron Spin Relaxation Effects—G. Feher and E. A. Gere. (*Phys. Rev.*, vol. 114, pp. 1245–1256; June 1, 1959.) The different relaxation processes connecting the four energy levels in Si doped with P have been investigated experimentally and considered in relation to theoretical work. Part 1: 4098 above.

537.311.33:546.28:539.12.04 4100
Magnetic and Electrical Properties of Reactor-Irradiated Silicon—E. Sonder. (*J. Appl. Phys.*, vol. 30, pp. 1186–1194; August, 1959.) Magnetic susceptibility measurements above 3°K and Hall effect and resistivity determinations between 50 and 300°K are reported for *n*-type Si samples irradiated with fission neutrons.

537.311.33:546.289 4101
Lattice Mobility of Thermal and Hot Carriers in Germanium—E. M. Conwell. (*Sylvania Tech.*, vol. 12, pp. 30–36; April, 1959.) A review of current theory relating to the influence of temperature and field on carrier mobility.

537.311.33:546.289 4102
Properties of some Germanium Single Crystals Grown from Solutions of Molten Metals—H. F. John. (*J. Electrochem. Soc.*, vol. 105, pp. 741–743; December, 1958.) A note giving the resistivities and mobilities of highly doped Ge crystals grown from In, Ga, Al and Sn-Sb solutions.

537.311.33:546.289 4103
Method of Etching Germanium to Precise Limits—B. T. Scofield. (*J. Sci. Instr.*, vol. 36, pp. 371–372; August, 1959.) Thickness reduction during etching is measured using the principle of the parallel plate viscometer.

537.311.33:546.289 4104
Diffusion and Solubility of Tantalum in Germanium—A. V. Sandulova and Khe-Yuĭ-Lyan. (*Dokl. Akad. Nauk SSSR*, vol. 128, pp. 329–332; September 11, 1959.)

537.311.33:546.289 4105
Use of Bismuth as a Donor-Type Impurity in Germanium Single Crystals—G. Mortimer. (*J. Electrochem. Soc.*, vol. 105, pp. 739–741; December, 1958.) Results indicate that there are no major difficulties in the use of Bi as a doping impurity for Ge.

537.311.33:546.289:535.215 4106
Impurity Photoconductivity in Germanium—H. Levinstein. (PROC. IRE, vol. 47, pp. 1478–1481; September, 1959.) The properties of Ge detectors with gold or zinc impurities are discussed, with details of the spectral response.

537.311.33:546.289:539.12.04 4107
Radiation Effects on Recombination in Germanium—O. L. Curtis, Jr. (*J. Appl. Phys.*, vol. 30, pp. 1174–1180; August, 1959.) Compilation of studies made of minority-carrier lifetime in bulk *n*- and *p*-type Ge following irradiation by gamma rays, fission neutrons, and monoenergetic 14-MeV neutrons.

537.311.33:546.289:539.12.04 4108
Electron-Bombardment-Induced Recombination Centres in Germanium—J. J. Loferski and P. Rappaport. (*J. Appl. Phys.*, vol. 30, pp. 1181–1183; August, 1959.) Experimental study of the rate of change of minority-carrier lifetime in Ge crystals bombarded by 1-MeV electrons.

537.311.33:546.289:539.12.04 4109
Energy Levels in Irradiated Germanium—E. I. Blount. (*J. Appl. Phys.*, vol. 30, pp. 1218–1221; August, 1959.) Tentative reconciliation of the differences found for the energy levels in Ge irradiated by different particles by consideration of clustering and association of defects.

537.311.33:546.289:539.12.04 4110
Energy, Orientation, and Temperature Dependence of Defect Formation in Electron Irradiation of *n*-Type Germanium—W. L. Brown and W. M. Augustyniak. (*J. Appl. Phys.*, vol. 30, pp. 1300–1309; August, 1959.) The threshold energy for the formation of stable acceptor centers, determined from conductivity measurements, is 355±5 kev, corresponding to 15.1 ev for the lightest mass Ge isotope at rest.

537.311.33:546.47–31 4111
The Diffusion and Precipitation of Indium in Zinc Oxide—D. G. Thomas. (*J. Phys. Chem. Solids*, vol. 9, pp. 31–42; January, 1959.) The indium was dissolved by coating the ZnO single crystals with a solution of indium nitrate and firing at temperatures between 800 and 1300°C.

537.311.33:546.49.241 4112
Some Semiconducting Properties of HgTe—J. Black, S. M. Ku and H. T. Minden. (*J. Electrochem. Soc.*, vol. 105, pp. 723–728; December, 1958.) The Hall coefficient and the resistivity have been measured as a function of magnetic field strength in the temperature range 80–650°K.

537.311.33:546.681.19 4113
Determination of the Effective Electron Mass in GaAs by the Infrared Faraday Effect—T. S. Moss and A. K. Walton. (*Proc. Phys. Soc.* (*London*), vol. 74, pp. 131–133; July 1, 1959.)

537.311.33:546.682.18 4114
Preparation and some Characteristics of Single-Crystal Indium Phosphide—T. C. Harman, J. I. Genco, W. P. Allred and H. L. Goering. (*J. Electrochem. Soc.*, vol. 105, pp. 731–735; December, 1958.) Two methods are described for the preparation of large single-crystals of InP, one in which the reacting of elements, purifying of compounds and growing of crystals is performed in one operation and the other in which crystals are pulled by a system similar to that used for InAs and GaAs. An electron mobility of 4000 cm²/v second at 300°K for a specimen with a carrier concentration of 1×10^{17}/cm³ was achieved.

537.311.33:546.682.18 4115
Electron Mobility in InP—M. Glicksman and K. Weiser. (*J. Electrochem. Soc.*, vol. 105, pp. 728–731; December, 1958.) A lattice mobility of 5000 cm²/v second at 290°K, varying with temperature at least as rapidly as T^{-2} in the temperature range 200°–300°K has been calculated from results.

537.311.33:546.682.18 4116
Low-Field Electrical Breakdown in *n*-Indium Phosphide—M. C. Steele. (*J. Phys. Chem. Solids*, vol. 9, pp. 93–94; January, 1959.)

537.311.33:546.682.86 4117
Transport of Electrons in Intrinsic InSb—H. Ehrenreich. (*J. Phys. Chem. Solids*, vol. 9, pp. 129–148; February, 1959.) "The mobility, thermoelectric power, Hall coefficient, and far infrared reflectivity of InSb are calculated, taking account of the nonparabolic conduction band as well as the correct wave functions of the electrons."

537.311.33:546.682.86 4118
Properties of *p*-Type Indium Antimonide: Part 2—Photoelectric Properties and Carrier Lifetime—C. Hilsum. (*Proc. Phys. Soc.* (*London*), vol. 74, pp. 81–86; July 1, 1959.) "Photoconductive and photoelectromagnetic effects were studied on *p*-type specimens of InSb with impurity concentrations varying from 10^{15} to 2×10^{17} cm^{-3}. Results were consistent with a simple theory, which assumes that the carrier relaxation time is independent of energy. Values for carrier lifetime were deduced. The fall in lifetime with increasing impurity concentration is explained as the result of two recombination processes, radiative recombination and recombination via traps." Part 1: 2466 of 1958 (Hilsum and Barrie).

537.311.33:546.682.86:538.63 4119
Transverse Magnetoresistance and Hall Effect in *n*-Type InSb—R. T. Bate, R. K. Willardson and A. C. Beer. (*J. Phys. Chem. Solids*, vol. 9, pp. 119–128; February, 1959.) Measurements confirm the existence of quantum effects for nondegenerate material in strong fields, and give strong-field magnetoresistance and Hall data in sem -quantitative agreement with the quantum treatment.

537.311.33:546.873.241 4120
Some Semiconductive Properties of Dilute Binary Solid Solutions of Bismuth in Tellurium and Tellurium in Bismuth—T. R. Piwkowski. (*Nature*, vol. 184, suppl. no. 6, pp. 355–356; August 1, 1959.)

537.311.33:678.7 4121
On the Possibility of Producing Polymeric Materials with Semiconductor Properties from Polyacrylnitryl—A. V. Topchiev, M. A. Geĭderikh, B. E. Davydov, V. A. Kargin, B. A. Krentsel', I. M. Kunstanovich and L. S. Polak. (*Dokl. Akad. Nauk SSSR*, vol. 128, pp. 312–315; September 11, 1959.) Description of semiconductors obtained by polymerization of acrylnitryl with a metallic organic catalyst LiC_4H_9 at low temperature, or by irradiation of the sample with gamma rays of 1.1×10^{22} ev. These semiconductors can have a forbidden band $\Delta E=1.7$ ev and an operating temperature up to 300°C. A graph shows the dependence of the conductivity on temperature.

537.312.62:534.22–8 4122
Change in Velocity of Sound between Normal and Superconducting States in Tin—D. F. Gibbons and C. A. Renton. (*Phys. Rev.*, vol. 114, pp. 1257–1261; June 1, 1959.) The measured change at a frequency of 80 kc is that expected from thermodynamic arguments.

538.22:538.569.4 4123
Spin-Level Inversion and Spin-Temperature Mixing in Ruby—R. H. Hoskins. (*Phys. Rev. Lett.*, vol. 3, pp. 174–175; August 15, 1959.) Describes measurements on a cavity completely filled with solid, 75 per cent sapphire, 25 per cent ruby, at 1.4°K. Successive inversion of spin levels of Cr^{+3} has been achieved by injection of a pulse of microwave power followed by a rapid variation of magnetic field.

538.221 4124
Investigation of Deformation Processes in Armco Iron by means of Internal Friction at Megacycle Frequencies—W. J. Bratina. (*Canad. J. Phys.*, vol. 37, pp. 579–590; May, 1959.)

538.221:534.2–8 4125
Dependence of Sound Velocity and Attenu-

ation on Magnetization Direction in Nickel at High Fields—G. A. Alers, J. R. Neighbours and H. Sato. (*J. Phys. Chem. Solids*, vol. 9, pp. 21–27; January, 1959.) Changes in velocity of transverse waves of the order of 0.1 per cent were observed.

538.221:538.569.4 4126
Observation of Nuclear Resonance in a Ferromagnet—A. C. Gossard and A. M. Portis. (*Phys. Rev. Lett.*, vol. 3, pp. 164–166; August 15, 1959.) The resonance occurred at 213.1 mc in a sample of finely divided-face centered cubic cobalt (Co^{59}) metal.

538.221:538.569.4 4127
Spin Fluctuation Scattering of Neutrons in Magnetite—T. Riste, K. Blinowski and J. Janik. (*J. Phys. Chem. Solids*, vol. 9, pp. 153–164; February, 1959.) Observations are found to agree with spin-wave theory over a wide range of temperatures.

538.221:538.652 4128
Investigation of the Thermal Expansion Coefficient β of Ferrites with regard to Magnetostrictive Effects—A. Zimmermann. (*Nachr-Tech.*, vol. 8, pp. 342–346; August, 1958.) The expansion coefficient as measured with the special equipment described is plotted as a function of temperature for Ni-Zn ferrites of various composition.

538.221:538.66 4129
Magnetothermal Effects in Iron and Silicon Iron—L. F. Bates and D. J. Sansom. (*Proc. Phys. Soc.* (*London*), vol. 74, pp. 53–64: July 1, 1958.)

538.221:621.3.017.31 4130
Eddy-Current Losses in Solid and Laminated Iron—P. D. Agarwal. (*Commun. and Electronics*, pp. 169–179; May, 1959. Discussion, pp. 179–181.) Accurate formulas are developed for the calculation of eddy-current losses and a single formula is given for the power factor reflected in the magnetizing winding by the magnetic circuit.

538.221:621.317.62 4131
Test Equipment for Ferrites with Square Hysteresis Loop—Drechsel. (*See* 4155.)

538.221:621.317.733 4132
A Double-T Bridge for the Measurement of the Dielectric Constant of Ferrites—G. Koepke. (*NachrTech.*, vol. 8, pp. 347–351; August, 1958.)

538.221:[621.318.124+621.318.134 4133
Microstructure and Properties of Ferrites—S. L. Blum. (*J. Amer. Ceram. Soc.*, vol. 41, pp. 489–493; November, 1958.) Domain behavior and its relation to microstructure is reviewed and the critical grain size necessary to maintain a single domain in Ni ferrite is calculated. Experimental data are given for $BaO \cdot 6Fe_2O_3$ and $NiFe_2O_4$ to show the relations between grain size, coercive force, permeability, ferromagnetic-resonance line width and dielectric losses.

538.221:[621.318.124+621.318.134 4134
Ferrites—High-Grade Materials in Communication Engineering—H. Reinboth. (*Nachr-Tech.*, vol. 8, pp. 338–341; August, 1958.) Properties and applications of various types of ferrite are reviewed.

538.221:621.318.124 4135
Analysis of the Semiconducting Properties of Cobalt Ferrite—G. H. Jonker. (*J. Phys. Chem. Solids*, vol. 9, pp. 165–175; February, 1959.) "From measurements of resistivity, activation energy, and Seebeck effect, an energy-level scheme is derived by which the semiconducting properties of $CoFe_2O_4$ can be described. These properties differ considerably from those of normal semiconductors, as the charge carriers are not free to move through the crystal lattice but jump from ion to ion."

538.221:621.318.134 4136
Disaccommodation of the Permeability of Nickel-Zinc Ferrites—A. Marais and T. Merceron. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2976–2978; May 25, 1959.) Disaccommodation is measured by the relative variation of permeability $\Delta\mu'/\mu'$ during a period of 30 min. Measurements made on Co-substituted Ni-Zn ferrites show two simultaneous relaxation effects of different origin.

538.221:621.318.134 4137
Phenomenological Theory of Magnetic Reversal of Storage-Type Ferrites—C. Heck. (*Elektrotech. Z., Edn A*, vol. 80, pp. 161–168; March 11, 1959.) The results of pulse tests on ring cores are analyzed and the effects of modifying the ferrite structure by heat treatment and the addition of oxides is investigated. See also 3397 of October (Heck and Reiner).

538.221:621.318.134:538.569.4 4138
Mechanism for Anomalous Magnetocrystalline Anisotropy Peaks in Ferromagnetic Crystals—C. Kittel. (*Phys. Rev. Lett.*, vol. 3, pp. 169–170; August 15, 1959.) Strong sharp resonance peaks in Y-Fe garnet at liquid-helium temperatures have been reported by Dillon (907 of March). A mechanism is suggested which involves an interaction between the uniaxial crystal field and an effective exchange field arising from magnetization of the ferric lattice along a different axis.

538.221:621.318.134:538.569.4 4139
Fine Structure in the Decline of the Ferromagnetic Resonance Absorption with Increasing Power Level—E. Schlömann and J. J. Green. (*Phys. Rev. Lett.*, vol. 3, pp. 129–131; August 1, 1959.) Describes measurements of the resonance absorption at a frequency of 9250 mc in single crystals of Y and Gd garnet.

538.221:621.318.134:621.372.029.6 4140
Ferrites and their Applications at Microwave Frequencies—J. Deutsch. (*Nachrichtentech. Z.*, vol. 11, pp. 473–481 and 503–507; September and October, 1958.) Gyromagnetic resonance in ferrite, and its application in reciprocal and nonreciprocal circuit elements are surveyed. 100 references.

539.23:546.621–31 4141
The Impedance, Rectification, and Electroluminescence of Anodic Oxide Films on Aluminium—A. W. Smith. (*Canad. J. Phys.*, vol. 37, pp. 591–606; May, 1959.) The oxide itself behaves as a good dielectric, but the presence of defects has an appreciable influence on the film properties.

621.315.612 4142
Ceramic Electrical Insulating Materials—M. D. Rigterink. (*J. Amer. Ceram. Soc.*, vol. 41, pp. 501–506; November, 1958.) Description of the structure and properties of porcelains, steatites and alumina ceramics, with a note on predicted advances.

621.357.6:621.3.049 4143
Electroforming of Intricate Electronic Components—E. B. Murphy. (*Electronics*, vol. 32, pp. 114–117; September 11, 1959.) Conducting parts of a complex shape can be made by electroplating on a suitable mandrel or mould which is afterwards removed.

621.791.34 4144
Solders for Nuclear and Space Environments—A. B. Kaufman. (*Electronics*, vol. 32, pp. 50–51; September 4, 1959.) Summary of properties of soft and silver solders suitable for extremes of temperature and radiation environments.

MATHEMATICS

517.7 4145
On Toroidal Functions—S. C. Loh. (*Canad. J. Phys.*, vol. 37, pp. 619–635; May, 1959.) The general theory of toroidal functions is discussed and a set of numerical tables for zonal and tesseral toroidal functions is included.

517.9:534.1 4146
On Simple Subharmonics—C. S. Hsu. (*Quart. Appl. Math.*, vol. 17, pp. 102–105; April, 1959.) Extension of a study by Rosenberg (3560 of 1958).

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74) 4147
An Error in the Determination of ΔT from the Lunar Ephemeris and the Frequency of Caesium in terms of U.T.+ΔT—C. A. Murray. (*Nature*, vol. 184, suppl. no. 7, pp. 441–442; August 8, 1959.) Recent observations of artificial satellites (4048 above) indicate that the true value of the earth's ellipticity e may be 1/298. The correction to be applied to an observed frequency of caesium resonance [3923 of 1958 (Markowitz *et al.*)] is $+8.8\ (e^{1}-294)\cos\Omega$ cps, where Ω is the longitude on the moon's node.

621.3.018.41(083.74) 4148
Proposed Feasibility Study of Frequency Shift in Sealed Atomic-Beam Frequency Standards—F. H. Reder. (Proc. IRE, vol. 47, pp. 1656–1657; Sepetmber, 1959.) Explanations are suggested for the steady frequency drift experienced after about 4000 h operation.

621.3.018.41(083.74):538.569.4 4149
Considerations on the Design of a Molecular Frequency Standards based on the Molecular-Beam Electric Resonance Method—V. W. Hughes. (*Rev. Sci. Instr.*, vol. 30, pp. 689–693; August, 1959.) Some general considerations are given followed by an account of the 100-kmc transition between the $J=0$ and $J=1$ rotational states of Li^6F^{19}. The transition is insensitive to external fields and requires only a low microwave power but the fraction of molecules in a single rotational state is small.

621.3.018.41(083.74):538.569.4 4150
A Microwave Frequency Standard employing Optically Pumped Sodium Vapor—W. E. Bell, A. Bloom and R. Williams. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 95–98; January, 1959. Abstract, Proc. IRE, vol. 47, p. 498; March, 1959.)

621.317.3:535.215:537.311.33 4151
Test Equipment for Photosensitive Semiconductors—Gereth and Müser. (*See* 4078.)

621.317.331.028.3:621.387 4152
The Use of Ionization Chambers as Sources of Current—D. Blanc, E. Fort, R. Lacoste and J. Lagasse. (*Compt. rend. Acad. Sci., Paris*, vol. 248, pp. 2984–2986; May 25, 1959.) An ionization chamber with parallel electrodes filled with air or argon at atmospheric pressure may be used as a current source for measuring resistances as high as $10^{13}\ \Omega$.

621.317.335.3 4153
The Weissfloch Transformation Theorem and its Application to the Determination of Dielectric Constants—G. Gobiet. (*Arch. elekt. Übertragung*, vol. 12, pp. 394–401; September, 1958.) The relation between Weissfloch's transformation parameter and the dielectric constant of a disk specimen treated as a quadripole in a parallel-wire transmission line or waveguide is discussed and a method of measurement is devised.

621.317.34:621.372.413 4154
Experimental Procedure foı the Determination of Cavity Parameters—A. Yariv and F. D. Clapp. (*Rev. Sci. Instr.*, vol. 30, pp. 684–687; August, 1959.) Measurements are made of the frequencies at which the voltage swr at the input to the cavity reaches a predetermined value. The knowledge of these frequencies coupled with a determination of the state of coupling to the cavity yields the necessary information. An exposition of some pertinent theoretical points is followed by a detailed description of the experimental procedure.

621.317.62:538.221 4155
Test Equipment for Ferrites with Square Hysteresis Loop—D. Drechsel. (*NachrTech.*, vol. 8, pp. 352–360; August, 1958.) The equipment described is used for testing quantities of ferrite rings under operating conditions. Automatic switching and ring-transport arrangements are provided.

621.317.7:621.314.7 4156
Transistor Simulator—J. Lüscher and P. Döme. (*Rev. Sci. Instr.*, vol. 30, pp. 656–659; August, 1959.) To measure the parameters of the transistor H equivalent circuit, a pulse is applied simultaneously to the transistor and a simulating circuit. Adjustment of the controls of the simulator gives the desired values.

621.317.727.089.6 4157
A Method of Calibration of Precision Voltage Dividers—C. B. Pinckney. (*Commun. and Electronics*, pp. 182–185; May, 1959.) The system is based on the extreme accuracy with which equal or nearly equal impedances may be compared. At frequencies up to 1 kc accuracies within 0.01 per cent were obtained for voltage ratios from 0.01 to 0.09 in one-hundredth steps and 0.1 to 0.9 in one-tenth steps.

621.317.733:538.221 4158
A Double-T Bridge for the Measurement of the Dielectric Constant of Ferrites—G. Koepke. (*NachrTech.*, vol. 8, pp. 347–351; August, 1958.)

621.317.742 4159
A Video Transmission Test Set for Steady State and Time Response Measurements—A. J. Seyler, and C. R. Wilhelm. (Proc. IRE, *Australia*, vol. 20, pp. 59–78; February, 1959.) The equipment described operates in the frequency range 100 kc–20 mc, measuring phase and time delays to an accuracy within 1 mμs and amplitudes to an accuracy within ±0.1 db.

621.317.755 4160
High Accuracy X-Y Pulse Measuring System—G. A. Haas and F. H. Harris. (*Rev. Sci. Instr.*, vol. 30, pp. 623–625; August, 1959.) This system gives a rapid voltage/current plot on a cro with an accuracy within a few tenths of 1 per cent, without the disadvantage of a constantly changing voltage.

621.317.761:621.372.413 4161
Frequency Meter Operating in the H_{011} Mode in the Range 3975–4275 Mc/s—R. L. O. Tattersall. (*P.O. Elect. Engr. J.*, vol. 52, pp. 129–135; July, 1959.) Describes a prototype for a series of three frequency meters. An accuracy within 1 part in 10^5 can be obtained and the response to unwanted modes is 15 db below the main response.

621.317.789.029.6 4162
Resistive-Film Calorimeters for Microwave Power Measurement—J. A. Lane. (IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, p. 177; January, 1959.) A summary of the essential features of a 3-cm-band calorimeter in the form of a differential air thermometer [3706 of 1955 (Gordon-Smith)] and of a transverse-film radiometer (1829 of 1956) which may be used at frequencies as low as 3 kmc. A preliminary note is given of experiments to extend the range of power measured by these instruments using a multirange dc amplifier.

621.317.789.029.6(083.74) 4163
Recent International Comparisons of Microwave Power Standards—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 43, p. 155; August, 1959.) Japanese and United States power standards showed a discrepancy greater than the estimated limits of error for the individual measurements, but results of a more recent comparison between United States and United Kingdom power standards agreed within 1.3 per cent.

621.317.794 4164
Phenomenological Description of the Response and Detecting Ability of Radiation Detectors—R. C. Jones. (Proc. IRE, vol. 47, pp. 1495–1502; September, 1959.) A discussion of the properties and the efficiency of radiation detectors assuming that they can be regarded simply as devices with input and output terminals.

621.317.794:535.15:621.396.965 4165
The Technique of Spatial Filtering—G. F. Aroyan. (Proc. IRE, vol. 47, pp. 1561–1568; September, 1959.) The detectability of objects in an optical or electro-optical system is improved by using a field stop or reticle at the focal plane of the imaging system. The space filtering properties associated with picket-fence, checker-board and star reticles are considered.

621.317.794:621.396.822 4166
Noise in Radiation Detectors—R. C. Jones. (Proc. IRE, vol. 47, pp. 1481–1486; September, 1959.) A general discussion of the eight kinds of noise which are found in the output of radiation detectors.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.247.4:621.397.9 4167
Electronic Production of Lines of Equal Density: Part 2—W. Krug and J. Schusta. (*Optik, Stuttgart*, vol. 15, pp. 550–559; September, 1958.) The photoelectric apparatus with mechanical scanning described in Part 1 (*ibid.*, vol. 15, pp. 145–152; February/March, 1958) has been improved. Details are also given of a television-optical arrangement for producing lines of equal density with application to a television microscope or telescope.

537.533.9:67.002.2 4168
Processing Materials with Electron Bombardment—A. Lawley. (*Electronics*, vol. 32, pp. 39–41; September 4, 1959.) Electron-beam melting and welding lead to greater purity, high-temperature stability and strength.

616–073.7:621.374.3 4169
Electronic Equipment for Pulse Generation Controlled by Heart Phases—E. Blüthgen and J. Bayer. (*NachrTech.*, vol. 8, pp. 423–426; September, 1958.) The instrument described incorporates thyratron multivibrator circuits with facilities for variable pulse delay and can be controlled by signals obtained from electrocardiograph electrodes or a heart microphone.

621.384.8:621.318.381:621.316.7.078.3 4170
Atomic-Beam Magnet-Stabilization System—G. O. Brink and N. Braslau. (*Rev. Sci. Instr.*, vol. 30, pp. 670–674; August, 1959.) The magnetic field is automatically controlled by means of a resonance in the beam. See 3072 of September (Bailey and Fellows).

621.385.833 4171
Correction of Aperture Error in Electrostatic Lenses by Space Charge—G. Haufe. (*Optik, Stuttgart*, vol. 15, pp. 521–537; September, 1958.) The spherical aberration of es lenses can be compensated by a space charge applied in a special lens. The electron-optical characteristics of such a lens when used in conjunction with ordinary electron lenses are investigated.

621.398 4172
A Comparison of Telemetry Systems—A. Cowie. (*J. Brit. IRE*, vol. 19, pp. 491–492; August, 1959.) Factors influencing the choice of a modulation system are reviewed and multichannel transmission is discussed with reference to time and frequency-multiplex systems.

621.398 4173
A Six-Channel High-Frequency Telemetry System—T. C. R. S. Fowler. (*J. Brit. IRE*, vol. 19, pp. 493–507; August, 1959.) Circuit details and operational results are given for a frequency-multiplex FM-AM system in the 465-mc band. Future uses of the system and possible methods for increasing its range are also discussed.

PROPAGATION OF WAVES

621.396.11 4174
System Loss in Radio Wave Propagation—K. A. Norton. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 53–73; July/August, 1959.) A summary of the applications of the concept of "system loss" in radio wave propagation. Examples are given which include calculations of losses for tropospheric forward scatter and ground-wave propagation, together with an analysis of the variation of path antenna gain with time in ionospheric scatter propagation. For a note on a C.C.I.R. recommendation relating to transmission loss, see Proc. IRE, vol. 47, pp. 1661–1662; September, 1959.

621.396.11:551.510.52 4175
Climatology of Ground-Based Radio Ducts—Bean. (*See* 4053.)

621.396.11:551.510.52 4176
Measurements of Phase Stability over a Low-Level Tropospheric Path—M. C. Thompson, Jr., and H. B. Janes. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 45–51; July/August, 1959.) The results of a 40-h period of measurements at 9400 mc over a 9.4-mile path are analyzed. The power spectral density of the phase variations is proportional to $f^{-2.8}$, and the long-term variations are correlated with refractivity measurements at the path terminals.

621.396.11:551.510.52 4177
Synoptic Variation of the Radio Refractive Index—B. R. Bean and L. P. Riggs. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 91–97; July/August, 1959.) An analysis of the variations during an outbreak of polar continental air, shows that the value of the refractivity, reduced to sea level, changes systematically with the passage of the cold front.

621.396.11:551.510.535 4178
Mode Expansion in the Low-Frequency Range for Propagation through a Curved Stratified Atmosphere—H. Bremmer. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 75–85; July/August, 1959.) An expansion of particular value in the study of gradual transitions in electron density at the lower edge of the ionosphere.

621.396.11:551.510.535 4179
Propagation of Long Electrical Waves in Magnetized Plasmas and their Passage through Plasma Layers—W. O. Schumann. (*Z. angew. Phys.*, vol. 10, pp. 428–433; September, 1958.) An analysis taking account of damping effects. See 1996 of June.

621.396.11:551.510.535 4180
A Theory of Ionospheric Radio Wave

Scattering under the Influences of Ion Production and Recombination—Maeda, Kato and Tsuda. (*See* 4059.)

621.396.11:551.510.535 4181
Ionospheric Radio Wave Scattering in the Electrodynamically Controlled Turbulence—Maeda, Kato and Tsuda. (*See* 4060.)

621.396.11.029.51 4182
Low-Frequency Propagation Paths in Arctic Areas—A. D. Watt, E. L., Maxwell and E. H. Whelan. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 99–112; July/August, 1959.) Field strengths from transmitters in the Labrador and Greenland area were measured at the earth's surface and during aircraft flights, and compared with theoretical values.

621.396.11.029.6:621.396.812 4183
Aperture-to-Medium Coupling on Line-of-Sight Paths: Fresnel Scattering—E. Levin, R. B. Muchmore and A. D. Wheelon. (IRE Trans. on Antennas and Propagation, vol. AP-7, pp. 142–146; April, 1959. Abstract, Proc. IRE, vol. 47, p. 1282; July, 1959.)

621.396.11.029.62/.63 4184
Microwave Propagation over Rough Surfaces—M. P. Bachynski. (*RCA Rev.*, vol. 20, pp. 308–335; June, 1959.) A general survey; the results of available experimental measurements made over all types of terrain are summarized and tabulated. Over 100 references.

621.396.11.029.63/.64 4185
Studies in Tropospheric Propagation Beyond the Horizon—A. B. Crawford, D. C. Hogg and W. H. Kummer. (*Bell Sys. Tech J.*, vol. 38, pp. 1067–1178; September, 1959.) Detailed description of experiments over a 171-mile land path on 460 and 4110 mc.

621.396.11.029.64 4186
A Contribution on the Origin of Multipath Propagation of Microwaves—G. Megla. (*NachrTech.*, vol. 8, pp. 389–391; September, 1958.) Cloud layers give rise to multiple reflections and the effect of these on the rece ved field strength at 10 cmλ is calculated.

621.396.11.029.65:551.578.1 4187
On the Measurement of Attenuation by Rain at 8.6 mm Wavelength—S. Okamura, K. Funakawa, H. Uda, J. Katō and T. Oguchi. (*J. Radio Res. Labs, Japan*, vol. 6, pp. 255–267; April, 1959.) Reflections were obtained from a target 400 m distant using FM radar equipment. The attenuation was measured as a function of precipitation rates of up to 140 mm/h. The results are in fair agreement with Ryde's theory but there is a difference between those obtained with horizontal and vertical polarization.

RECEPTION

621.396.621:621.376.3:621.314.7 4188
A Domestic F.M. Receiver using Diffused-Base Mesa Transistors—M. Applebaum and E. Midgley. (*Electronic Engrg*, vol. 31, pp. 448–453; August, 1959.) The characteristics of diffused-base mesa transistors are discussed and their performance in the rf and IF stages of a receiver is given.

621.396.621:621.376.33 4189
Determination of the Expression for a Signal at the Output of a Limiter—I. Medici. (*Note Recensioni Notiz.*, vol. 7, pp. 510–521; September/October, 1958.) The general expressions derived can be used to calculate the effects of noise, fading and amplitude distortion of the FM carrier.

621.396.621.54:621.376.33 4190
The Distortion Factor due to Incomplete Limiting in V.H.F. F.M. Receivers—E. G. Woschni. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 18–25; July, 1958.) Extension of previous investigations (2873 of 1958) to take account of signal distortion due to incomplete limiting.

621.396.8:621.396.666 4191
The Dependence of Combiner Diversity Gain on Signal-Level Distribution—B. Easter, C. H. Maddock and R. G. Medhurst. (Proc. IRE, vol. 47, pp.1651–1652; September, 1959.) Discrepancies in the predicted improvement of combiner over selector diversity and variations in the observed value may be due to the form of signal-level distribution involved.

621.396.812.3:621.3.087.4 4192
Analysing Multipath Delay in Communications Studies—J. F. Lyons, Jr. (*Electronics*, vol. 32, pp. 52–55; September 4, 1959.) Describes an automatic analyzer which triggers at changes in signal amplitude due to multipath delays of pulse transmissions and prints the delay times between triggerings.

621.396.828 4193
Wide-Band Radio-Interference Suppression Equipment for Life Installations and Screened Rooms—M. Ortloff. (*Elektrotech. Z., Edn. B*, vol. 11, pp. 136–139; April 21, 1959.) Various types of interference suppressors conforming to the German VDE specifications are described. They include units giving 60 db suppression in the range 150 kc–30 mc, and 80–100 db in the range 30–1000 mc, and adaptors to extend the range to lower and higher frequencies, *e.g.* for mains interference suppression in screened test enclosures.

STATIONS AND COMMUNICATION SYSTEMS

621.376.54:621.376.3 4194
The Influence of Frequency Modulation of the High-Frequency Carrier on Pulse Width Modulation—R. Ebermann. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 48–59; September, 1958.) The displacement of the leading and trailing edges of high-frequency pulses whose carrier is frequency-modulated is investigated, and the effect on crosstalk attenuation of simultaneous PWM is considered.

621.391 4195
The Coding Law of Information Theory—F. H. Lange. (*Hochfrequenz. u. Eletkroak.*, vol. 67, pp. 1–4; July, 1958.) The calculation and realization of optimum coding is considered in communication systems with constant flow of information with or without a store between source and coding device.

621.391:621.376 4196
The Significance of 90° Phase Shift in Special Modulation Methods—H. Schlesier. (*Hochfrequenz. u. Elektroak.*, vol. 67, pp. 42–48; September, 1958.) S.s.b. modulation by means of phase-shifting, and quadrature modulation are considered. Wide-band 90° phase-shift circuits are discussed.

621.396.2:621.384.3 4197
Infrared Communications Receiver for Space Vehicles—W. E. Osborne. (*Electronics*, vol. 32, pp. 38–39; September 18, 1959.) The principles of an experimental infrared transistor receiver are discussed with reference to its full circuit diagram. The receiver discriminates against solar radiation.

621.396.3:621.396.43:523.5 4198
Metre-wave Propagation by Meteoric Ionization—J. Grosskopf. (*Nachrichtentech. Z.*, vol. 11, pp. 455–460; September, 1958.) The principles of operation of the JANET system are discussed with reference to the papers published in Proc. IRE, vol. 45, December, 1957 [see *e.g.* 908 of 1958 (Forsyth *et al.*)].

621.396.41:621.396.8 4199
C.C.I.T.T. Recommendations for Multichannel Radio Relays and White Noise—C. A. Parry. (*Commun. and Electronics*, pp. 107–117; May, 1959.) The channel signal/noise ratio is considered in detail. Nonlinear noise is shown to be dependent on both the number of channels and the signal power. Thermal noise in the output is influenced by the amplification of the RF signal.

621.396.5:621.395.34 4200
A New Method of Signalling for Single-Channel and Mobile Radiotelephone Systems—B. W. G. Penhall and J. D. Thomson. (Proc. IRE, *Australia*, vol. 20, p. 279; May, 1959.) A preliminary note on a carrier-frequency-shift system controlled by the dc impulses from a telephone handset.

621.396.61:621.3.018.41(083.74) 4201
Power Requirements and Choice of an Optimum Frequency for a Worldwide Standard-Frequency Broadcasting Station—A. D. Watt and R. W. Plush. (*J. Res. Nat. Bur. Stand.*, vol. 63D, pp. 35–44; July/August, 1959.) "Calculations are presented for the expected transmission characteristics and atmospheric noise levels in the 8- to 100-kc band. When these are combined with carrier-to-noise requirements for a given precision of frequency comparison, it is indicated that a minimum radiated power in the order of 10 to 100 kw for frequencies in the vicinity of 20 kc will be required to provide worldwide coverage. Minimum observation times of 15 to 30 min appear to be required for these transmitter powers in order to obtain a precision of frequency comparison of 1 part in 10^9 for typical transmission paths. Carrier-to-noise requirements and the factors determining this ratio are considered for typical receiving systems."

621.396.65 4202
A 215-Mile 2720-Mc/s Radio Link—L. H. Doherty and G. Neal. (IRE Trans. on Antennas and Propagation, vol. AP-7, pp. 117–126; April, 1959. Abstract, Proc. IRE, vol. 47, p. 1282; July, 1959.)

621.396.65:621.396.822 4203
Correlation Function and Power Spectra of Radio Links affected by Random Dielectric Noise—D. S. Bugnolo. (IRE Trans. on Antennas and Propagation, vol. AP-7, pp. 137–141; April, 1959. Abstract, Proc. IRE, vol. 47, p. 1282; July, 1959.)

621.396.721 4204
An 80-Watt F.M. Transmitting and Receiving Installation for Outside Sound Broadcasts—H. Bohlmann an A. Rettig. (*Rundfunktech. Mitt.*, vol. 2, pp. 210–219; October, 1958.) The installation consists of two similar sets of apparatus. Radiation is vertically polarized and four switchable fixed frequencies in the 7-m band are used. A range of about 6 miles is obtainable.

SUBSIDIARY APPARATUS

621.311.6.072.2 4205
Time-Controlled Unit-Function, Constant-Voltage Generator—Y. Ettinger and H. Edels. (*J. Sci. Instr.*, vol. 36, pp. 362–364; August, 1959.) The output voltage is variable from 40 to 1000 v and is independent of current over the range 0–12 amperes.

621.311.69:535.215–15 4206
Infrared-Detector Silicon Solar-Cell Power Supply—L. W. Schmidt and J. L. Davis. (Proc. IRE, vol. 47, pp. 1519–1520; September, 1959.)

621.314.1:621.373.52:621.314.7 4207
Design of Transistor Power Converters—T. R. Pye. (*Electronics*, vol. 32, pp. 56–58;

September 4, 1959.) Description of two dc/dc converters: a Si type delivering 15 w and a Ge type delivering 100 w both from a 24 v source.

621.316.722.078:621.387 4208

The Design of Thyratron-Stabilized D.C. Supplies—B. G. Higdon and M. E. Bond. (*Electronic Engrg*, vol. 31, pp. 475–479; August, 1959.) The components used in a simple stabilizer and a feedback-loop unit for hv supplies at currents greater than 0.5 amperes are discussed.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2:621.3.018.782 4209

Waveform Distortion in Television Links: Part 1—Introduction to Waveform Distortion—I. F. Macdiarmid. (*P.O. Elect. Engrg. J.*, vol. 52, pp. 108–114; July, 1959.) Different forms of distortion may be traced using a pulse-and-bar type of test waveform.

621.397.24 4210

Some Aspects of Television Transmission over Long Distance Cable Links—H. Mumford. (*J. Brit. IRE*, vol. 19, pp. 509–519; August, 1959.) The basic properties of the systems used for long-distance transmission by coaxial cable are outlined and some of the basic transmission limits for such systems are shown to be in agreement with subjective laws which are discussed. Several methods of transmitting synchronizing data are reviewed and compared.

621.397.24:621.315.28 4211

Prospects for Transatlantic Television by Cable—R. J. Halsey and A. R. A. Rendall. (*P.O. Elect. Engrg. J.*, vol. 52, pp. 136–139; July, 1959.) By reducing television bandwidth to 0.5–1 mc it should be possible to build a coaxial system within a few years.

621.397.5:535.623 4212

Experiments on the Adaptation of the N.T.S.C. Colour Television System to the European 625-Line Standard—J. Davidse. (*Nachrichtentech. Z.*, vol. 11, pp. 461–466; September, 1958.) Tests were made on experimental 625-line equipment using chrominance subcarrier frequencies of 4.10 or 4.43 mc and variable bandwidths for the color signals. The effects of bandwidth reduction and choice of subcarrier frequency on reproduction and compatibility and on cross-talk between luminance and chrominance information are discussed.

621.397.6 4213

The Development of the German Standards Converter Technique—H. Bödeker. (*Rundfunktech. Mitt.*, vol. 2, pp. 220–223; October, 1958.) Two types of equipment for use in Eurovision program exchanges are described; they incorporate image-orthicon or vidicon tubes respectively. For English version see *E.B.U. Rev.*, Pt. A, pp. 14–17; October, 1958.

621.397.611:621.375.4 4214

A Transistor Video Amplifier with Keyed Black-Level Control, Clamping, and Adjustable Black-Level "Set-Up"—H. Anders. (*Rundfunktech. Mitt.*, vol. 2, pp. 224–233; October, 1958.) The amplifier described is intended for use in television cameras and film scanners where a reference voltage is available during line flyback. Drift transistors are used in the equipment which has an output of 1 v peak-to-peak into 75Ω, 28 db gain, and a power consumption of 0.6 w.

621.397.611.2 4215

Sensitivity and Motion Capturing Ability of Television Camera Tubes—R. G. Neuhauser. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 68, pp. 455–461; July, 1959. Discussion.)

621.397.611.2 4216

On the Detective Quantum Efficiency of Television Camera Tubes—R. C. Jones. (*J. Soc. Mot. Pict. Telev. Eng.*, vol. 68, pp. 462–466; July, 1959.) The detective quantum efficiency is represented as a function of the wavelength of the photocathode radiation, of the irradiation of the cathode and of the line-number of the target. It is calculated for one vidicon and two image orthicon camera tubes.

621.397.62 4217

The Design of Dual-Standard Television Receivers for the French and C.C.I.R. Television Systems—C. J. Hall. (*J. Brit. IRE*, vol. 19, pp. 457–468; August, 1959.) Typical receivers for the two systems differ in all parts of the circuit except the power supply, frame time-base and audio-frequency sections. Consequently, direct circuit switching is not considered practicable and attempts are made to use common circuits for both systems without degradation of performance. Where this is impossible, duplicate circuits with switched h.t. supplies are suggested. Suitable circuits and techniques to achieve a dual-standard receiver are described and a complete receiver design is outlined.

621.397.62:535.623:621.385.832 4218

Quality-Control Determinations of the Screen Persistence of Colour Picture Tubes—J. M. Forman and G. P. Kirkpatrick. (*RCA Rev.*, vol. 20, pp. 293–307; June, 1959.) Equipment is described in which a photomultiplier tube behind a narrow slit monitors a section of a television raster, producing an output which is displayed on a calibrated oscilloscope for persistence measurements.

621.397.62:537.531 4219

Assessment of X-Radiation from Television Receivers—A. Ciuciura. (*J. Brit. IRE*, vol. 19, pp. 469–482; August, 1959.) Statistical results are obtained by combining the X-radiation properties of a cr tube, in terms of e.h.t., beam current and spread, with those of e.h.t. supplies defined in terms of mean potential, spread and internal impedance.

621.397.621.2:621.314.7 4220

Peak Flyback Voltage in Transistorized TV Horizontal Deflection Circuits—W. F. Palmer. (Proc. IRE, vol. 47, pp. 1655–1656; September, 1959.) The voltage is shown to be about 10 times the supply voltage; the peak/saturation voltage ratio should be at least 100:1.

621.397.8 4221

Reduction of Co-channel Interference by Precise Frequency Control of Television Picture Carriers: Part 2—W. L. Behrend. (*RCA Rev.*, vol. 20, pp. 349–364; June, 1959.) Additional subjective viewing tests are reported. Part 1: 2618 of 1957.

621.397.8 4222

The Physical Basis of the Assessment of Picture Quality—W. Kroebel. (*Rundfunktech. Mitt.*, vol. 2, pp. 234–245; October, 1958.) The subjective assessment of picture quality can be expressed by a physical parameter. Various methods of quality assessment are discussed. See also 3138 of September (Kroebel *et al.*) and 1393 of April.

TRANSMISSION

621.396.61:621.396.674.3:627.95 4223

Ferromagnetic Transmitter Aerial for Cases of Distress at Sea—K. Baur and G. Ziehm. (*Telefunken Ztg.*, vol. 31, pp. 150–161; September, 1958. English Summary, p. 204.) Details are given of an emergency transmitter with internal ferrite-rod magnetic dipole. It is designed for operation at 2.182 mc and initial trials have shown that bearings could be taken over a range of 5 nautical miles of a floating transmitter operating at a frequency of 2.05 mc with 3.2 μw output.

621.396.61.072.7:621.396.74 4224

The Problem of Synchronizing Transmitter Frequencies—H. Ehlers and H. Thies. (*Rundfunktech. Mitt.*, vol. 2, pp. 201–209; October, 1958.) The development, present state, and limitations of common-channel operation and the requisite synchronization techniques are discussed. Details are given of the systems adopted by various regional broadcasting authorities in the German Federal Republic, and a method of comparing local frequency standards with standard-frequency transmissions is described.

TUBES AND THERMIONICS

621.314.63 4225

A New Semiconductor Component for Electronic Circuits—G. Zielasek. (*Elektrotech. Z.*, *Edn. B.*, vol. 11, pp. 227–228; May, 21, 1959. Correction, vol. 11, p. 274; June. 21, 1959.) A new type of *p-n* junction diode, called Variode, is described. It is intended for use in voltage-limiter, current gating and control circuits for relatively high currents. Unlike the Zener diode it operates in the forward direction and at lower voltages. An application of a Ge Variode as a regulating device in a motor-vehicle electrical system is described in *Elektronische Rundschau*, vol. 13, pp. 378–381; October, 1959.

621.314.63 4226

Direct Observation of Phonons during Tunnelling in Narrow Junction Diodes—N. Holonyak, Jr, I. A. Lesk, R. N. Hall, J. J. Tiemann and H. Ehrenreich. (*Phys. Rev. Lett.*, vol. 3, pp. 167–168; August 15, 1959.) Successive maxima in the forward *i/v* characteristics of Ge and Si tunnel diodes are attributed to optical-transition phonons which are involved in the tunnel process.

621.314.63:621.318.57 4227

A Silicon-Controlled Rectifier: Part 1—Characteristics and Ratings—D. K. Bisson and R. F. Dyer. (*Commun. and Electronics*, pp. 102–106; May, 1959.) Data are given for a three-terminal semiconductor device with a *p-n-p-n* junction configuration having characteristics similar to a thyratron.

621.314.7 4228

The Possibilities and Problems of Junction-Transistor Applications—G. Meyer-Brötz. (*Telefunken Ztg.*, vol. 31, pp. 162–174; September, 1958. English Summary, pp. 204–205.) A review of circuitry. 128 references.

621.314.7 4229

Junction Transistor Characteristics and Manufacturing Problems—W. Engbert. (*Telefunken Ztg.*, vol. 31, pp. 175–178; September, 1958. English summary, p. 205.) The results achieved by the diffusion method and its limitations are reviewed.

621.314.7 4230

Surface-Immune Transistor Structure—H. Nelson. (*RCA Rev.*, vol. 20, pp. 222–228; June, 1959.) A form of transistor construction is described in which the base region is brought to an external surface through the central region of the collector. Si power transistors of this form are characterized by a high degree of minority carrier conservation and immunity from surface effects.

621.314.7 4231

Stabilization of Transistor Gain over Wide Temperature Ranges—R. A. Schmeltzer. (*RCA Rev.*, vol. 20, pp. 284–292; June, 1959.)

An analysis of the effects of junction temperature on the gain of a Ge *p-n-p* alloy-junction transistor. If the ac resistance of the driving source is of a suitable low value it is possible to achieve a gain stability of ±0.125 db over the temperature range −70° to +80°C in a stage having a gain of 30 db.

621.314.7 4232
Determining Transistor High-Frequency Limits—J. Lindmayer and R. Zuleeg. (*Electronics*, vol. 32, pp. 31–33; August 21, 1959.) Technique using coaxial elements extends measurement of the maximum frequency of oscillation up to 1 kmc.

621.314.7:535.215 4233
The Mode of Operation of *n-p-n* Phototransistors—A. Hoffmann. (*Z. angew. Phys.*, vol. 10, pp. 416–418; September, 1958). Two equivalent circuits of the phototransistor are considered consisting of a normal transistor and a photodiode, with the photodiode connected in parallel, a) with base and collector, and b) with base and emitter. A genuine amplification of the primary photocurrent is only obtained in the first circuit; this is confirmed by experiment.

621.314.7:621.317.7 4234
Transistor Simulator.—Lüscher and Döme. (*See* 4156.)

621.314.7:621.396.621:621.376.3 4235
A Domestic F.M. Receiver using Diffused-Base Mesa Transistors—Applebaum and Midgley. (*See* 4188.)

621.314.7.012 4236
Transistor High-Frequency Parameter f_1—L. G. Cripps. (*Electronic Radio Eng.* vol. 36, pp. 341–346; September, 1959.) The frequency, f_1, at which the earthed-emitter short-circuit current gain has fallen to unity is discussed as a parameter for defining the hf characteristics of a transistor. A method is described for measuring f_1 to an accuracy within 2 per cent at frequencies up to 200 mc.

621.383 4237
Classification and Analysis of Image-Forming Systems—W. K. Weihe. (Proc. IRE, vol. 47, pp. 1593–1604; September, 1959.) Both photon-sensitive and temperature-sensitive devices are considered. Classification is by the type of transducer and the method of reading-in information.

621.383.2 4238
Photoemissive Image-Forming Systems—R. S. Wiseman and M. W. Klein. (Proc. IRE, vol. 47, pp. 1604–1606; September, 1959.) The properties of some evaporated, semi-transparent photoemissive surfaces and the electron-optical systems used to form a visible image are reviewed.

621.383.4:535.215–15 4239
Single-Crystal Infrared Detectors based upon Intrinsic Absorption—F. F. Ricke, L. H. DeVaux and A. J. Tuzzolino. (Proc. IRE, vol. 47, pp. 1475–1478; September, 1959.) A discussion of the use of crystals such as InAs having an energy gap of 0.4 ev or less.

621.383.4:535.215–15 4240
The Measurement and Interpretation of Photo-detector Parameters—R. F. Potter, J. M. Pernett and A. B. Naugle. (Proc IRE, vol. 47, pp. 1503–1507; September, 1959.) An outline of measuring techniques and the results obtained.

621.383.4:535.215–15 4241
Detectivity and Preamplifier Considerations for Indium Antimonide Photovoltaic Detectors—G. R. Pruett and R. L. Petritz. (Proc. IRE, vol. 47, pp. 1524–1529; September, 1959.) Equations are derived which indicate the most suitable type of preamplifier required to obtain maximum sensitivity.

621.383.4:621.397.3:535–15 4242
An Infrared Pickup Tube—G. A. Morton and S. V. Forgue. (Proc. IRE, vol. 47, pp. 1607–1609; September, 1959.) The principle of frame storage and sequential read-out is used in a tube containing a thin layer of photoconductor as the sensitive target surface.

621.385.029.6 4243
Microwave Tube Techniques in Europe—H. Huber. (*Le Vide*, vol. 13, pp. 220–246; September/October, 1958. In French and English.) A review in which cathodes, circuit construction and vacuum techniques are discussed. 76 references.

621.385.029.6 4244
The C. W. Magnetron Valve Type 7091—W. Schmidt. (*Elektronische Rundschau*, vol. 12, pp. 309–314; September, 1958.) Technical data on a 2.5-kw magnetron for operation at 2–4 kmc.

621.385.029.6:537.533 4245
The Energy Relations in Electron Beams—H. Kogelnik. (*Arch. elekt. Übertragung*, vol. 12, pp. 419–427; September, 1958.) In the second-order perturbation theory of the one-dimensional electron beam the theorem of ac power conservation becomes identical with Chu's theorem of the linear theory. The significance of the residual second-order terms in the expression of total kinetic power is discussed. See also 323 of 1958 (Paschke).

621.385.029.6:537.533 4246
The Propagation of Perturbations along Magnetically Focused Electron Beams—F. Paschke. (*RCA Rev.*, vol. 20, pp. 254–283; June, 1959.) An analysis of beam behavior for magnetic field values between Brillouin field and infinity.

621.385.029.6:537.533:621.375.9 4247
Parametric and Pseudo-parametric Amplifiers—P. A. Clavier. (Proc. IRE, vol. 47, p. 1651; September, 1959.) A proposal is made for an electron-beam parametric amplifier with transverse modulation on the beam.

621.385.029.65 4248
Development ot Magnetrons for Millimetre Wavelengths—S. Aoi, S. Nakajima, S. Mito and K. Baba. (*Le Vide*, vol. 14, pp. 7–18; January/February, 1959. In French and English.) A description of the construction, operation and characteristics of the rising-sun-type split-anode magnetron for wavelengths of 5.7 and 7.0 mm.

621.385.032.21 4249
Reversible Poisoning by Sulphur, Oxygen and other Gases of Oxide-Coated Cathodes at High Temperatures—N. A. Surplice. (*Brit. J. Appl. Phys.*, vol. 10, pp. 359–363; August, 1959.) The results support the hypothesis that high-temperature reversible poisoning is caused by the absorption of negative ions on the internal and external surfaces of the porous oxide-coating.

621.385.032.213.6 4250
A Technique for Processing Nickel Matrix (Moulded) Thermionic Cathodes—E. M. Boone, W. H. Cornetet, Jr, D. Kahng and S. Taylor. (Proc. IRE, vol. 47, pp. 1650–1651; September, 1959.)

621.385.1:621.373.423.029.6 4251
Electron Oscillations in Vacuum Tubes—F. Möhring. (*Elektronische Rundschau*, vol. 12, pp. 301–304; September, 1958.) The mechanism causing transit-time oscillations in conventional tubes is discussed and the frequency characteristics of the phenomenon are examined with reference to experimental curves.

621.385.14–713 4252
The Conduction of Heat from Heavily Loaded Anodes in Thermionic Valves.—W. Heinze, H. Göllnitz and W. Baer. (*NachrTech.*, vol. 8, pp. 398–410; September, 1958.) Detailed calculation of heat convection along cooling fins of various cross-section, and determination of cooling-surface design for optimum heat dissipation. The process of evaporation cooling is considered and load limits of transmitter tubes with different cooling surfaces are compared.

621.385.2 4253
The A.S.T.M. Planar Reference Diode for Testing Electron Tube Materials and Processes—R. P. Misra. (*Le Vide*, vol. 13, pp. 256–263; September/October, 1958. In English and French.) A description of the construction, special cleaning and processing of a planar diode designed as a control by the American Society for Testing Materials.

621.385.832.032.35 4254
Phosphors for Cathode-Ray Tubes in Industrial and Low Scanning Speed Display Systems—M. D. Dudley. (*J. Brit. IRE*, vol. 19, pp. 485–489; August, 1959.) The properties of available phosphors are discussed in relation to the type of display required. A particular combination of phosphors is described which has a persistence controllable by infrared radiation.

MISCELLANEOUS

061.4:621.37/.39 4255
The National Radio Show—in Retrospect—(*Wireless World*, vol. 65, pp. 437–447; October, 1959.) An analysis of recent developments in antennas, television and radio receivers, and sound reproducers seen at the Radio Show (*see* 3897 of November).

061.4:621.396 4256
Airshow Electronics—(*Wireless World*, vol. 65, pp. 429–431; October, 1959.) Review of aviation electronics at the S.B.A.C. exhibition.

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IRE NATIONAL CONVENTION RECORD

Volume VII, 1959

The Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N.Y.

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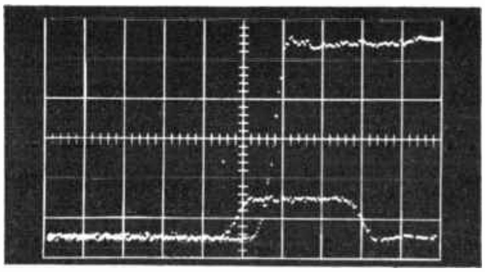
NEW INTERNALLY TRIGGERED

TEKTRONIX ®

0.6-Nanosecond Risetime

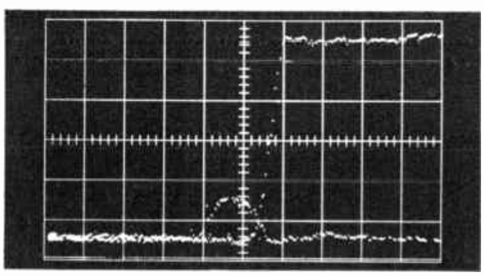
(approximately 600-MC Bandwidth)

The waveform photographs below show the ability of the Tektronix Sampling System to display a wide range of pulses. These photographs were purposely chosen to illustrate the system's abilities under marginal conditions.



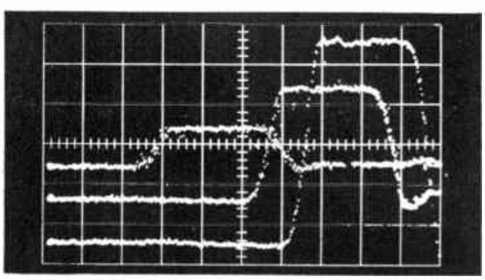
1

The alternate pulse feature of the Type 110 pulse generator is being used to generate a large, long pulse, and a short, small pulse. The trigger take-off system's sensitivity is set for maximum. The signal level is 100 mv/cm, and the sweep speed is 1 nsec/cm. There is clearly less than 1 nsec time difference in triggering on the 100 mv, 3 nsec and the 500 mv long step signals.



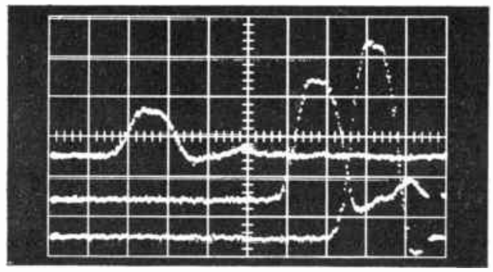
2

This picture shows the same conditions as in Fig. 1, except the small pulse is now only 1 nsec wide. The time shift relative to the large step is just over 1 nsec.



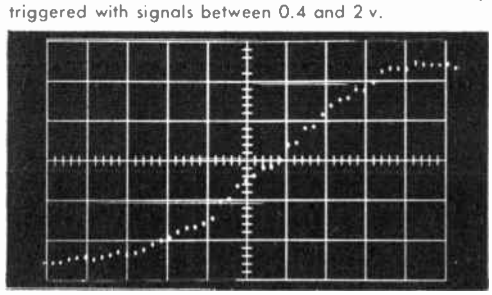
3

The system is operating at maximum sensitivity, 20 mv/cm. A triple exposure, positioned vertically to align the 50% points, allows easy measurement of the time slip. Under these extreme conditions, the smallest pulse has an energy of about 24 millipicojoules. The trigger take-off system then removes approximately 1 millipicojoule for application to the switched system of amplifiers and the trigger regenerator.



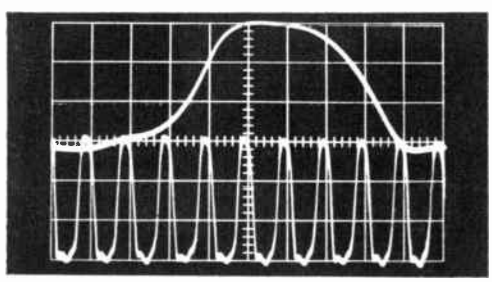
4

The amplifiers in the trigger channel (used in the previous 3 pictures) are switched out. The sensitivity is 2 v/cm. The smallest of the 1 nsec wide pulses furnishes approximately 0.4 v to the trigger regenerator, through the trigger take-off system. This picture is of interest since this is the narrow-pulse response which is obtainable with both the 110 and N Units, when externally triggered with signals between 0.4 and 2 v.



5

The leading edge of the large pulse of Figure 3 is displayed with the 1 nsec/cm sweep speed magnified ten times. This gives an equivalent sweep speed of 100 picoseconds/cm. The risetime of the complete system—110 pulse generator, 110 trigger take-off, 113 delay cable and the N unit—is well under 0.6 nsec.



6

Double exposure shows a 60-mv, 100-mc continuous pulse train at equivalent sweep times of 1 nsec/cm and 10 nsec/cm. The Type 110 derives a trigger from the signal, permitting the Tektronix Sampling System to operate without external triggers, counting down from 100-mc to the 100-kc sampling rate of the N Unit.

28 Fields of Special Interest-

The 28 Professional Groups are listed below, together with a brief definiton of each, the name of

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Annual fee: $2.

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Mr. Lewis M. Sherer, Chairman, RTCA, Washington, D.C.

31 Transactions, *5, *6, & *9, and *Vol. ANE-1, Nos. 2 and 3; Vol. 2, No. 1-3; Vol. 3, No. 2; Vol. 4, No. 1, 2, 3; Vol. 5, No. 2, 3, 4; Vol. 6, No. 1.

Antennas and Propagation

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

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

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7 Transactions, PGAC-3-4-5-6, AC-4, No. 1.

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Annual fee: $2.

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23 Transactions. *7, 8; BTR-1, No. 1-4; BTR-2, No. 1-2-3; BTR-3, No. 1-2; BTR-4, No. 2, 3-4; BTR-5, No. 1, 2.

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14 Transactions, No. 2, 4, 6, 7, 8, 9, 10, 11, 12, 13, 14.

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Annual fee: $3.

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Annual fee: $2.

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15 Transactions, CS-2, No. 1; CS-5, No. 2, 3; CS-6, No. 1, 2; CS-7, No. 1, 2, 3.

Component Parts

Annual fee: $3.

The characteristics, limitation, applications, development, performance and reliability of component parts.

Mr. J. J. Drvostep, Chairman, Sperry Gyroscope Co., Great Neck, N.Y.

17 Transactions, Vol. CP-3, No. 2; CP-4, No. 1, 2, 3-4; CP-5, No. 1, 2, 3, 4; CP-6, No. 1, 2, 3.

Education

Annual fee: $3.

To foster improved relations between the electronic and affiliated industries and schools, colleges, and universities.

Dr. R. L. McFarlan, Chairman, 20 Circuit Rd., Chestnut Hill 67, Mass.

8 Transactions. Vol. E-1, No. 3, 4; E-2, No. 1, 2, 3, 4.

Electron Devices

Annual fee: $3.

Electron devices, including particularly electron tubes and solid state devices.

Dr. W. M. Webster, Chairman, RCA Labs., Princeton, N.J.

26 Transactions, *Vol. ED-1, No. 3-4; ED-3, No. 2-4; ED-4, No. 2-3, 4; ED-5, No. 2, 3, 4; ED-6, No. 1, 3.

Electronic Computers

Annual fee: $2.

Design and operation of electronic computers.

Mr. Richard O. Endres, Rese Engineering Co., 731 Arch St., Philadelphia, Pa.

31 Transactions, EC-6, No. 2, 3; EC-7, No. 1, 2, 3, 4; EC-8, No. 1, 2, 3.

Engineering Management

Annual fee: $3.

Engineering management and administration as applied to technical, industrial and educational activities in the field of electronics.

Dr. Henry M. O'Bryan, Sylvania Elec. Products, 730 3rd Ave., New York 17, N.Y.

16 Transactions, EM-3, No. 2, 3; EM-4, No. 1, 3, 4; EM-5, No. 1-4; EM-6, No. 1, 2, 3.

Engineering Writing and Speech

Annual 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. T. T. Patterson, Jr., Chairman, RCA Bldg. 13-2, Camden, N.J.

4 Transactions, Vol. EWS-1, No. 2; EWS-2, No. 1, 2.

Human Factors in Electronics

Annual fee: $2.

Development and application of human factors and knowledge germane to the design of electronic equipment.

Mr. Curtis M. Jansky, Chairman, Royal McBee Corp., Port Chester, N.Y.

THE INSTITUTE OF RADIO

-IRE's 28 Professional Groups

the group chairman, and publications to date.

* Indicates publications still available

Industrial Electronics

Annual fee: $3.

Electronics pertaining to control, treatment and measurement, specifically, in industrial processes.

Mr. J. E. Eiselein, Chairman, RCA Victor Dev., Camden, N.J.

10 Transactions, *PGIE-1-3-5-6-7-8, 9, 10.

Information Theory

Annual fee: $3.

Information theory and its application in radio circuitry and systems.

Dr. Peter Elias, Chairman, MIT, Cambridge 39, Mass.

18 Transactions, PGIT-4, IT-1, No. 2-3; IT-2, No. 3; IT-3, No. 1, 2, 3, 4; IT-4, No. 1, 2, 3, 4; IT-5, No 1 2, 3

Instrumentation

Annual fee: $2.

Measurements and instrumentation utilizing electronic techniques.

Mr. L. C. Smith, Chairman, Image Electronics Inc., 2300 Washington St., Newton Lower Falls 62, Mass.

15 Transactions, 4; Vol. 1-6, No. 2, 3, 4; Vol. 1-7, No. 1, 2; Vol. 1-8, No. 1, 2.

Medical Electronics

Annual fee: $3.

The use of electronic theory and techniques in problems of medicine and biology.

Mr. W. E. Tolles, Chairman, Airborne Instruments Lab., Mineola, N.Y.

15 Transactions, 8, 9, 11, 12, ME-6, No. 1, 2, 3.

Microwave Theory and Techniques

Annual fee: $3.

Microwave theory, microwave circuitry and techniques, microwave measurements and the generation and amplification of microwaves.

Dr. A. A. Oliner, Microwave Research Institute, 55 Johnson St., Brooklyn 1, N.Y.

26 Transactions, MTT-4, No. 3-4; MTT-5, No. 3, 4; MTT-6, No. 1, 2, 3, 4; MTT-7, No. 2, 3.

Military Electronics

Annual fee: $2.

The electronics sciences, systems, activities and services germane to the requirements of the military. Aids other Professional Groups in liaison with the military.

Mr. Henry Randall, Chairman, Office of Asst. Secy. Defense, Pentagon, Washington, D.C.

7 Transactions, MIL-1, No. 1; MIL-2, No. 1; MIL-3, No. 2, 3, 4

Nuclear Science

Annual fee: $3.

Application of electronic techniques and devices to the nuclear field.

Dr. A. B. Van Rennes, Chairman, Bendix Aviation Corp., Detroit 35, Mich.

14 Transactions, NS-1, No. 1; NS-3, No. 2, 3; NS-4, No. 2; NS-5, No. 1, 2, 3, NS-6, No. 1, 2, 3.

Production Techniques

Annual fee: $2.

New advances and materials applications for the improvement of production techniques, including automation techniques.

Mr. L. M. Ewing, Chairman, General Electric Co., Syracuse, N.Y.

6 Transactions, No. 2-3, 4, 5, 6.

Radio Frequency Techniques

Annual fee: $2.

Origin, effect, control and measurement of radio frequency interference.

Mr. J. P. McNaul, Chairman, Signal Corps, USA's RDL, Ft. Monmouth, N.J.

1 Transaction, RF-1, No. 1.

Reliability and Quality Control

Annual fee: $3.

Techniques of determining and controlling the quality of electronic parts and equipment during their manufacture.

Mr. P. K. McElroy, Chairman General Radio Co., West Concord, Mass.

16 Transactions, *3, 5-6, 10, 11, 12, 13, 14, 15, 16.

Space Electronics and Telemetry

Annual fee: $2.

The control of devices and the measurement and recording of data from a remote point by radio.

Mr. C. H. Hoeppner, Chairman, Radiation, Inc., Melbourne, Fla.

13 Transactions, TRC-1, No. 2-3; TRC-2, No. 1; TRC-3, No. 2, 3; TRC-4, No. 1; SET-5, No. 1, 2, 3.

Ultrasonics Engineering

Annual fee: $2.

Ultrasonic measurements and communications, including underwater sound, ultrasonic delay lines, and various chemical and industrial ultrasonic devices.

Dr. Wilfred Roth, Chairman, Roth Lab., Hartford, Conn.

7 Transactions, PGUE, 5, 6, 7.

Vehicular Communications

Annual fee: $2.

Communications problems in the field of land and mobile radio services, such as public safety, public utilities, railroads, commercial and transportation, etc.

Mr. A. A. MacDonald, Chairman, Motorola, Inc., 4545 W. Augusta Blvd., Chicago 51, Ill.

13 Transactions, 5, 8, 9, 10, 11, 12, 13.

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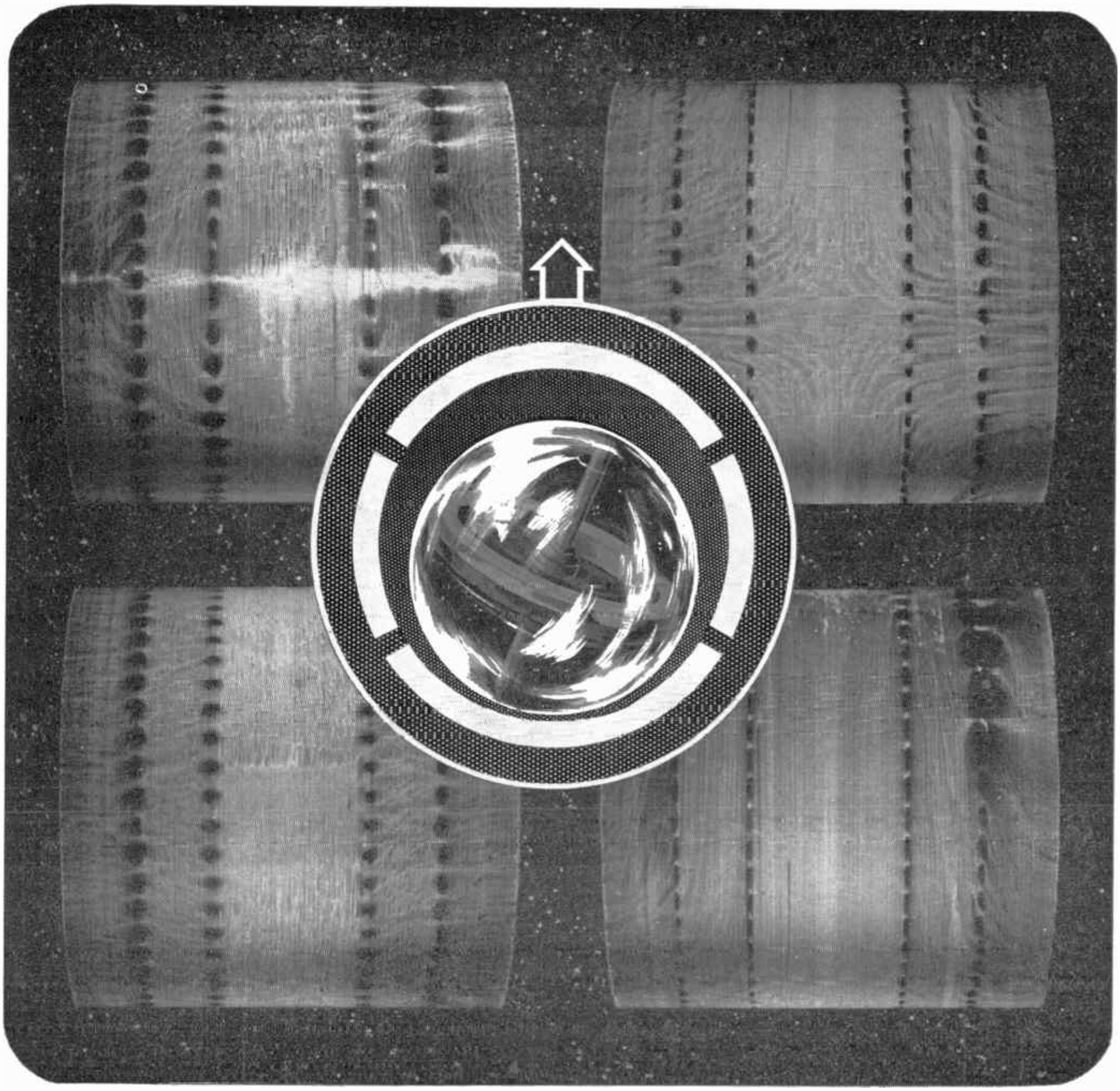
IMPORTANT DEVELOPMENTS AT JPL



GAS LUBRICATION

Research in gas lubrication and on performance and application of gas bearings is an important current activity at Jet Propulsion Laboratory.

The photographs shown are actual visualizations of gas flow patterns (obtained by an ultraviolet fluorescence technique) on a shaft under varying loads. Those on the left show pattern on an unloaded bearing — those on the right when bearing is loaded under 80 lbs. at 40 psig supply pressure.

These research experiments relate directly to the use of frictionless bearings in space vehicle components.

This is another example of the variety of supporting research and development being carried on at JPL to advance the national space exploration program.

CALIFORNIA INSTITUTE OF TECHNOLOGY

JET PROPULSION LABORATORY

A Research Facility operated for the National Aeronautics and Space Administration

PASADENA, CALIFORNIA

Employment opportunities for Engineers and Scientists interested in basic and applied research in these fields:

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Send professional resume, with full qualifications and experience, for our immediate consideration

IRE People

(Continued from page 92A)

Skydyne, Inc. announces the appointment of **William F. Maccallum** (A'59) as its Sales Manager. He has served in different capacities for over fifteen years with the company, beginning in the Purchasing Department, eventually becoming the Director of Purchases. He then became Assistant Sales Manager, the position which he held until he was appointed Head of the Sales Department.

❖

Appointment of a new management group at Schaevitz Engineering was announced today (Monday) by Herman Schaevitz, president of the electronic manufacturing company.

G. E. Merer

In the new organizational set-up, **George E. Merer** (S'47–SM'52) has been named as director of sales and marketing.

He was a staff engineer specializing in military electronics for Cubic Corporation. His experience includes ten years with the U. S. Air Force as a research and development engineer during which time he served as chairman of the Atlas Missile Weapons System Steering Committee and supervised numerous radar and communications projects. He has a degree in physics from Brooklyn College.

Mr. Merer is a member of the Armed Forces Communications and Electronics Association, American Ordnance Association, National Aeronautics Association and the Radio Technical Commission for Aeronautics.

❖

David L. Wyand (SM'53–M'57) has recently been appointed Manager, Customer Services at Eitel-McCullough, Inc., San Carlos, Calif.

Wyand was a member of the Government Marketing Department prior to this recent promotion. He joined Eimac in 1954 while enrolled in the cooperative work-study program at the University of California in Berkeley until 1956. During this period he worked in the development of Eimac's C-band amplifier klystrons and reflex klystrons. In 1957 he was graduated from that university with the B.S. degree in electrical engineering. He also received a bachelor's degree in marine engineering in 1951 from the California Maritime Academy in Vallejo, California. Prior to joining the Marketing Division at Eimac, Eyand served as a field engineer for the Westinghouse Electric Corporation in Baltimore, Maryland. From 1951 to 1953 he served as an Electronic Repair Officer in the U. S. Navy in the Far East.

❖

(Continued on page 120A)

4½" IMAGE ORTHICONS

provide better picture quality than
ever previously attained.
Pioneered, developed and manufactured by
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suppliers of Image Orthicons to the world.

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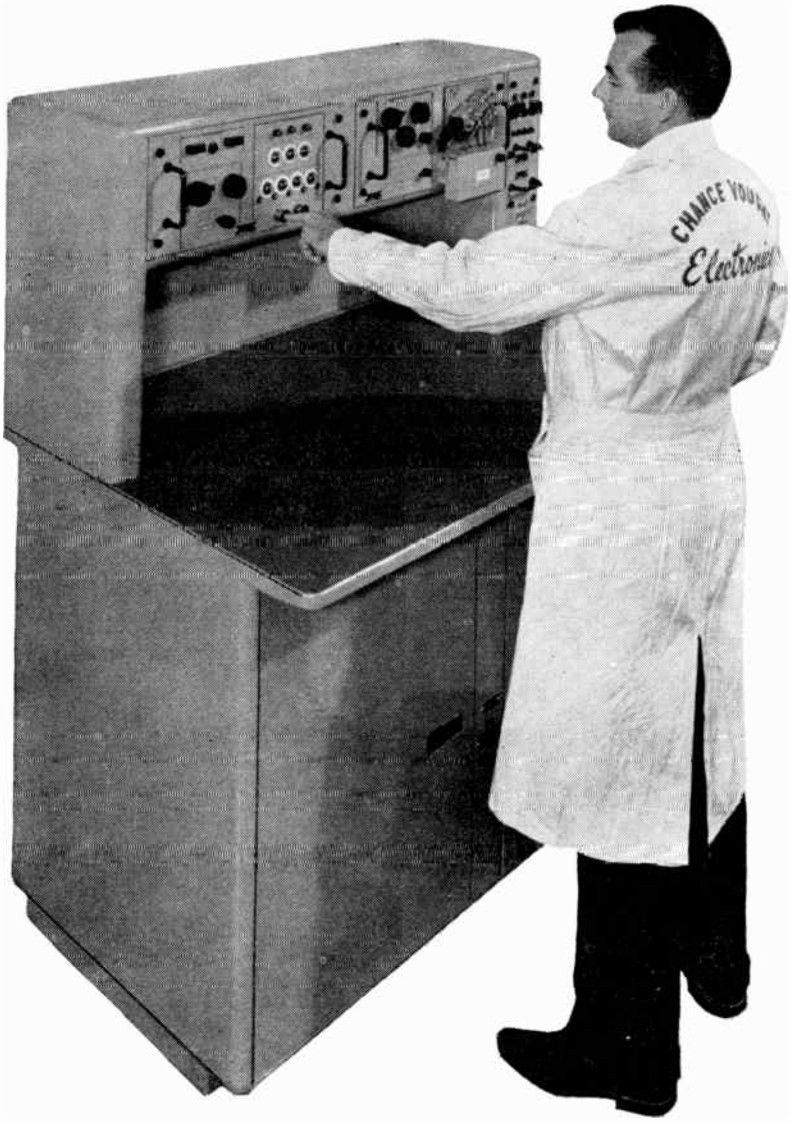
Have you gotten our letter about this test set ?



Transistorized . . . programmed with perforated tape — this set is a versatile means for testing drone radio command guidance systems. Its automatic features are typical of test and checkout equipment produced by Chance Vought's Electronics Division. These people have provided GSE support for radar and inertial guidance; flight stabilization; warhead arming and fuzing; rocket engine and telemetering systems. Altogether, over 4,500 articles of GSE equipment have been delivered in these and other programs.

This capability is available to the company or service that needs it. A recent letter from Vought Electronics describes experience, labs and representative products in detail. It was mailed to military agencies and weapons developers—primarily to guidance system developers.

If you haven't received this letter . . . if you want to know more about an experienced source of GSE support for small and large jobs, write:

VOUGHT ELECTRONICS

A DIVISION OF CHANCE VOUGHT AIRCRAFT, INCORPORATED • DALLAS, TEXAS

GROUND SUPPORT ELECTRONICS • ANTENNAS • POWER CONTROLS

sta-bil'i-ty

Poised in recovery or skimming straight and level — in any attitude — stability augmentation from American-Standard steadies the Kaman H-43B Huskie. This USAF rescue helicopter adds one more high performance vehicle to the growing list of applications for our subminiature inertial components. In aircraft — the F-106, in missiles — the Terrier and Tartar, and in submarines — the Polaris class, American-Standard, Military Products Division supplies gyros fundamental to accurate positioning and control.*

Company sponsored development in the Military Products Division has produced a family of gyroscopes, accelerometers and miniature components whose quality, performance and reliability make them unique in the field of navigation and stabilization systems.

Opportunities for Engineers

Winning this degree of acceptance has created new career opportunities in advanced design, production and applications engineering. American-Standard, Military Products Division, Norwood, Massachusetts.

*American-Standard and Standard® are trademarks of American Radiator & Standard Sanitary Corporation.

AMERICAN-Standard
MILITARY PRODUCTS DIVISION

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.

ELECTRICAL ENGINEER

Evaluate instrument systems, establish calibration and operation procedures. Procurement, acceptance testing and shakedown of new instruments required in wind tunnel testing of models and propulsion systems of rockets and missiles. Establish preventive maintenance procedures, monitor data quality and improve instrumentation. Send resume and salary requirements to Technical Employment, ARO, Inc., Tullahoma, Tenn.

TEACHING POSITION

Excellent teaching opportunity will be available beginning Feb.-Sept. 1960. Advanced degree required. Attractive full year contract available. Salary range for 9 months is $5,000 to $7,000. Location is in the midwest at a medium size private university. Complete resume to Box 2006.

INSTRUMENT OR ELECTRONIC ENGINEERS

Electronic or electrical engineer or physicist with instrumentation experience to do research and development work on the design, construction and evaluation of instrumentation for sampling and analysis of atmospheric air pollutants. Responsibilities for projects from design to prototype models. Salaries $5500, to $8200. Another, salary $10,130, in design and development of radiation measuring devices to direct program involving identification and measurement of radionuclides in environmental samples. Good prospects for growth and advancement. Federal Civil Service benefits and requirements. Apply U.S. Public Health Services, Robert A. Taft Sanitary Engineering Center, 4676 Columbia Parkway, Cincinnati 26, Ohio.

ENGINEERS

The Pratt & Whitney Aircraft Division of United Aircraft Corp., East Hartford, Conn. has openings for graduate engineers in the areas of Propulsion Systems Performance Analysis, Heat Transfer Research, Ultra High Temperature Materials, Dynamics, Vibrations, Structures Research, Experimental Testing, Technical Report Writing and Propulsion System Control Engineering. Many of these openings are in our advanced Development Groups where we are presently conducting studies in solid and liquid propellants, ion propulsion, arc jet, plasma jet, and other advanced forms of propulsion. Openings are available at both the Conn. and Florida facilities. For more information, contact Mr. Henry M. Heldmann, Employment Office.

PROFESSORS

Rank of Assistant Professor or Associate Professor, depending on qualifications. Salary $5500 to $8500 for session, 10 months nominal, 9 months actual. Start February or September 1960. Duties will include offering graduate courses and helping to develop research facilities. Opportunities for curriculum experimentation. Various sources of additional income available. Substantial allowance for relocation. Exceptionally good retirement plan. Fully accredited Electrical Engineering Dept. in medium sized university (700 undergraduate students in engineering) located in a very pleasant uncrowded city of 350,000. Address resume to Dean Otto Zmeskal, College of Engineering, University of Toledo, Toledo 6, Ohio.

INSTRUCTOR OR ASSISTANT PROFESSORS

Retirement of a staff member creates a vacancy in the Electrical Technology Dept. for February 1960. Teaching area is mainly in electronics at the technican level. Minimum requirements are BEE. or B.S. in E.E. and 2 years of industrial experience. Starting salary $5000 to $7000. Opportunity for an additional $2000 through evening and summer teaching. Excellent pension system and other fringe benefits. Write to Prof. J. De France, Dept. of Elec. Tech., New York City Community College, 300 Pearl St., Brooklyn 1, New York.

SCIENCE AND ENGINEERING

Opportunity at Robert College, Istanbul, Turkey for qualified men in engineering, mathematics, physics and chemistry interested in combining teaching and the development of limited research and consulting activities with the opportunity to live and travel in a vital part of the world: Strengthening staff, modernizing undergraduate engineering curricula, beginning graduate programs in engineering, developing undergraduate and later graduate programs in sciences, constructing new science and engineering building to prepare engineers for the industrial and technological development of Turkey and the Middle East. Address inquiries to Dean Howard P. Hall of the College of Engineering, Robert College, Bebek P.O. 8, Istanbul, Turkey, with copy to Near East College Assoc., 40 Worth St, Rm. 521, New York 13, N. Y.

DEVELOPMENT ENGINEER

Development Engineer to head a small development group in the field of small electronic components. Degree or its equivalent, and experience in this field required. Write Philadelphia Plant Employment, International Resistance Co., 401 North Broad St., Philadelphia 8, Pa.

STAFF IN ELECTRICAL ENGINEERING

Attractive positions combining teaching (graduate and undergraduate) and research are available. Appointment at any level from Assist. Professor up, depending on qualifications. A person with interests in electrical energy conversion and modern electrical machines is especialy sought, but openings in other fields are available. Applicants must have Ph.D. or equivalent evidence of research potential. Write Chairman, Div. of Engineering, Brown University, Providence 12, R. I.

TECHNICAL SALES ENGINEER

Knowledge of Government operations and experience in microwave tube field desired. Retired service personnel would be considered. Good position in growing company. Please call or write American Radio Company, Inc., 445 Park Ave., New York 22, N.Y. Pl-3-5046.

PRODUCTION FOREMAN

Production Foreman—Electronic Transformer: 5 years (+) experience in coil winding business. Familiar with general machineshop equipment and vacuum potting techniques. New company. Excellent opportunity. Sunny San Diego. Apply Atlas Transformer Co., 1839 Moore St., San Diego 1. Calif.

ASSOCIATE PROFESSOR

Electrical Engineering faculty being expanded in rapidly growing department, graduating first

(Continued on page 110A)

Positions Open

(Continued from page 109A)

class in 1960. Current positions available to rank of Associate Professor and to 9 months salary of $6000, depending upon education and experience. Background emphasis preferably upon electronics and advanced circuit theory. Opportunity for research and other industrial programs in the area. Send full background to Chairman, Electrical Engineering, University of Bridgeport, Bridgeport, Conn.

ENGINEERS

ASSISTANT DIRECTOR RESEARCH & DEVELOPMENT: E.E. graduate with an advanced degree, who has history in electronics, audio, transducers, pulse circuitry and electromechanical devices. Supervisory experience desired in organizing and planning R&D programs.

CHIEF RESEARCH ENGINEER: Electronic engineering graduate trained in audio, video and transistor circuits. Supervisory experience in R&D planning.

SENIOR MECHANICAL ENGINEER: Experienced in product design & development of small mechanisms, precision and electromechanical devices, speed reduction systems.

SENIOR & JUNIOR ELECTRONIC ENGINEERS will find ample opportunities and challenges in Gray Manufacturing Co., 16 Arbor St., Hartford, Conn.

ASSISTANT PROFESSOR

Assistant Professor of Electrical Engineering, University of North Dakota, Grand Forks, North Dakota. Position open September 1960. Must have M.S. degree and some experience in teaching or industry. Will teach electronic and circuitry courses to undergraduates primarily, with some graduate teaching available, if desired. Submit resume to Chairman, Electrical Engineering Dept.

ASSISTANT & ASSOCIATE PROFESSORS

Assistant & Associate Professors, Ph.D., 200 graduate students. Ideal dry mountain climate. Income $10,000 up with research. Chairman, E.E. Dept., University of New Mexico, Albuquerque, New Mexico.

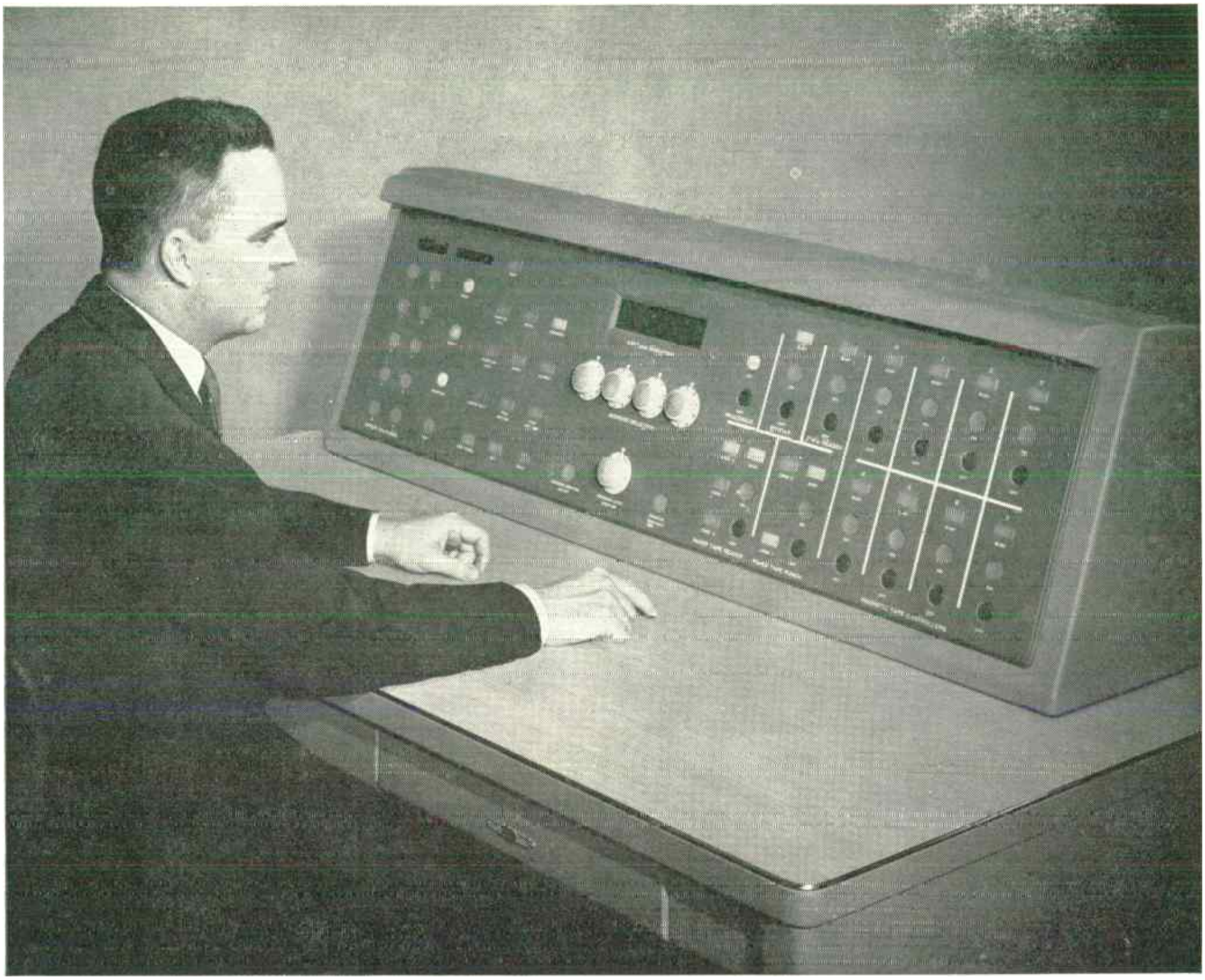
ASSISTANT & ASSOCIATE PROFESSORS

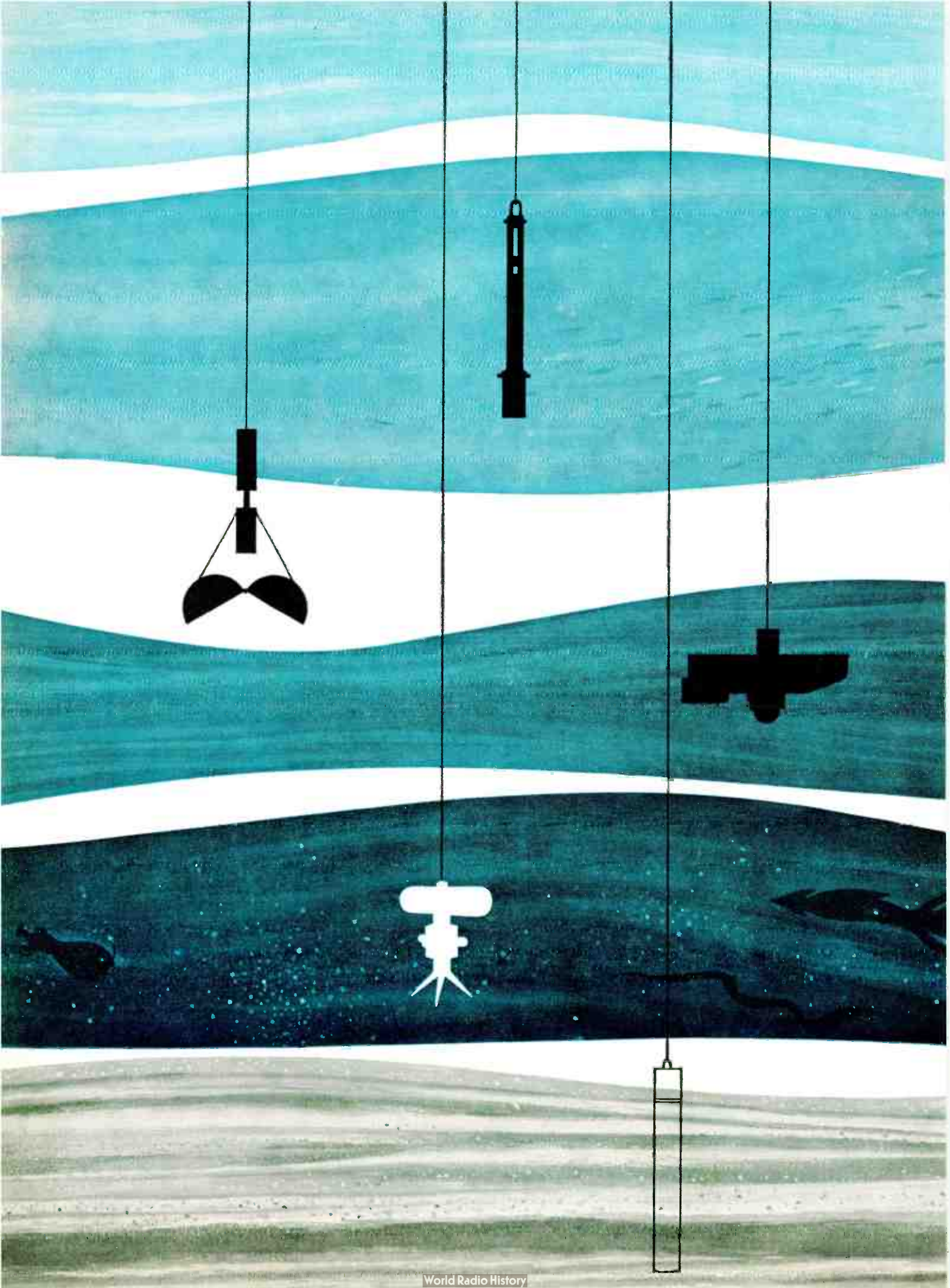
Applications are invited for Assistant & Associate Professors of E.E. Candidates should be well qualified academically and should have some research experience, preferably in Control Systems, Solid-state devices, Microwaves, Telecommunications. Salary scales are competitive with industrial and research establishments. Additional stipends are offered for summer research work. Address applications to Chairman, Dept. of E.E., McMaster University, Hamilton, Ontario, Canada.

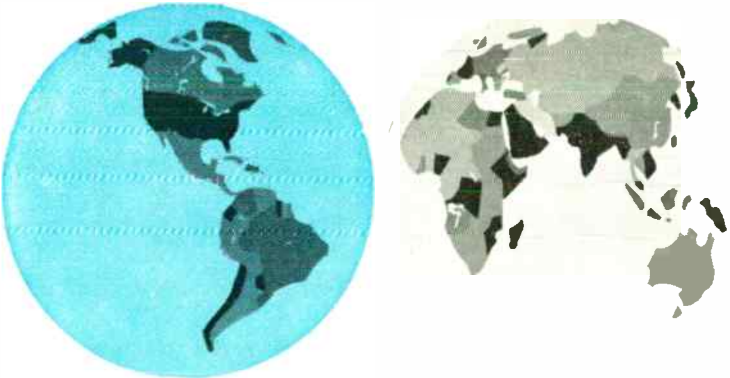
STAFF ENGINEER

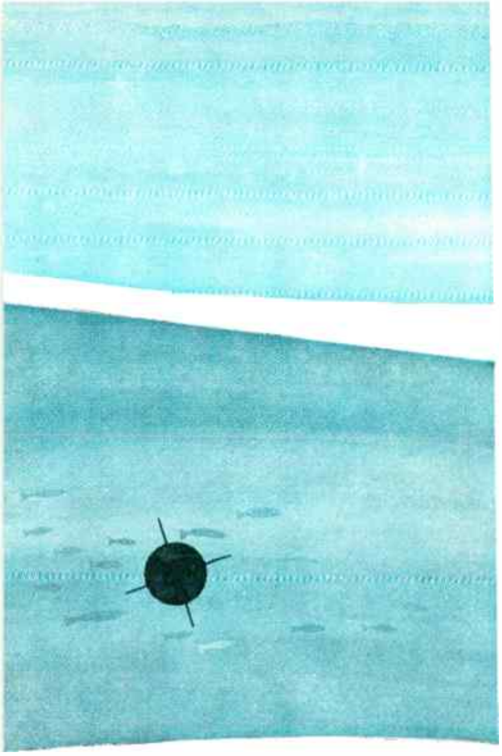
A high level staff engineering position is available for an experienced engineer who desires a position without line responsibilities. The position requires ability to study systems and circuits proposed and under development with a view to steering engineering effort along productive paths. A superior educational background and considerable experience are required in carrier telephone, electronic switching, microwave systems, and related circuitry. Salary open. Northern California area. All replies will be kept confidential. Reply to Box 2008.

(Continued on page 114A)





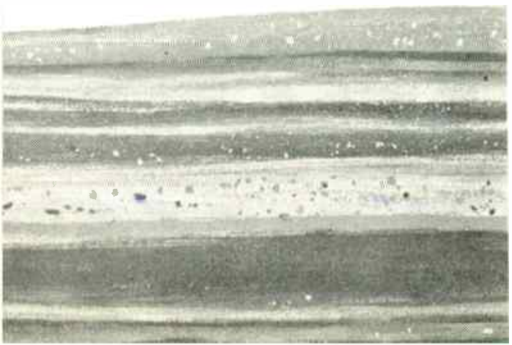




Lockheed's interest in the virtually unknown 360,000,000 cubic miles of this planet's oceans, stems naturally out of its underwater environmental development work with the Navy's POLARIS Fleet Ballistic Missile.

Proposed studies in the increasingly important field of oceanography include: oceanographic research vessels; measuring instruments; data collection systems; underwater communication and navigation; and basic research regarding natural phenomena and military aspects of the deep sea.

EXPLORING THE WORLD OF WATER



Division Diversification—Oceanography is typical of Lockheed Missiles and Space Division's broad diversification. The Division possesses complete capability in more than 40 areas of science and technology—from concept to operation. Its programs provide a fascinating challenge to creative engineers and scientists. They include: celestial mechanics; computer research and development; electromagnetic wave propagation and radiation; electronics; the flight sciences; human engineering; magnetodynamics; man in space; materials and processes; applied mathematics; operations research and analysis; ionic, nuclear and plasma propulsion and exotic fuels; sonics; space communications; space medicine; space navigation; and space physics.

Engineers and Scientists—Such programs reach far into the future and deal with unknown and stimulating environments. It is a rewarding future with a company that has an outstanding record of progress and achievement. If you are experienced in any of the above areas, or in related work, we invite your inquiry. Please write: Research and Development Staff, Dept. A-33, 962 W. El Camino Real, Sunnyvale, California. U.S. citizenship required.

Lockheed

MISSILES AND SPACE DIVISION

Systems Manager for the Navy POLARIS FBM; the Air Force AGENA Satellite in the DISCOVERER Program; MIDAS and SAMOS; Air Force X-7; and Army KINGFISHER

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Positions Open

(Continued from page 110A)

ELECTRONICS INSTRUCTOR

Post High School Institution—Degree required—Permanent, hospitalization and non-contributory pension system provided. Start February 1, 1960. Write giving complete resume and salary required to New York Trade School, 304-326 East 67 St., New York 21, N. Y. Att: Director, Electronic Training.

MICROWAVE SPECIALIST

Microwave physicist needed for applying microwave techniques to the study of plasma flows and ionized regions around high speed models and in shock tubes. Measurements and study of the radio-frequency energy emitted by the passage of high-speed models and of the transmission and reflection characteristics of the wake are required in order to evaluate the effects of these characteristics and also as an aid to further the knowledge of flow phenomena at extreme speeds. Applicant should have advanced degree with a good background in microwave propagation and field theory as well as ability to work with microwave hardware. He should be capable of taking the initiative in the application of microwave techniques and in the interpretation of results. Write Personnel Officer, NASA, Ames Research Center, Moffett Field, Calif.

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.

FIELD TECHNICAL REPRESENTATIVE

Desires position as Technical Representative or position with Civil Service in Japan. Have 3 months in grade as Electronic Technician GS-9. Career status. Have 2½ years experience as Technical Representative. Member of IRE. Have 1st Class FCC Radiotelephone license. Considerable and varied electronics experience. Versatile. Box 1089 W.

ENGINEER

4½ years military field engineering on aircraft control and warning, airborne fire control, and range instrumentation radar systems.

(Continued on page 116A)

Positions Wanted

By Armed Forces Veterans

(Continued from page 114A)

4 years AF radar repair on AC & W Systems. Desires Field Manager position with field service organization in U.S. or Far East. Will consider good assignment as field service representative. Age 28, married, 1 child. Box 2040 W.

ENGINEER

Engineer with B.S., M.S., and E.E. degrees with 15 years experience in micro-wave tube and circuit design and development. Desires project responsibility in Southern California area. Married; served in Navy as Radar Officer. Box 2041 W.

ELECTRONIC ENGINEER

BSEE. 1956. 1/Lt. USAF. Familiar with jet aircraft navigation problems. Knowledge of bomb-nav. systems. Some design experience in bench power for missile components. Top secret clearance. Desires position in San Francisco Bay area. Box 2042 W.

ENGINEERING MANAGER

16 years experience in radar, control, and computers. Last 5 years manager of computer design. Prominent through extensive publications and significant patent in digital computer systems. Advanced degree in E.E. Sigma Xi. Senior Member IRE. Seeking increased responsibility. Under 40, married. Box 2046 W.

TECHNICAL WRITER

4 years handbook experience in digital computers, missiles, data processing; 2 years in field service. Desires position in Western Europe. Age 34, married. BA.; secret clearance; 4 languages. Box 2047 W.

ADMINISTRATIVE ENGINEER

Engaged in project administration in connection with R&D activities. Experience involved multi-million dollar subcontract contractual and technical administration and development of administrative operating and control procedures. BBa., MBA. 9 years diversified engineering. Engineering includes test equipment designs and systems engineering. Schooled in military electronics. Desires responsible administrative position with a growing future. Box 2048 W.

SENIOR ELECTRONICS ENGINEER

BEE., MEE. Had project responsibility in audio, video, computer circuits and equipment; conscientious supervisor. Desires managerial responsibility and substantial challenge. Box 2058 W.

ELECTRONICS ENGINEER

BEE. 1952. MEE. 1955. Age 31. Desires Project Engineering or managerial position with growth potential. 4 years experience, to Project Engineer level, designing and developing large radar systems. Some teaching and research experience. Former Signal Corps officer. Licensed Professional Engineer in New York. Box 2059 W.

ENGINEER

Desires position not wholly technical involving possible overseas travel. LTJG USNR. Tau Beta Pi. Eta Kappa Nu. Unmarried. BS. and MS. in E.E. from large midwestern universities. Box 2060 W.

IRE People

(Continued from page 104A)

Carl Pilnick (A'54–M'55–SM'57) has been elected vice president and director of research and development at Consolidated Avionics Corporation, according to an announcement by Harry R. Glixon, the firm's president.

C. Pilnick

In his new post, he is responsible for technical supervision of all development projects undertaken by the Westbury subsidiary of Consolidated Diesel Electric Corporation. The firm is active in the fields of automatic tset equipment, data reduction systems, low frequency electronic equipment and transistorized power supplies.

Joining Avionics' engineering staff in 1957, he was promoted to senior project engineer and chief staff engineer prior to becoming vice president. He has the B.S. degree in electrical engineering from the College of the City of New York and the M.S. degree from Stevens Institute of Technology.

Before joining Consolidated Avionics, Mr. Pilnick was associated with the Teleregister Corporation in Stamford, Conn., and Bruno-New York Industries Corporation in New York, N. Y., among others.

❖

Harry E. Schauwecker (A'53) is the newly elected President of Valor Instruments, Inc., Gardena, Calif. Previously, as Director of New Products Development, he was responsible for the development of the expanding line of transistorized instruments.

H. Schauwecker

Prior to joining Valor, he was a transistor specialist at Gilfillan Brothers Inc. and Bell Telephone Laboratories. A lecturer in Engineering at U.C.L.A., he has taught an advanced course on Transistor Applications for several years.

❖

Dr. A. E. Middleton (A'52–M'58), a pioneer in the study of semiconductors and the electrical properties of germanium, has joined the research staff of The Harshaw Chemical Company. His new duties will be as Senior Research Adviser on solid state physics, an area increasingly important to many of the products manufactured by Harshaw.

(Continued on page 122A)

IRE People

(Continued from page 120A)

Dr. Middleton received the Ph.D. degree from Purdue University in 1944. His thesis was concerned with the resistivity, Hall effect, and thermoelectric power of germanium, germanium alloys, silicon and tellurium. Since that time he has continued work in solid state physics. He was one of the organizers of the semi-conductor section of the Electrochemical Society.

❖

Personnel shifts in the engineering department at Levinthal Electronic Products, Subsidiary Radiation Inc., have

A. J. MORRIS

H. G. HEARD

moved **Albert J. Morris** (M'47–SM'51) to the post of senior vice president, engineering. Previously vice president and chief engineer, he was co-founder with Dr. Elliott Levinthal of the company, which became part of Radiation Inc. last May.

The post of chief engineer has been filled by the appointment of **Harry G. Heard** (S'47–A'50–M'56). Formerly on the staff of the University of California Radiation Laboratory, he was the research engineer responsible for operation and development of the 6.4 BEV Bevatron and was responsible for coordinating the efforts of all scientific and engineering groups for the development and utilization of the accelerator as a basic research tool. He was also credited with investigative work and developments leading to fundamental advances in the effectiveness of this machine.

A graduate of the University of California with both a B.S. and M.S. in electrical engineering, Mr. Heard holds memberships in Sigma Xi, Tau Beta Pi, Eta Kappa Nu, and the American Physical Society.

❖

Reuben O. Schlegelmilch (A'42–M'44–SM'48) has been appointed technical director of the defense products group of Westinghouse Electric Corp. Prior to this appointment, he was director of research and development at the Air Force's Rome Air Development Center. In his new post he will have his office at the Westinghouse defense products group's headquarters in Washington, D. C.

He received the B.S. degree from the University of Wisconsin and the M.S. degrees from Rutgers University and M.I.T. He now is completing his dissertation for the Ph.D. degree in electrical engineering from Syracuse University.

(Continued on page 126A)

IRE People

(Continued from page 122A)

He has also served as an instructor at Rutgers and Cornell Universities, and with the engineering experimental station at the University of Illinois. Before joining RADC, he worked on micorwave radar development at Research Enterprises, Ltd. (Canada), and at the M.I.T. radiation laboratories.

Mr. Schlegelmilch is a member of the AIEE and the American Management Association.

⁂

Packard Bell Computer Corporation has named **Theodore J. Smith** (S'54–A'56) sales manager, it has been announced by Max Palevsky, director and vice president.

T. J. SMITH

Mr. Smith holds the B.S. degree in engineering from the University of Maryland, and is a graduate of the Air Force Ground Electronics School at Keesler Air Force Base, Miss. He has also completed several courses in business administration and electrical engineering at the University of Mississippi and the University of California at Los Angeles.

Before joining Packard Bell Computer, he served as chief applications engineer for the Nuclear Division of American Electronics, Los Angeles, Calif. Previously, he was an assistant to the vice president for engineering at Gilfillan, Inc., Los Angeles.

⁂

In an expansion of the Van Groos Company, J. C. Van Groos announces the appointment of **Carl Sutliff** (S'55–M'59) to the position of Field Engineer. He has been associated with Bendix Aviation, Pacific Division, Associated Missiles, a division of Marquardt, and most recently has worked as a sales engineer in the Los Angeles area.

Mr. Sutliff obtained the B.S. degree from California State Polytechnic College.

⁂

Eugene N. Torgow (S'49–A'49–SM'54) has joined the Engineering Division of the Polytechnic Research and Development Co., Inc., as a Department Head of Special Products.

E. N. TORGOW

He was born on November 26, 1925 in New York, N. Y. He received the B.E.E. degree from the Cooper Union School of Engineering, New York, N. Y., in 1946, and the M.E.E. degree from the Polytechnic Institute of Brooklyn, N. Y., in 1949.

(Continued on page 128A)

professional opportunities at Honeywell Aero

INERTIAL SYSTEM DEVELOPMENT

Systems Analyst—employs mathematical techniques such as operational calculus, matrix algebra, and difference equations to the solution of problems concerning performance characteristics of various system configurations including analysis for error introduced by sensors and computer, requirements for alignment, and optimization of the system configuration.

Digital System and Logic Designer—requires familiarity with capabilities of various digital computer configurations and ability to employ system and logic relations in specifying necessary configuration for solving inertial navigation problem.

Electronic and Mechanical Designers—engineers with background in transistor circuitry, inertial sensor development and evaluation, and precision mechanical equipment design are needed to perform component development and evaluation, and to design mounting and alignment equipment.

APPLIED RESEARCH

Programmer Analyst—mathematician with experience in the use of medium and large scale digital computers for analysis of scientific problems.

Human Factors Engineer—capable of analysis and direction of experiments in human motor skills, and application to man-machine systems involving automatic control techniques.

Systems Analyst—capable of conducting research studies involving new techniques of space navigation and guidance.

DESIGN AND DEVELOPMENT

Flight Control Systems—analytical, systems, and component engineers to work in areas such as advanced flight reference and guidance systems. Positions range from analyzing stability and control problems, systems engineering—through design, testing, and proof of electrical and mechanical equipment —including flight test and production test.

Advanced Gyro Design—Engineers with two and up to twenty years' experience in precision gyro and accelerometer development, servo techniques, digital techniques, solid state electronic development, advanced instrumentation and magnetic component design.

Electronic Circuit Designers—experienced in the areas of analog/digital computers, transistor circuits, servos, instrumentation, and/or gyro stabilization.

For the less experienced professional engineer, there are opportunities in the Evaluation Laboratory which lead to careers in any of the above fields.

To investigate any of the above professional opportunities at the Aeronautical Division, please write in confidence to Bruce Wood, Dept. 281B

To explore professional opportunities in other Honeywell operations coast to coast, send your application to H. K. Eckstrom, Honeywell, Minneapolis 8, Minnesota.

Honeywell H HONEYWELL

AERONAUTICAL DIVISION

1433 Stinson Blvd. NE, Minneapolis 13, Minnesota

IRE People

(Continued from page 126A)

During 1946–47, he served with the U. S. Army Air Force, doing radar repair and maintenance in the Pacific Theater. In 1948, he jointed the staff of the Microwave Research Institute of the Polytechnic Institute of Brooklyn, where he worked on research and development of microwave attenuators and power measuring devices. From 1951 to 1953, he established and supervised a microwave laboratory at the Allen B. Du Mont Laboratories, East Paterson, N. J.

In 1953, he returned to the Microwave Research Institute and was engaged in strip line and waveguide filter development. He became leader of the Component Research Section in 1957.

❖

Two Northern California electronics executives, **Calvin K. Townsend,** (A'49) and Dr. **John V. N. Granger** (S'42–A'45–M'46–SM'51–F'56), have joined the board of directors of WESCON.

Mr. Townsend has been named by the Western Electronic Manufacturers Association and Dr. Granger was chosen by the San Francisco Section of the IRE.

Mr. Townsend, a native of Lima, Ohio, is executive vice-president and chairman of the board of Jennings Radio Manufacturing Corp., San Jose, Calif. He has served twice on the board of the Western Electronic Manufacturers Association and was its president in 1957.

A graduate of the University of California, he was general manager of the Aircraft Accessories Corp., Kansas City Division, during World War II. He returned to California in 1944 to become general partner and manager of Jennings Radio, which was formed in 1942.

He serves on the boards of Electro Engineering Works of San Leandro, Calif., the San Jose Rotary Club and the Greater San Jose Chamber of Commerce.

Dr. Granger is president of Granger Associates, Palo Alto, Calif. He is a native of Marion, Iowa.

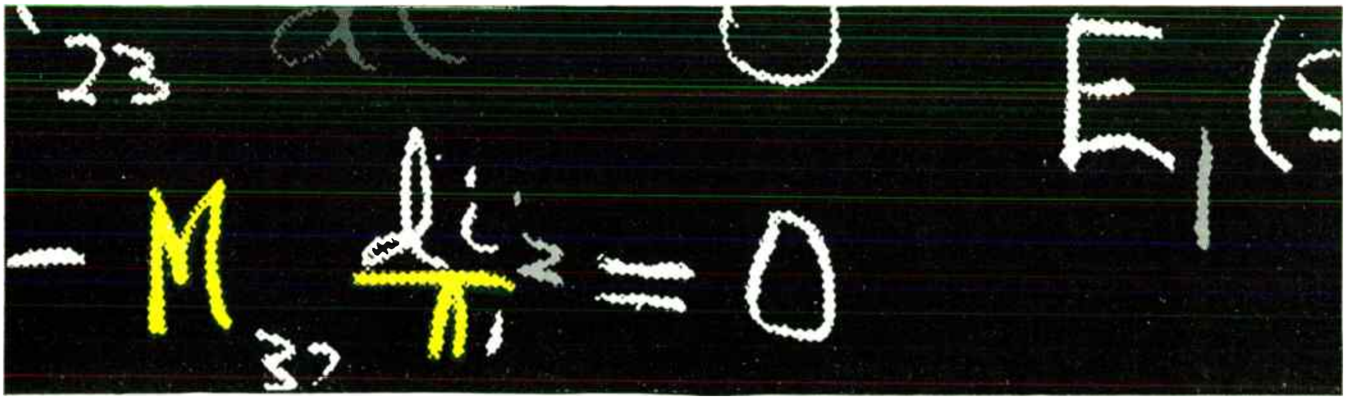
He graduated from Cornell College in Iowa and is presently a trustee of the college. He earned the M.S. degree in communications engineering at Harvard University and early in World War II was with Harvard's Radio Research Laboratory. In 1944 he joined the American-British Laboratory at Great Malvern, Eng.

He entered Harvard again in 1946, becoming a research fellow and group leader in the Electronic Research Laboratory. He received the Ph.D. degree from Harvard in 1948.

He joined Stanford Research Institute in the spring of 1949 to supervise the antenna research program. He became assistant chairman of SRI's Engineering Division and head of the Radio Systems Laboratory.

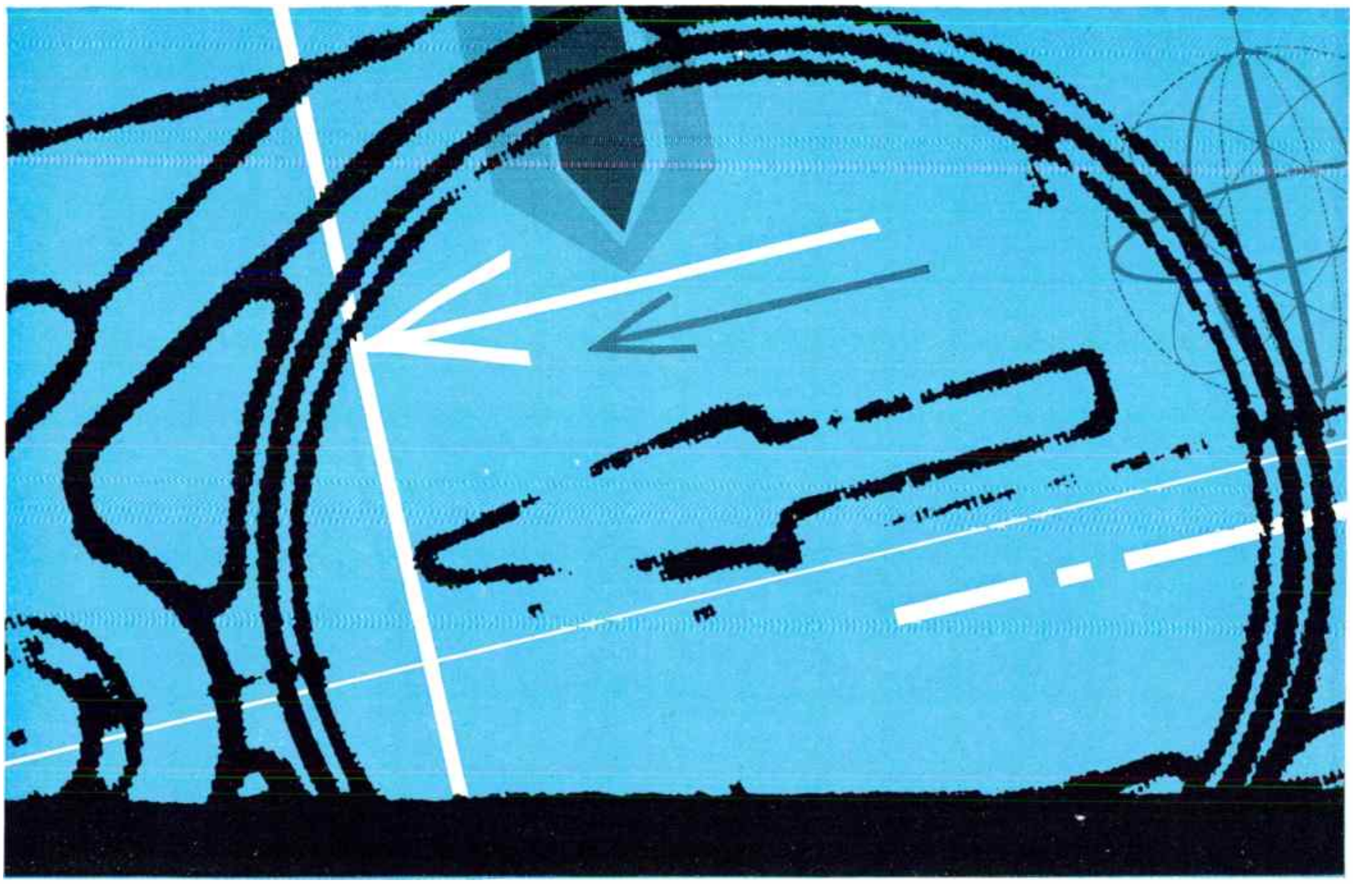
For his professional achievements, Dr. Granger was named "The Outstanding Young Electrical Engineer for 1952" by

(Continued on page 132A)



General Motors pledges

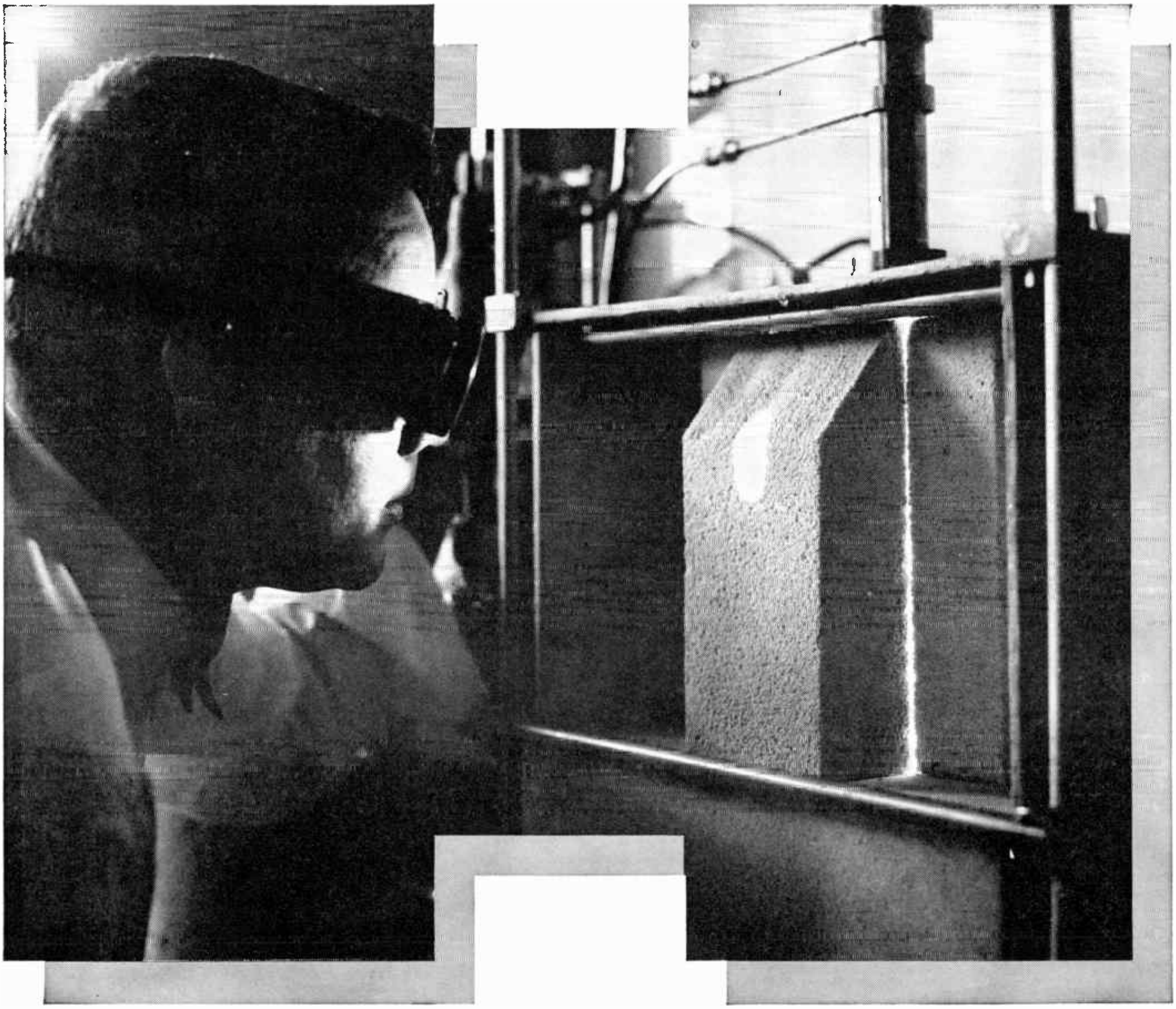
AC QUESTMANSHIP



AC Seeks and Solves the Significant—AC Design and Development is moving far ahead in new technology—the result of GM's commitment to make ever larger contributions to the defense establishment. AC plans to resolve problems even more advanced than AChiever inertial guidance for Titan / This is AC QUESTMANSHIP. It's a scientific quest for the development of significant new components and systems . . . to advance AC's many projects in guidance, navigation, control and detection / Dr. James H. Bell, AC's Director of Navigation and Guidance, sees this as a "creative challenge". His group takes new concepts and designs them into producible hardware having performance, reliability and long life. He strongly supports the fact that an AC future offers scientists and engineers "a great opportunity to progress with a successful and aggressive organization" / If you have a B.S., M.S., or Ph.D. in the electronics, scientific, electrical or mechanical fields, plus related experience, you may qualify for our specially selected staff. If you are a "Seeker and Solver", write the Director of Scientific and Professional Employment, Mr. Robert Allen, Oak Creek Plant, Box 746, South Milwaukee, Wisconsin.

GUIDANCE / NAVIGATION / CONTROL / DETECTION / AC SPARK PLUG *The Electronics Division of General Motors*

alchemy in the 20th



Development *of new and highly reliable direct display storage and microwave tubes is being carried on at the Electron Tube Division, Hughes Products.*

century

In medieval times, alchemists spent lifetimes trying to transform the commonplace into the precious without success. Today, in the Hughes Research Laboratories, scientists and engineers using advanced methods and equipment, are synthesizing many new and precious materials to be used in solid state research.

The photograph, for example, shows a Hughes Research Laboratories scientist observing the growth of an yttrium aluminum garnet crystal using the flame fusion method. This work, as well as other crystal growth techniques, is part of a major Hughes materials research and supply program for the synthesis of ruby, sapphire, spinel, ferrite, garnet and related oxide single crystals. These materials provide atomic circuit elements which are the key to quantum electronics.

As solid state research at the Hughes Research Laboratories continues to expand and intensify, a supply of new and tailor-made materials in the form of high-quality single crystals is essential. Effective utilization and improvement of existing and new crystal growth methods requires extensive knowledge of their range of applicability, crystal growth mechanisms and the relationships between growth parameters and perfection of the resulting materials.

Materials research is contributing significantly to existing ferrimagnetic, paramagnetic and absorption spectroscopic studies...and is opening new areas of investigation by providing materials not previously available as large single crystals.

Other Hughes activities cover practically every part of the electronics spectrum...providing stimulating outlets for creatively-oriented engineers. These include: Space Vehicles, Nuclear Electronics, Ballistic Missiles, Advanced Data Handling and Display Systems, Infrared Devices, Three-Dimensional Radar...and many others.

The variety and advanced nature of Hughes projects provides an ideal environment for the scientist or engineer who wishes to increase his professional stature.

Feedback *in the form of performance data and suggestions for modifications to advanced Hughes Systems is provided by Hughes Field Engineers.*

Newly instituted programs at Hughes have created immediate openings for engineers experienced in the following areas:

Electroluminescence	Field Engineering
Infra-red	Equipment Engineering
Plasma Physics	Commercial Sales
Digital Computers	Microwave & Storage Tubes
Systems Design & Analysis	Communications Systems
Circuit Design & Evaluation	Micro Electronics

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Hughes General Offices, Bldg. 6-E1, Culver City, Calif.

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IRE People

(Continued from page 128A)

Eta Kappa Nu and in 1955 received the annual recognition award of the Seventh Region of the IRE.

His memberships include the Research Society of America, the Institute of Aeronautical Sciences and Sigma Xi. He is on the Board of Governors of the Committee for Art at Stanford.

❖

Dr. Peter B. Myers (A'51–M'56) has accepted a position as Staff Scientist with the Semiconductor Division of Motorola Inc. In his new position, he will be engaged in advanced research and development of semiconductor integrated circuits, and will serve as a division-wide consultant on mechanical problems.

P. B. MYERS

For the past eight years he has been associated with Bell Telephone Laboratories doing research work on instrumentation for high frequency characterization of transistors, application and characterization of thin evaporated magnetic films and studies of symmetrical silicon alloyed junction transistors for cross-point use.

Dr. Myers received the Ph.D. degree in Physics from Oxford University Eng., in 1950 and is a member of the Society for Quality Control, the Philosophy of Science Association, the New York Academy of Sciences and Sigma Xi.

❖

Dr. John R. Whinnery (A'41–SM'44–F'52), professor of electrical engineering, has been appointed Dean of the College of Engineering on the Berkeley campus of the University of California, Chancellor Glenn T. Seaborg has announced.

J. R. WHINNERY

He succeeds Dean Morrough P. O'Brien, who retired from the position on June 30, 1959, after 15 years as head of the College.

Under its new administrative chief, the College will continue to place major emphasis on the scientific basis of engineering—a policy which has been in effect for many years. Continued stress will also be placed on the extensive program of graduate studies and research.

(Continued on page 135A)

Top Openings
for

ELECTRONIC ENGINEERS

and

SENIOR ELECTRONIC ENGINEERS

Must have experience in the development and design of semiconductor circuitry, analogue or digital components and sub-systems and/or data handling, conversion and processing equipment. Detailed knowledge of many of the following transistorized modules is required: modulators, demodulators, choppers, inverters, converters, power supplies, DC and AC amplifiers, flip-flops, multivibrators, squarers, pulse amplifiers, pulse gates, etc. Degree or advanced degree in electronics is preferred.

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SEND YOUR RESUME

Please send resume of your qualifications at the earliest opportunity to Director, Engineering Personnel, Dept. P-2, Electronic Systems Development Corporation, 1484 E. Main Street, Ventura, California.

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- Advanced Pulse and Video Circuit Development
- Advanced Inertial Navigational System Development
- Optical and Infra-Red Equipment Engineering
- Electron Optics

MECHANICAL DIVISION General Mills

IRE People

(Continued from page 132A)

In his new position, Dean Whinnery will be responsible for the education of the 1809 undergraduates and 589 graduate students enrolled in engineering at Berkeley during the current semester. As an example of the growth of the College in recent years, comparable enrollment figures for 1953 were 1415 and 199.

Dean Whinnery came to his new position after three years as chairman of the Department of Electrical Engineering at Berkeley. He received the B.S. degree on the Berkeley campus in 1937 and then spent the following nine years performing research in microwave electronics with the General Electric Company.

In 1946, he returned to the Berkeley campus and obtained his doctorate there two years later. His faculty appointments at Berkeley have included lecturer in 1946, associate professor in 1948, and professor of electrical engineering in 1952.

In 1951–52, he took a leave of absence from the University to serve as head of microwave tube research for the Hughes Aircraft Co. During the 1958–59 academic year he studied noise problems in microwave tubes at the Swiss Federal Technical Institute on a Guggenheim Fellowship.

More than half of his research career has been devoted to the study of microwave tubes. The remainder of his work has centered on microwave circuits and antennas. With Dr. M. R. Currie, he was responsible for the development of the cascade backward-wave amplifier. He has also made extensive contributions to the understanding of noise problems in microwave amplifiers.

❖

Dr. R. M. Warner, Jr., (SM'59) has joined the Motorola Semiconductor Products Division in Phoenix, Ariz. as Chief Engineer, Diode Development. He spent the preceding seven years in semiconductor device development at Bell Telephone Laboratories, Murray Hill, N. J. Dr. Warner is co-inventor of the field effect tetrode and current limiting diode, as well as several other semiconductor devices.

He was born in Barberton, Ohio, and received the B.S. degree in physics from Carnegie Institute of Technology in 1947, and the M.S. and Ph.D. degrees in physics from Case Institute of Technology in 1950 and 1952, working in nuclear physics. He served as physics instructor at both institutions and worked in the Pittsburgh Plate Glass Co. and the Corning Glass Works for a total of two years. He spent over 3 years in the Army Signal Corps, serving as radio officer during World War II in both overseas theatres. His affiliations include the American Physical Society, American Association for the Advancement of Science, Sigma Xi, Tau Beta Pi, and Phi Kappa Phi.

❖

(Continued on page 136A)

IRE People

(Continued from page 135A)

Sidney Weiser (A'51) has been appointed as Director of Engineering for USI Robodyne, a division of U. S. Industries, Inc.

S. WEISER

He has been chief engineer at USI Robodyne and was formerly director of research for Federal Manufacturing and Engineering Co. He has an extensive background in the research, development, design and production of a wide variety of products and systems. They include magnetic tape recorders, cameras, microphones, automatic slide changers, actuating mechanisms, air conditioning apparatus, electric heat generators, and mechanical presses with interlocking controls.

His participation in Defense programs includes responsibility for: production of "Matador" Missile Guidance System; development, design and production of the Test Console for the M-1 Bombing System Computer; Reliability Testing Consoles for the "Hustler" Bomber Fire Control System; and Sidereal Clock and Stability Test Consoles for Gyros used on the "Polaris" Missile System.

Mr. Weiser is a graduate (cum laude) of the Polytechnic Institute of Brooklyn and has done graduate work in production control, physics, optics, photography, VHF techniques, and advanced theory of ultrashort electromagnetic waves at the same institution and at Pratt Institute. He is a member of the American Society of Mechanical Engineers, the Society of Motion Picture and Television Engineers, the Armed Forces Communications Association and Tau Beta Phi, the honorary engineering society. His patents, granted and pending, include a microscope with Frictional Vernier Drive, a linear actuating device, and TransfeRobot equipment which automates assembly, inspection, gauging and packaging of a wide variety of products.

Professional Group Meetings

AERONAUTICAL AND NAVIGATIONAL ELECTRONICS

Philadelphia—October 22

"Bendix Doppler Navigation System," W. A Visher, Bendix, Radio Div.

(Continued on page 138A)

Professional Group Meetings

(Continued from page 136A)

Antennas and Propagation

San Francisco—October 13

"A VLF Satellite Experiment," R. A. Helliwell, Stanford Univ.

Antennas and Propagation Microwave Theory and Techniques

Columbus—October 27

"Recent Advances in Radio Astronomy and Space Communications," J. D. Kraus, The Ohio State Univ.

Philadelphia—September 29

Tour of Philadelphia Communication Facilities, E. P. Grimm, City of Philadelphia.

Audio

Milwaukee—September 15

Field Trip-Tour of facilities and inspection of the new RCA Video Tape Recorder, L. Wittenberg, WISN-WISN-TV.

Syracuse—September 15

"Stereo System Specifications," R. B. Dome, General Electric Co.

Automatic Control

Long Island—October 20

"Electronic Computers in Control Systems," J. Truxal, Polytechnic Inst., of Brooklyn.

Los Angeles—October 14

"Attitude Control System for Space Probe Launching Vehicles," H. Low, Space Tech. Labs.

"Reaction Wheel Control Systems for Space Vehicles," H. Patapoff and R. W. Froelich, Space Tech. Labs.

Broadcasting

Omaha-Lincoln—Oct. 30 1959

Tour of Transmitter and Studio; Multiplex FM Demonstration, Tour Leader, J. Katz, KQAL-FM.

Circuit Theory

Los Angeles—October 22

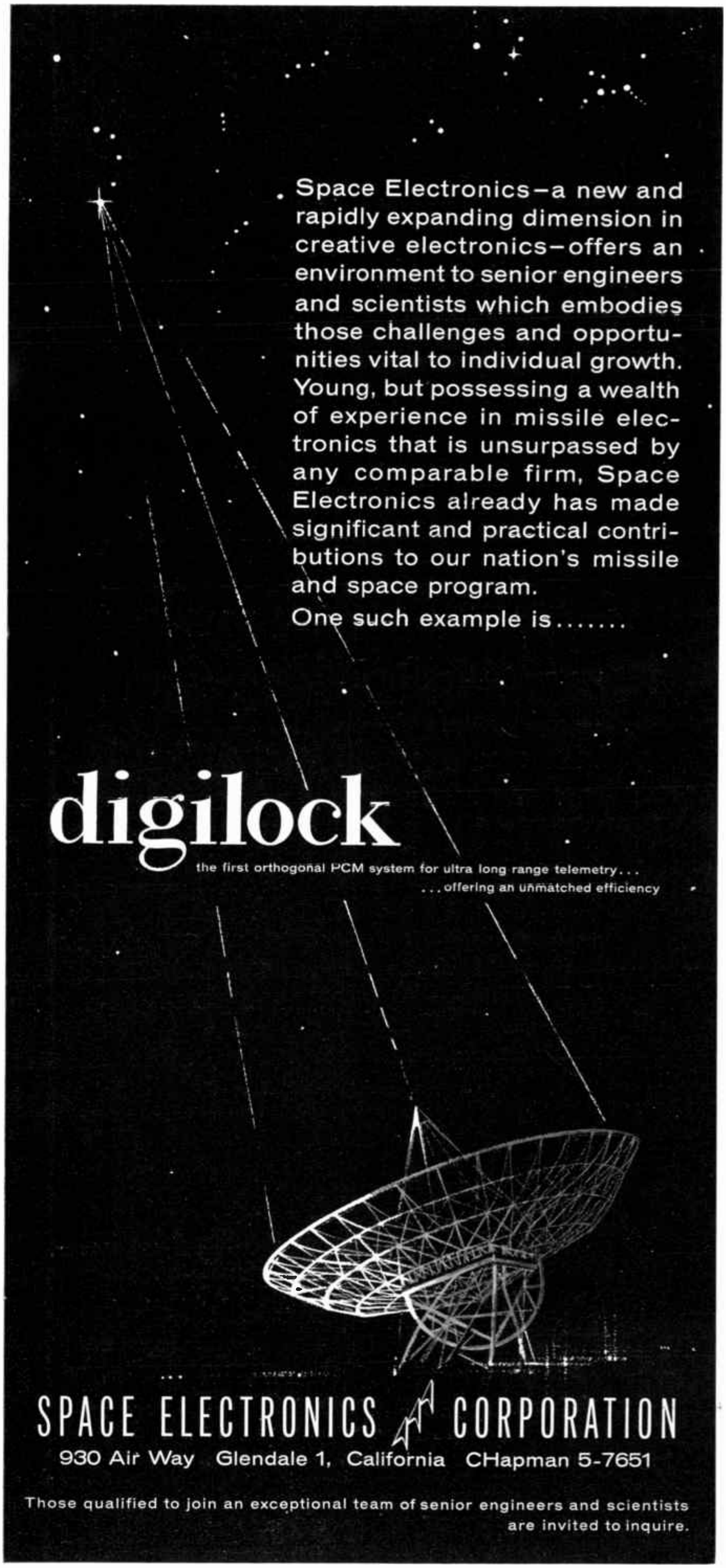
"Active RC Synthesis," I. M. Horowitz, Hughes Aircraft Co.

"Certain Applications of Matrices to Circuit Theory," L. A. Pipes, Univ. of Calif.

Syracuse—October 27

"Active RC Filters," J. M. Sipress, Bell Telephone Labs.

(Continued on page 140A)



Professional Group Meetings

(Continued from page 138A)

Component Parts

Baltimore—October 20

"Tantalum Capacitors," G. Iaggi, Fansteel Metallurgical Corp.

Philadelphia—October 20

"Electrical and Electronic Applications of Glass," G. McLellan, Corning Glass Works.

Electronic Computers

Boston—October 22

"The Design of a New High-speed Computer at the University of Illinois," W. J. Poppelbaum, and J. Robertson, Univ. of Illinois.

Omaha-Lincoln—October 7

Guided Tour through IBM 705 Computer Equipment, Tour Leaders, W. Larsen and R. Vrchlavsky, Customer Engineers, IBM.

Philadelphia—November 3

"New System Philosophy in Computers—The RCA Airborne Digital Computer," A. Baker, RCA.

Electron Devices

San Francisco—October 19

"Integrated Electronics," D. Jenny, David Sarnoff Res. Labs., RCA.

Syracuse—October 21

Organizing Meeting.

Electron Devices/Microwave Theory and Techniques

San Francisco—September 30

"Traveling-wave Masers," A. Siegman, Stanford Elec. Res. Lab.

Engineering Management

Boston—October 15

"Supervisory Pitfalls," L. A. Schmidt, Melpar, Inc.

San Francisco—October 13

"Maintaining A Complex Communication Network," L. Cornell, Pacific Telephone and Telegraph.

Information Theory

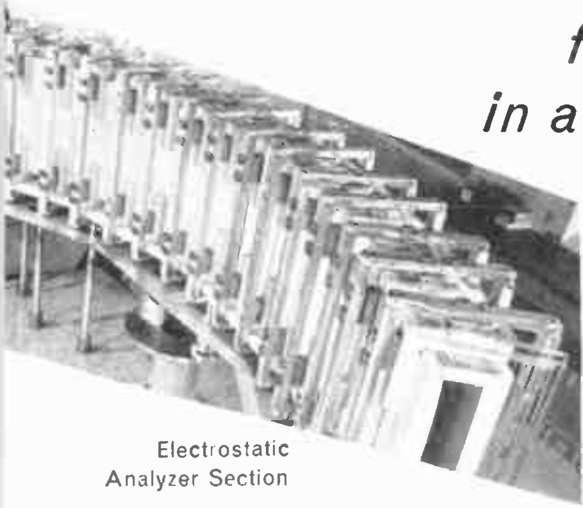
Los Angeles—October 12

"Sequential and Inconsequential Machines," R. Bellman, RAND Corp.

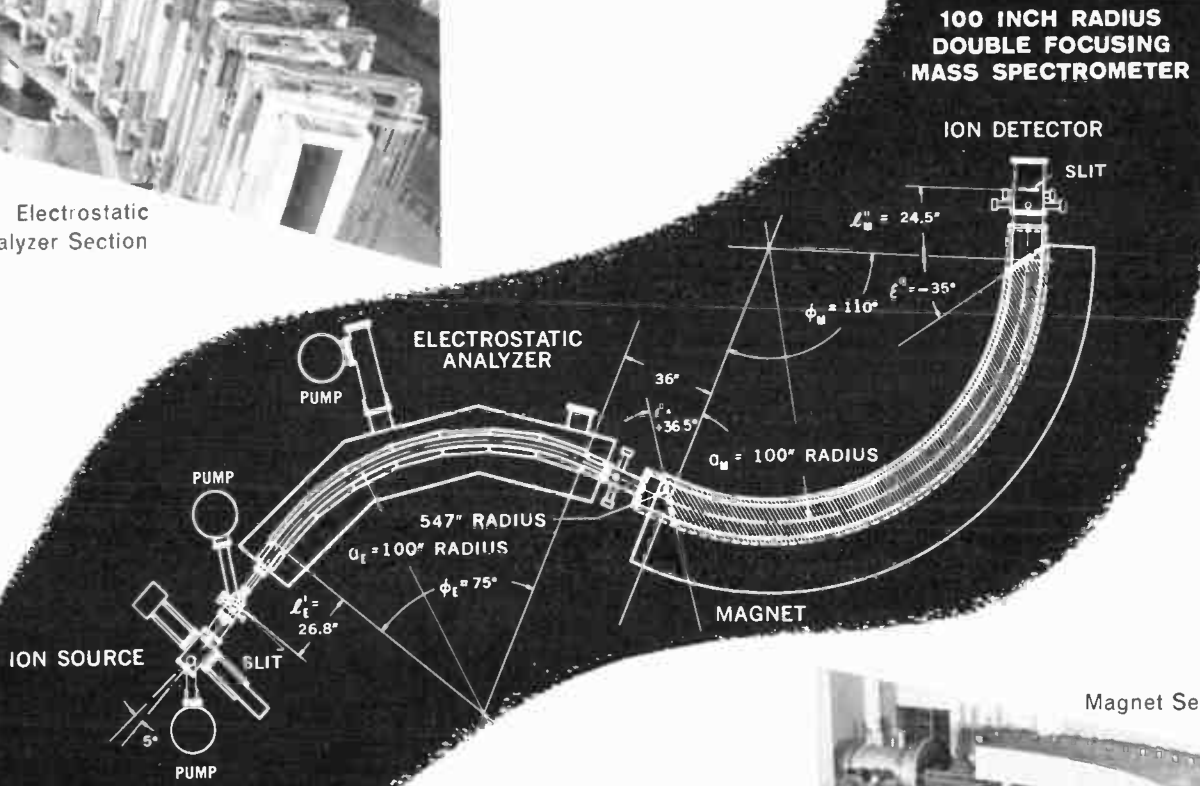
"Some Special Search Problems," O. Gross and S. M. Johnson, RAND Corp.

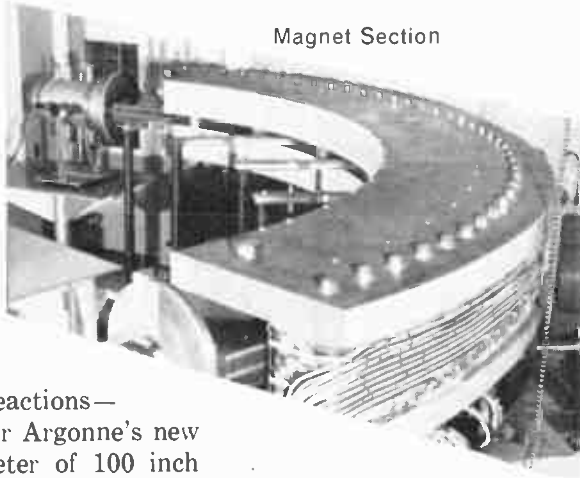
(Continued on page 142A)

finding the needle in a nuclear haystack



Electrostatic Analyzer Section





Magnet Section

The search for a rare heavy isotope, Cm^{247} in nature; determination of spallation products from high-energy nuclear reactions— These are typical analytical problems for Argonne's new double focusing mass spectrometer of 100 inch radius. This huge instrument combines a spherical electrostatic analyzer and a magnetic analyzer in the Mattauch-Herzog arrangement to produce focusing of the ion beam both in direction and in velocity. High transmission and high resolution are achieved simultaneously. The mass spectrometer has been designed to transmit more than 95 per cent of the ions produced in a conventional ionization source with a resolution of 10,000. Specialized equipment and services available at Argonne were required for building this unique research tool. There are unusual opportunities here for those who would create and use such instruments.

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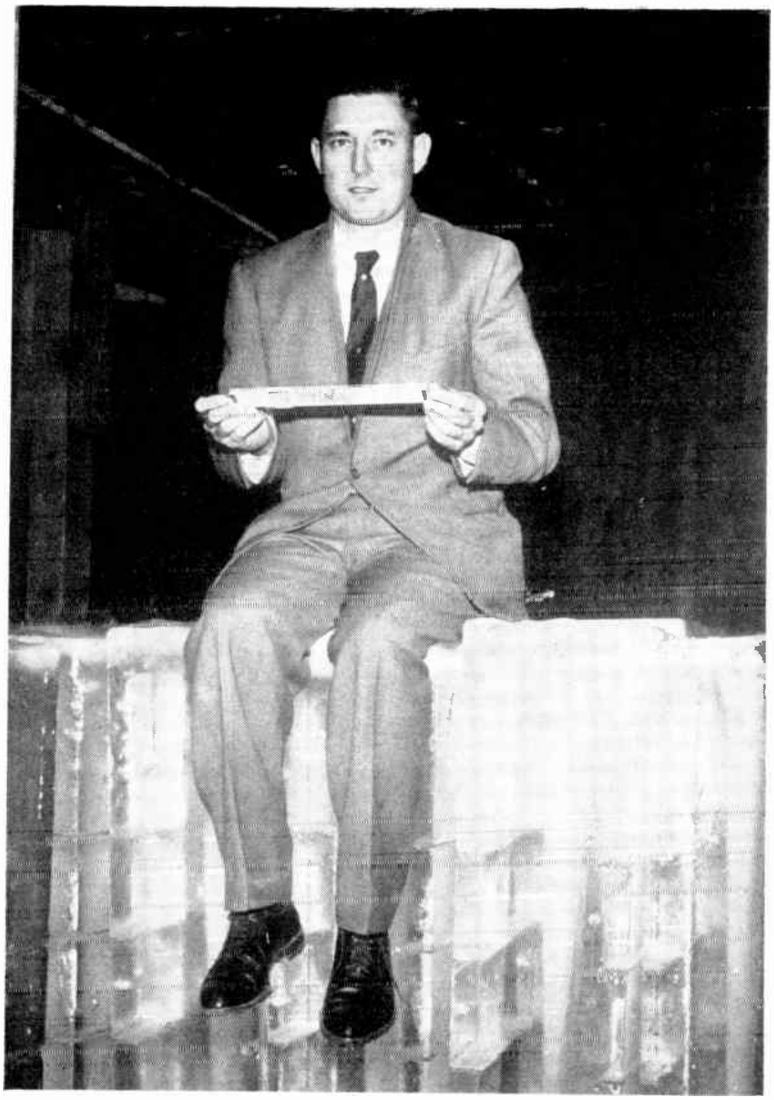


Professional Group Meetings

(Continued from page 140A)

Medical Electronics

Boston—November 4

"Measuring Mental Stability in Space," J. H. Mendelson, Boston Naval Hosp.

Los Angeles—October 15

"Electronics Improves Optical Images for Diagnosis," F. L. Barker, Westinghouse Electric Corp.

"Log-E-Tronic Printer," L. Matlovsky, Los Angeles County Hosp.

Microwave Theory and Techniques

Long Island—September 22

"Some Propagation Phenomena in Ferrite Loaded Waveguide," H. Seidel, Bell Telephone Labs.

Omaha-Lincoln—October 7

"Designing A Microwave System: II," C. O. Jett, Union Pacific Railroad.

Tour of Communications Transceiver Plant, Tour Leader, A. W. Akerson, Union Pacific Railroad.

San Francisco—September 30

"Recent Developments in Masers," A. Siegman, Stanford Univ.

San Francisco—October 20

"The Tunnel Diode," H. Kromer, Varian Assoc.

Military Electronics

Indianapolis—July 9

"Microminiaturization," E. Keonjian, American Bosch Arma Corp.

Long Island—October 27

"Operational Uses of Bright Display Systems," T. K. Vickers, Hazeltine Elec. Corp.

"Hazeltine Large Screen Display for SAGE System," G. Stone, Hazeltine Elec. Corp.

"Scan Conversion Techniques," R. Sorenson, Hazeltine Elec. Corp.

San Francisco—October 6

"Physical and Psychological Simulation of Manned Aircraft Flight," M. D. White and R. M. Barnett, Ames Res. Center.

Nuclear Science

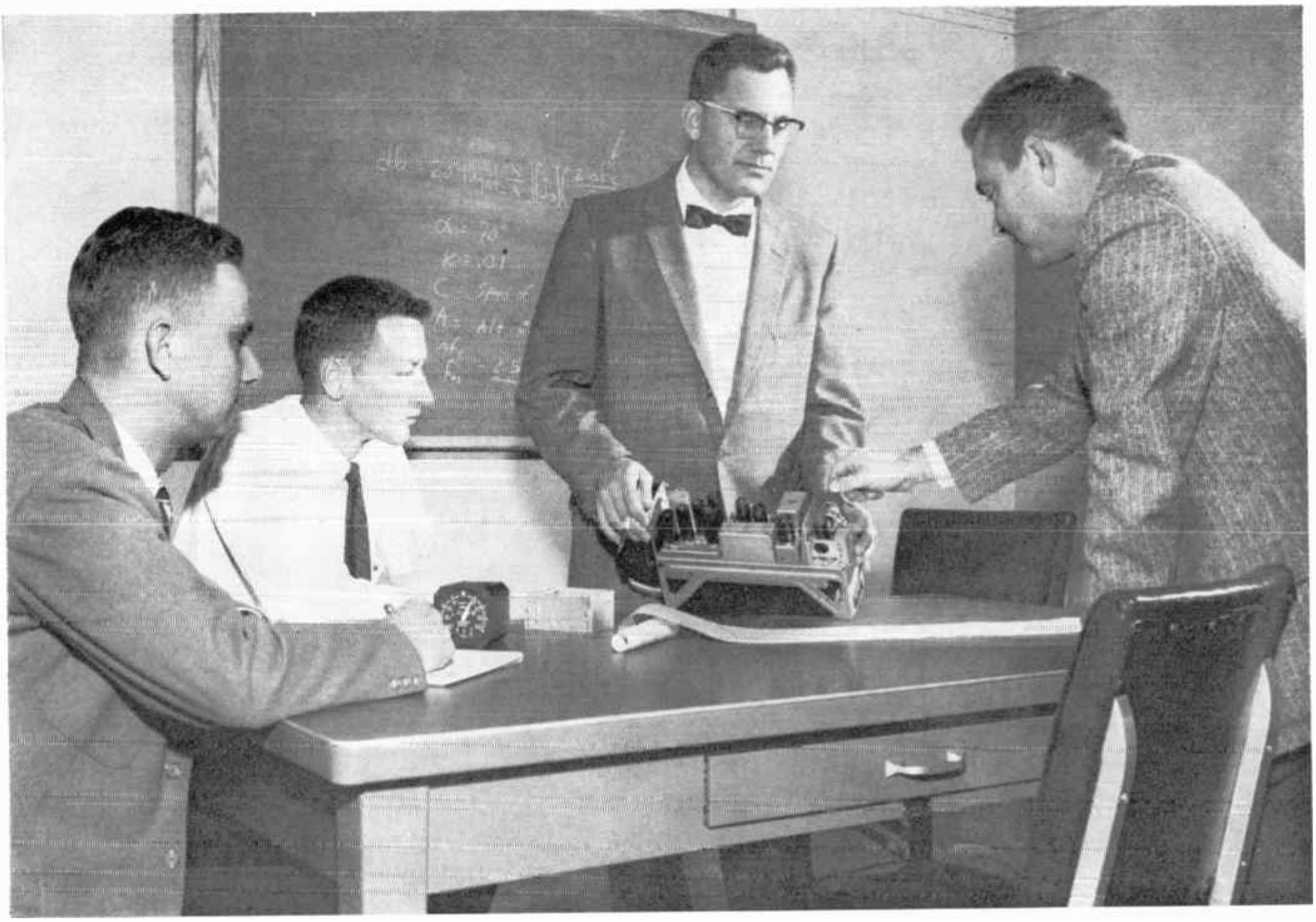
Oak Ridge—September 17

"Status Report of Project Sherwood," E. S. Bettis, Oak Ridge National Lab.

Oak Ridge—October 15

"Very Low Temperature Experiments," J. W. T. Dabbs, Jr., Oak Ridge National Lab.

(Continued on page 144A)



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"The Role of the Mechanical Engineer in Electronic Design," T. A. Smith, RCA.

Reliability and Quality Control

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"Maintainability and Reliability Relationship in an Air Vehicle Weapon System," R. J. Hering, North American Aviation.

"Design Reviews to Improve System Reliability," H. Houghton, Hughes Aircraft Co.

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Tour of Pacific Telephone Central Office, Host, J. O'Brein.

Philadelphia—September 29

Discussion of program for 1959-60 G. Ashendorf appointed program Chairman.

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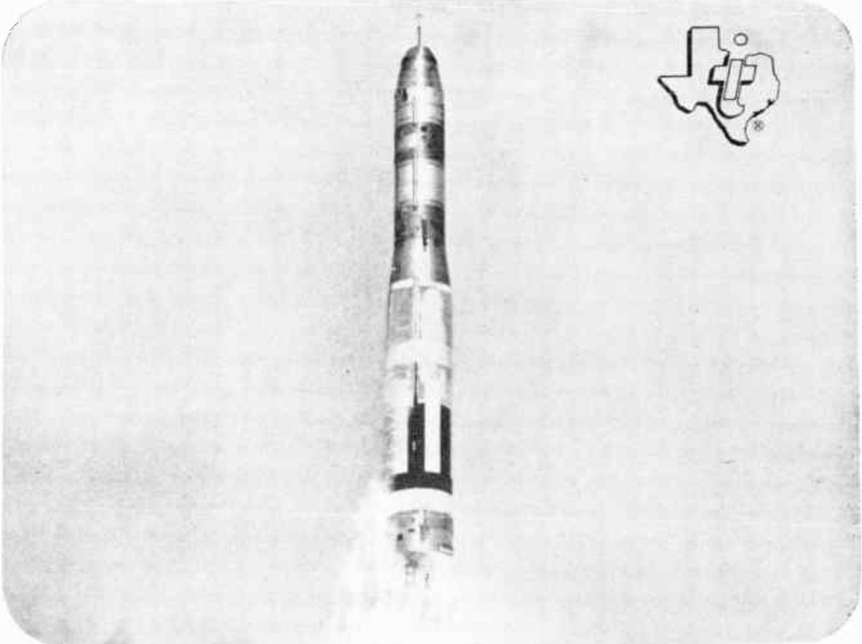

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Cobbold, R. S. C., Saskatoon, Saskatchewan, Canada
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Collins, F. M., Niagara Falls, N. Y.
Colonias, J. S., Portland, Ore.
Cosentino, F. J., Bethpage, L. I., N. Y.
Cronburg, W. E., Winchester, Mass.
Danheiser, E. L., Riverton, N. J.
Danis, P. H., Marlboro, Mass.
DaParma, E. J., Floral Park, L. I., N. Y.
Demkowski, W. P., Jr., Syracuse, N. Y.
DePietro, R. A., East Hartford, Conn.
Devlin, R. P., Glen Burnie Park, Md.
Difeo, V. P., Akron, Ohio
Dimon, G. H., Jr., Rome, N. Y.
Donohoe, F. J., Kokomo, Ind.
Drenik, J., Haddonfield, N. J.
DuBois, N. L., Thornwood, N. Y.
Eck, H. J., Seattle, Wash.
Eddas, R., Portland, Ore.
Eelma, A., Minneapolis, Minn.
Einfalt, J. R., Chelmsford, Mass.
Emery, W. K., Santa Monica, Calif.
Evans, R. T., Melbourne, Fla.
Facchine, F. B., Springfield, Va.
Faris, D. W., Atlanta, Ga.
Fineman, T. P., West Orange, N. J.
Foley, W. M., Wilkes-Barre, Pa.
Forsee, G. B., Indianapolis, Ind.
Frank, C. A., Yardley, Pa.
Gally, S. K., Pasadena, Calif.
Galouchko, A., Rimouski, Que., Canada
Geldmacher, R. C., Lafayette, Ind.
Gibble, C. P., Saugerties, N. Y.
Goehner, R. H., Bloomfield, N. J.
Gonsalves, L. M., New Bedford, Mass.
Goodwin, J. M., State College, Pa.
Gordon, S. S., No. Syracuse, N. Y.
Green, B. K., Dayton, Ohio
Greenstein, L. J., Chicago, Ill.
Grummt, W. E., Philadelphia, Pa.
Hadzoglov, J., E. Cleveland, Ohio
Halaby, S. A., Corning, N. Y.
Halstenberg, R. V., La Jolla, Calif.
Hamet, L. N., Youngwood, Pa.
Hansen, C. T., Buffalo, N. Y.
Harley, W. N., Jr., Smyrna, Ga.
Harmon, L. W., San Francisco, Calif.
Harrison, W. E., Jr., Baltimore, Md.
Hartmann, D. P., West Allis, Wis.
Hickey, P. R., Tujunga, Calif.
Hinchey, F. A., Las Cruces, N. M.
Hoffman, A. N., Plymouth Meeting, Pa.
Holzel, R. W., Clinton, N. Y.
Hoos, W. R., Baltimore, Md.
Howe, J. W., San Francisco, Calif.
Howell, R. S., Poughkeepsie, N. Y.
Hungerford, E. A., Jr., Chappaqua, N. Y.
Hunt, R. W., Kingston, N. Y.
Ioannou, J. T., Dearborn, Mich.
Irving, R., Fairfax, Va.
Irwin, J. M., Syracuse, N. Y.

James, B. G., San Carlos, Calif.
Janas, S., Jr., Elmhurst, Ill.
Johnson, C. E., Los Angeles, Calif.
Johnson, E. G., Woburn, Mass.
Johnson, G. B., San Diego, Calif.
Johnson, H. P., Ontario, Calif.
Johnson, R. R., Somerville, N. J.
Kaufmann, K. G., Pennsauken, N. J.
Kelleher, R. B., Jr., Quincy, Mass.
Kerr-Wilson, G. A., Ottawa, Ont., Canada
King, E. L., San Fernando, Calif.
Kirwin, G. R., Albany, N. Y.
Klees, W. B., Chicago, Ill.
Knolinski, R. E., Kokomo, Ind.
Kreckel, K. H., Rochester, N. Y.
Krone, H. V., Urbana, Ill.
Kuehn, T. J., Latrobe, Pa.
Labby, W. R., Milwaukee 19, Wis.
Lachs, G., Rochester, N. Y.
Lane, E. L., Albuquerque, N. M.
Laurie, K. A., Quebec, P. Q., Canada
Lee, A. N., Sacramento, Calif.
Lerner, S. D., Kew Gardens, N. Y.
Levine, S. L., Pacific Palisades, Calif.
Longworth, R. A., Baldwinsville, N. Y.
Ludwig, D. R., Brookline, Mass.
Lufkin, P. M., Detroit, Mich.
Maggio, A. F., Jr., East Syracuse, N. Y.
Maher, J. E., Jr., Milwaukee, Wis.
Mairson, T., Nashua, N. H.
Maltby, R. J., Kitchener, Ont., Canada
Mann, R. C., Seattle, Wash.
McClead, D. A., Maracaibo, Venezuela
McCoy, J. F., Mineola, L. I., N. Y.
McGee, G. C., Long Beach, Calif.
McGee, R. A., Utica, N. Y.
McKeag, M. F., Wallingford, Conn.
McKenna, M. F., Lancashire, England
Medler, L., Nutley, N. J.
Melan, E. H., Poughkeepsie, N. Y.
Merz, D. M., Schenectady, N. Y.
Michael, R. N., Erie, Pa.
Michels, J. A., Glencoe, Ill.
Montgomery, H. C., Poughkeepsie, N. Y.
Moore, A. L., Takoma Park, Md.
Moore, G. E., Palo Alto, Calif.
Murray, N. H., West Medford, Mass.
Muth, R. M., Syracuse, N. Y.
Myers, W. C., Hillsboro, Ore.
Nachtrieb, T. W., Haddonfield, N. J.
Nelischer, J. E., Belleville, Ont., Canada
Noles, M. H., Houston, Tex.
Ohn, T., Chicago, Ill.
Oldfield, P., Verdun, Que., Canada
Olfman, M., Montreal, Canada
Pagerey, P. F., Rochester, N. Y.
Panas, J. F., Philadelphia, Pa.
Peach, J. A., Rochester, N. Y.
Penkowski, L. J., Garfield Heights, Ohio
Phillip, R. J., Poughkeepsie, N. Y.
Piel, S. A., Fort Worth, Tex.
Polomoscanik, M., Ilion, N. Y.
Ponsness, R. D., Santa Cruz, Calif.
Potter, J. M., Syracuse, N. Y.
Powell, R. J., Sunnyvale, Calif.
Prensky, W. C., Levittown, Pa.
Pruefert, H., Palo Alto, Calif.
Prunty, R. B., Haddonfield, N. J.
Pullman, B. C., Philadelphia, Pa.
Pyle, F. J., Glendale, Calif.
Rahn, R. H., San Salvador, El Salvador, C. A.
Ramsey, R. B., Ferguson, Mo.
Rapoza, T. J., Levittown, N. J.
Ratter, W. F., Mountain View, Calif.
Reichbach, M., Jerusalem, Isreal
Risely, R. E., San Bernardino, Calif.
Robbins, W. G., Clyde, N. Y.
Roberts, B., Kansas City, Mo.
Rosenberger, G. B., Saugerties, N. Y.
Rosenblum, A., Deal, N. J.
Rousseau, A. J., Brockton, Mass.
Rucker, B. T., Endicott, N. Y.
Rudisill, M. E., Burlington, N. C.
Rudolph, L. R., Springfield, Va.
Ruff, G. A., State College, Pa.
Saffery, C. J., Glen Cove, N. Y.
Sampsom, P. M., Upland, Calif

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Membership

(Continued from page 155A)

Sanders, J. G., Lansdale, Pa.
Schiff, G. W., Waltham, Mass.
Schowalter, D. E., Wichita, Kan.
Schowengerdt, R. N., University City, Mo.
Shiber, H. G., West Acton, Mass.
Schultz, W. C., Buffalo, N. Y.
Scully, J. K., Pomona, Calif.
Sharp, D. D., Framingham, Mass.
Shearin, A. M., Greensboro, N. C.
Silverstein, A. N., Springfield, N. J.
Singer, S., Poughkeepsie, N. Y.
Slapsys, A. P., Toronto, Ont., Canada
Slobodin, R. J., Falls Church, Va.
Smolensky, S. M., Andover, Mass.
Snyder, G. L., New Hartford, N. Y.
Speer, S., Brooklyn, N. Y.
Speth, A. J., Yonkers, N. Y.
Squillace, R. E., Tampa, Fla.
Stafford, W. C., Indianapolis, Ind.
Strandrud, H. T., Seattle, Wash.
Strum, R. D., Monterey, Calif.
Szczepkowski, T. G., Philadelphia, Pa.
Tenny, R. M., Philadelphia, Pa.
Tilton, W. D., Denver, Colo.
Tuly, J. F., San Jose, Calif.
Tyzenhouse, E. J., Cocoa Beach, Fla.
Uncapher, M. E., Buffalo, N. Y.
Unruh, W. O., Urbana, Ill.
Van Allen, R. L., Alexandria, Va.
Vorne, A. H., Chicago, Ill.
Wagoner, E. V., Jr., San Pedro, Calif.
Wang, J. C. H., Hyattsville, Md.
Ware, L. E., Jr., Michigan City, Ind.
Waters, L. E., Kingston, N. Y.
Weisman, J., Glen Burnie, Md.
Wellnitz, A. G., St. Paul, Minn.
Wells, C. P., East Lansing, Mich.
Wells, D. R., Long Beach, Calif.
Wells, H. N., Palo Alto, Calif.
Weston, J. E., Johnson City, N. Y.
Wilcox, E. J., Charleston, S. C.
Williams, J. B., Downey, Calif.
Williams, J. F., Inglewood, Calif.
Williamson, W. L., Eau Gallie, Fla.
Wolff, M. F., Hempstead, L. I., N. Y.
Woodward, K. R., Ft. Worth, Tex.
Ziegler, J. M., Massapequa Park, N. Y.
Zimmerman, R. S., Prairie Village, Kan.

Admission to Associate

Allison, J. K., New York, N. Y.
Angle, D. R., Claremont, Calif.
Auburn, D. F., San Diego, Calif.
Austin, T. D., Brownsville, Tex.
Avey, R. J., Jr., Washington, D. C.
Barker, D. R., Holloman AFB, N. M.
Bastien, R. C., St. Albans, L. I., N. Y.
Baumgartner, V. E., Dallas, Tex.
Blasco, J., Montreal, Que., Canada
Blender, M. L., Providence, R. I.
Brien, C. J., Higganum, Conn.
Brumlik, J., Chicago, Ill.
Calvert, D. T., Dallas, Tex.
Cleaver, B. L., Fort Worth, Tex.
Coughlin, L. I., Aurora, Ill.
Crawford, G. H., Schefferville, Que., Canada
De Martini, F., Milano, Italy
Dever, F. J., Philadelphia, Pa.
Edmonds, S. T., St. Albans, L. I., N. Y.
Edwards, J. N., Williamstown, Mass.
Ezzell, K. M., Biloxi, Miss.
Fasnacht, M. B., Columbia, Pa.
Frisiani, A. L., Genoa, Italy
Gee, W., Mountain View, Calif.
Grandi, D. C., Milano, Italy
Grant, M. G., San Francisco, Calif.
Greenstein, M., Chicago, Ill.
Hattori, Y., Philadelphia, Pa.

(Continued on page 158A)

Membership

(Continued from page 156A)

Hitchcock, E. T., Alhambra, Calif.
Irskens, T. E., Tucson, Ariz.
Jacobson, L. A., Port Jefferson Sta., L.I., N.Y.
Johnston, W. R., Jr., Philadelphia, Pa.
Jones, V. E., Jr., Chatsworth, Calif.
Kizer, D. F., Memphis, Tenn.
Kosseff, J. W., Van Nuys, Calif.
Lach, W. J., Hollywood, Calif.
Lachenbruch, D., Radnor, Pa.
Leavitt, W. A., Jr., Waltham, Mass.
Meredith, J. H., Winston Salem, N. C.
Nixon, R. C., Rochester, N. Y.
Nold, J. M., Pomona, Calif.
Northwood, W. E., Manhattan Beach, Calif.
Petrone, L., Torino, Italy
Robinson, F. A., Jr., Twin Falls, Idaho
Sestak, A. T., Milwaukee, Wis.
Schenck, L. S., Mount Kisko, N. Y.
Schmitt, P. T., Baltimore, Md.
Schwartz, A. J., Boulogne-Billancourt, Seine, France
Shaw, C. P., Los Angeles, Calif.
Skybyk, D., Montreal, Que., Canada
Slough, C. M., Richmond, Va.
Smith, C. H., San Pablo, Calif.
Smith, C. M., Triangle, Va.
Steinhoff, N. K., Freeport, Ill.
Stern, J., Pittsburgh, Pa.
Sumner, L. E., Mims, Fla.
Thatcher, R. N., Large, Pa.
Villar, R. A., Lodi, N. J.
Walker, G. A., Akron, Ohio
Wouters, H. M., Utrecht, Netherlands

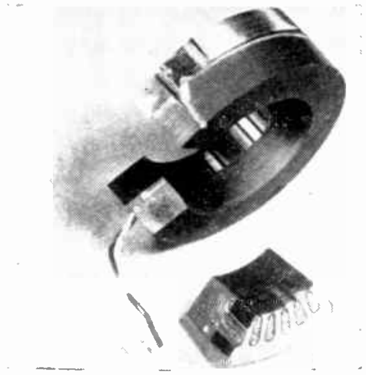
NEWS New Products

(Continued from page 36A)

Epoxy Casting Compound

A new shock resistant epoxy casting system, HYSOL 6622, has been developed by **Houghton Laboratories**, Olean, N. Y.

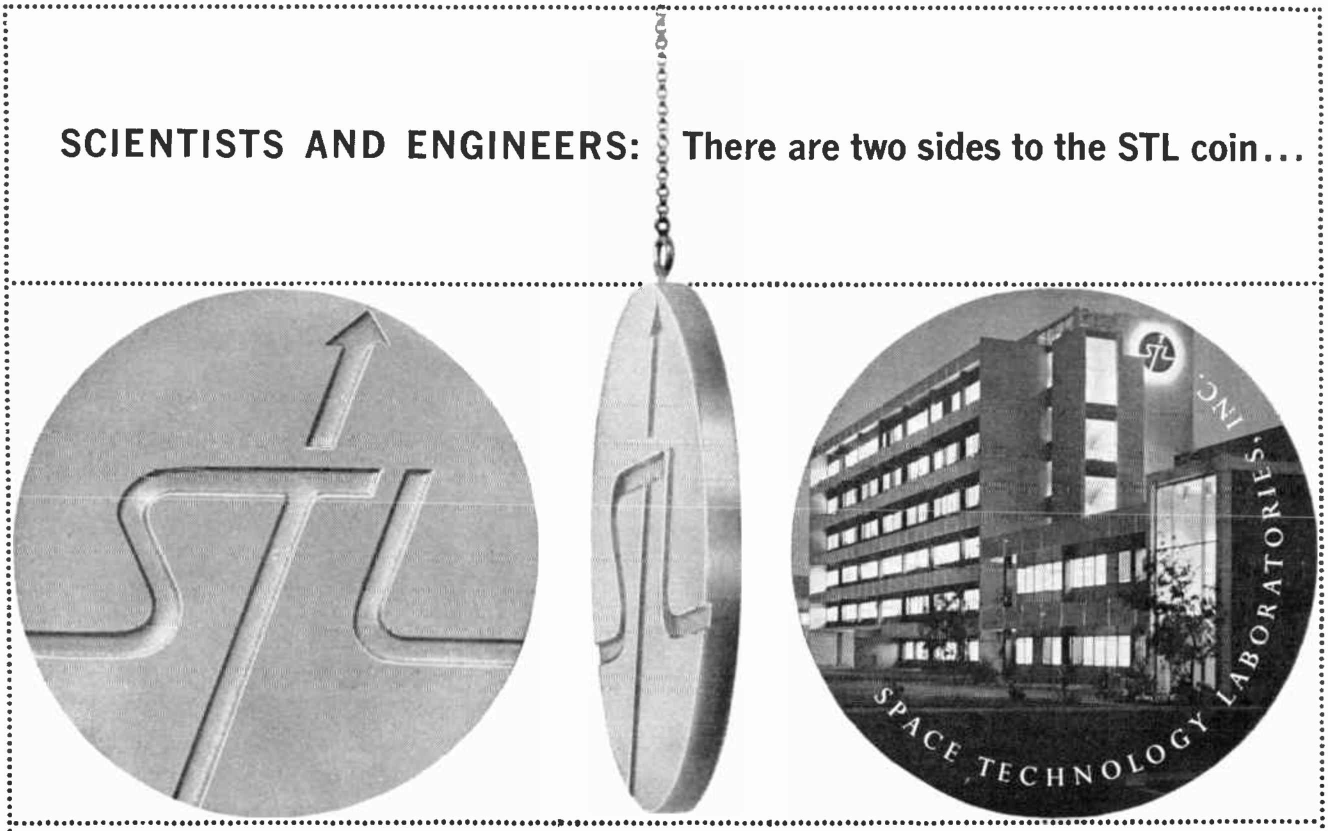
Especially designed for embedding electric motor stators and transformers, it is also suitable for casting or potting large masses where exotherms cannot be tolerated. It is excellent for potting circuits and transformers having strain sensitive elements.



HYSOL 6622-105 (filled) has successfully passed the requirements of MIL-T-27A, offers resistance to thermal and mechanical shock through its flexibility. Castings containing large steel inserts of various configurations have been cycled

(Continued on page 160A)

SCIENTISTS AND ENGINEERS: There are two sides to the STL coin...



What STL does:

Space Technology Laboratories is making significant contributions in theoretical analysis, research, development and technical management of advanced ballistic missile and space systems. STL conducts advanced space flight experiments under the executive management of the Air Force on behalf of such agencies as ARPA and NASA. In addition STL's leadership in military applications of space technology is illustrated by its successful accomplishments as the contractor responsible for over-all systems engineering and technical direction of the Atlas, Titan, Thor, and Minuteman portions of the Air Force Ballistic Missile Program.

What STL offers:

For scientists and engineers with outstanding capabilities, STL offers unusual growth opportunities in many areas of technical activity, including:

Electronic and Electromechanical Systems
Vehicle Engineering and Development
Propulsion and Guidance Systems
Computer Technology
Systems Engineering and Technical Direction
Telecommunications
Airborne Systems
Ground Support Equipment

The Technical Staff of STL is the largest professional group in the nation devoted exclusively to research, development, and systems engineering in the field of ballistic missiles, space projects, and related advanced technology. If you want to apply your skills and talents, in these advanced areas working with leaders in your field, investigate positions at STL now. Please send your resume today to *Mr. Richard A. Holliday.*

Space Technology Laboratories, Inc., P.O. Box 95004
Los Angeles 45, California

SPACE TECHNOLOGY LABORATORIES, INC.

NEWS New Products

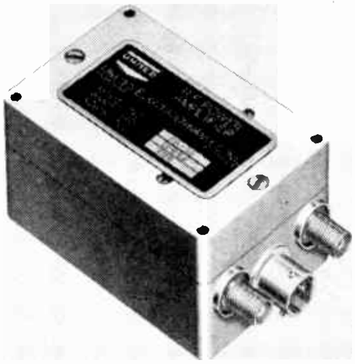
(*Continued from page 158A*)

between −65°C and 150°C without cracking. HYSOL 6622-105 is an easily handled material with a pot life of several days at room temperature and simplified curing characteristics.

For complete details write for the new 6622 product bulletin.

Telemetry Power Amplifier

A super-ruggedized 10-watt amplifier for use in telemetry applications is announced by **United ElectroDynamics**, 200 Allendale Rd., Pasadena, Calif.



(*Continued on page 162A*)

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 160A)

Unit withstands operating temperatures up to 200°F, vibration of 20 g, 20 to 2000 cps, acceleration of 10 g, and shock of 120 g. This hermetically sealed amplifier operates at all altitudes, and exceeds Mil specifications for salt spray and dust.

Unit measures $2'' \times 1\frac{9}{16}'' \times 3''$, and weighs 9 ounces, and is believed to be the smallest and lightest 10-watt power amplifier on the market today.

Designated the PA-10, this amplifier was flight-proven in numerous missile programs, and was also tested in the laboratory. It is used to amplify to 10 watts the signal of any 2-watt transmitter

Literature will be sent on request by the manufacturer.

Ferrite Isolator

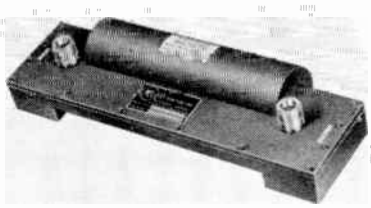
A new S-band Ferrite Isolator spanning a full octave in frequency range has recently been developed using new ferrite and waveguide techniques by **Kearfott Company Inc., Microwave Div.,** Van Nuys, Calif., 14844 Oxnard St., Van Nuys, Calif., manufacturers of precision microwave components and test equipment.

(Continued on page 164A)

NEWS New Products

(Continued from page 162A)

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



The all new Model WD-2106 Octave S-Band Isolator marks a revolutionary advancement in the microwave field by permitting the use of only one isolator for a full octave frequency range instead of covering the band in increments using a number of isolators. Ideal for use in telemetry, radar systems and transponders, the new unit offers exceptional reliability, excellent isolation to insertion loss characteristics and compactness of design.

Outstanding characteristics include a Frequency Range of 2.1 to 4.3 kmc. Isolation is 20 db minimum; insertion loss is indicated at 2 db maximum; other features are input VSWR of 1.5 maximum with Type N connector; with peak power at 1000 watts maximum and average power at 5.0 watts maximum. Temperature ambient is 65°C maximum.

Limit Stop Assembly

PIC Design Corp., 477 Atlantic Ave., East Rockaway, L.I., N. Y., a subsidary of **Benrus Watch Co., Inc.**, has announced the introduction of a compact, heavy duty limit stop assembly.

Available from stock, these "X2" type units are made to MIL-Specifications and are available in both ball bearings and oilless bronze bearings with a wide range of mechanical stop limits from 30 degrees to 4,530 degrees.

The units are self adjustable for fine "zero" adjustment. They are good for stopping torques up to 500 ounce/inches.

Detailed information and prices are contained in the new 416-Page PIC master catalog #20a, available, without charge from the firm.

(Continued on page 166A)

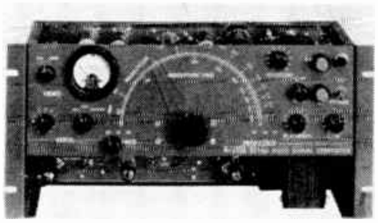
**NEWS
New Products**

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

(Continued from page 164A)

Keyed Video Signal Generator

Foto-Video Laboratories, Inc., 36 Commerce Rd., Cedar Grove, N. J., announces its new V-233A Keyed Composite Video Signal Generator, successor to the V-3 series signal generators.



Like its predecessors, the V-233A features a phase-locked sine wave variable-frequency oscillator for all frequencies between 90 kc and 10 mc. The oscillator phase locks solidly to horizontal blanking pulses, and is followed by a keying amplifier which adds clean blanking and (if desired) sync pulses. The keying amplifier may be used independently to key blanking and sync into external signals such as sweep-generator signals, staircase signals, or saw-tooth generator signals.

The internal oscillator may be used to add its signal to the external signal to form a "modulated" stair-step or saw-tooth waveform. When the phase-locked sine wave is added to a keyed external sweep generator signal a marker is placed on the sweep signal which may be positioned at any point within the range of the instrument.

To permit checking low-frequency response for smear, a window signal generator has been incorporated to add to its flexibility.

The unit may be driven from external triggers such as provided by a standard EIA sync generator. An internal 15.75 kc square wave generator is built-in for field use. A front panel peak-to-peak voltmeter provides set-up and monitoring facilities. A meter-function selector switch selects black-level to peak white, or sync tip-to-peak white measurements. The instrument is complete with electronically-regulated power supply and is available with or without field case. The price is $995.

Leveler Amplifier

A new broadband dc leveler amplifier featuring high gain is now available from **Alfred Electronics, Inc.**, 897 Commercial St., Palo Alto, Calif., manufacturer of broad band microwave instruments and power supplies.

(Continued on page 168A)

BASIC BUILDING BLOCKS FROM KEARFOTT

ANALOG-TO-DIGITAL CONVERTERS

Kearfott's rugged shaft position-to-digital converters are resistant to high shock and vibration and high and low temperature environments. Ideally suited for missile applications, these converters are available for many uses, including latitude, longitude, azimuth or conventional angular shaft displacement conversion and decimal count conversion. Exclusive drum design provides large conversion capacity in smallest size. Combination counter converter assemblies for both visual and electrical readout also available.

TYPICAL CHARACTERISTICS

Kearfott Unit No.	P1241-11A
Code	Cyclic Binary
Range	0-32,768 (2^{15})
Bits per Revolution	16
Revolutions for Total Range	2,048
Volts D.C.	10.5
Current (ma.)	20
Inertia (gm. cm.^{2})	20
Unit Diameter (in.)	$1\frac{7}{16}$
Unit Length (in.)	3
Life	10^6 Revolutions or 10^3 hours
Static Torque (in.-oz.)	2 (break) 1 (running)
Weight (oz.)	5
Maximum Speed (RPM)	600

Write for complete data.

BASIC BUILDING BLOCKS FROM KEARFOTT

20 SECOND SYNCHRO

This synchro, just one of a broad line offered by Kearfott, provides the extreme accuracy required in today's data transmission systems. Kearfott synchro resolvers enable system designers to achieve unusual accuracy without the need for 2-speed servos and elaborate electronics. By proper impedance, matches up to 64 resolver control transformers can also operate from one resolver transmitter.

TYPICAL CHARACTERISTICS **SIZE 25**

Type Resolver	Transmitter	Control Transformer
Part Number	Z5161-001	Z5151-003
Excit. Volts (Max.)	115	90
Frequency (cps)	400	400
Primary Imped.	400/80°	8500/80°
Secondary Imped.	260/80°	14000/80°
Transform. Ratio	.7826	1.278
Max. Error fr. E.Z.	20 seconds	20 seconds
Primary	Rotor	Stator

Write for complete data.

BASIC BUILDING BLOCKS FROM KEARFOTT

INTEGRATING TACHOMETERS

Kearfott integrating tachometers, special types of rate generators, are almost invariably provided integrally coupled to a motor. They feature tachometer generators of high output to-null ratio and are temperature stabilized or compensated for highest accuracy integration and rate computation. Linearity of these compact, lightweight tachometers ranges as low as .01% and is usually better than ± .1%.

TYPICAL CHARACTERISTICS

	Size 11 (R860)
Excitation Voltage (400 cps)	115
Volts at 0 rpm (RMS)	.020
Volts at 1000 rpm (RMS)	2.75
Phase shift at 3600 rpm	0°
Linearity at 0-3600 rpm	.07
Operating Temperature Range	−54° +125°

Write for complete data.

Miniature Floated Gyro

Electrohydraulic Servo Valve

Scanalog 200-Scan Alarm Logging System

Engineers: Kearfott offers challenging opportunities in advanced component and system development.

KEARFOTT DIVISION

GENERAL PRECISION INC.

LITTLE FALLS, NEW JERSEY

Midwest Office: 23 W. Calendar Ave., La Grange, Ill.
South Central Office: 6211 Denton Drive, Dallas, Texas
West Coast Office: 253 N. Vinedo Avenue, Pasadena, Calif.

NEWS New Products

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

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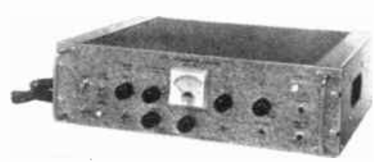


Used with an external crystal detector and directional coupler, the Model 700 can hold RF output from single frequency or swept microwave sources constant to ±0.1 db at a fixed frequency.

Because of the high amplifier gain (6×10^6), variations with frequency are entirely a function of the crystal-coupler characteristics. With proper RF components, power varies less than ±½ db over L, S, C and X bands at slow and fast sweep speeds.

The amplifier is also a broadband attenuator with a 30 db dynamic range over which output power is constant. Model 700 operates with CW signals and internal square wave generator will modulate the RF signal.

Input may be 3 mv to 1 volt. Output is 100 volt maximum in the range of −50 to +100 volts. Frequency response is dc to 100 kc.

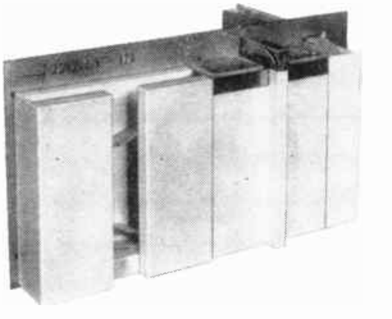
For use with microwave tubes that are operated with the control electrode at a negative potential, a floating otuput with 5 kv insulation is provided. Gain is 5×10^6 and frequency response is dc to 50 kc with output voltage of 450 volts maximum.

For more information, please write to the firm.

Paper Tape Perforator

Paper tapes can be punched at rates to 60 characters per second with a new self-contained tape perforator developed by Tally Register Corp., 5300 14th Ave., N. W., Seattle 7, Wash.

Designed to accept tape of varying widths up to 8 channels, the perforator prepared tape from keyboards, tape reproducers, digital counters and digital data handling systems.



The perforator is available in two models Model 420PF, a panel mounted

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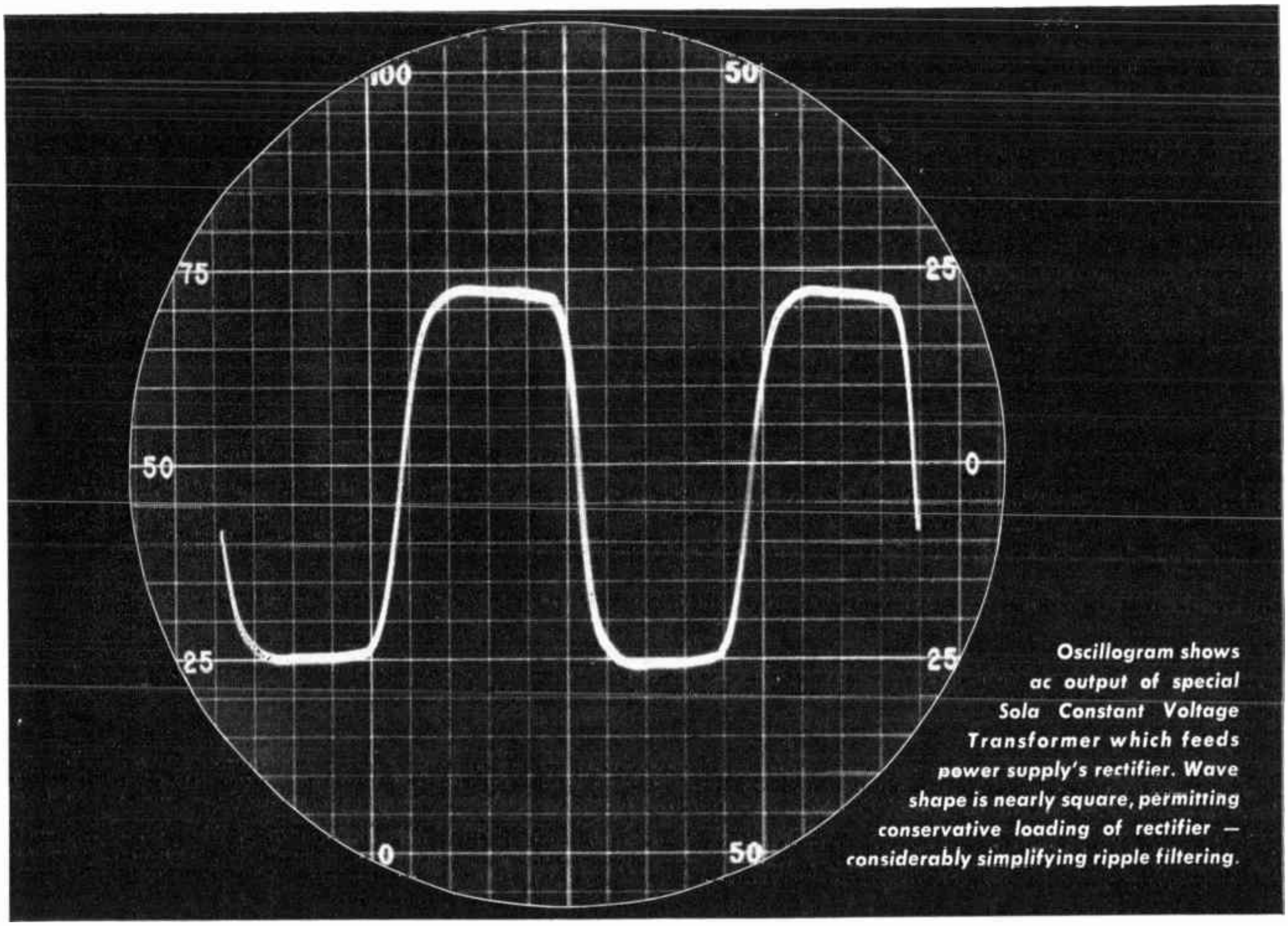


Oscillogram shows ac output of special Sola Constant Voltage Transformer which feeds power supply's rectifier. Wave shape is nearly square, permitting conservative loading of rectifier — considerably simplifying ripple filtering.

Square-wave output of special transformer gives high efficiency in Sola's regulated dc power supply

Sola engineers (men with a keen eye for a trim wave shape) designed a special constant voltage transformer having nearly a square-wave output. Then they linked the transformer with two other components to produce a regulated dc power supply which has notable efficiency.

They fed the regulated output of this transformer into a semiconductor rectifier . . . the low-peak characteristic of the square wave results in a conservative loading on the economical rectifier assembly. It can deliver considerable amounts of current as long as you don't over-voltage it—and over-voltaging just doesn't happen when the input to the rectifier is Sola-regulated to within ±1%.

The rectified voltage feeds into the third component in this happy combination—the high-capacitance filter. The capacitor's filtering job is made easier because the rectified square wave contains a comparatively small amount of ripple. Final dc output from the filter has less than 1% rms ripple . . . for many applications there is no need for a voltage-dropping, efficiency-cutting choke coil.

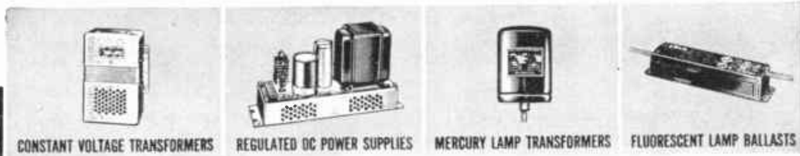
The Sola Constant Voltage DC Power Supply has output in the ampere range, regulates within ±1% even under ±10% line voltage variations, and is suitable for intermittent, variable, and pulse loads. It has low output impedance, is very compact, and provides about all you could ask for in maintenance-free dependability.

Hundreds of ratings of these dc power supplies have been designed and produced to meet widely varying electrical and mechanical requirements of equipment manufacturers. In addition, there are six stock variable-output models and six stock fixed-output models with ratings from 24 volts at six amps to 250 volts at one amp.

For complete data write for Bulletin 1A-DC

SOLA

B P C



A DIVISION OF BASIC PRODUCTS CORPORATION

SOLA ELECTRIC CO., 4633 W. 16th St., Chicago 50, Ill., Bishop 2-1414 • In Canada, Sola Electric (Canada) Ltd., 24 Canmotor Ave., Toronto 20, Ont.

Now meets and exceeds

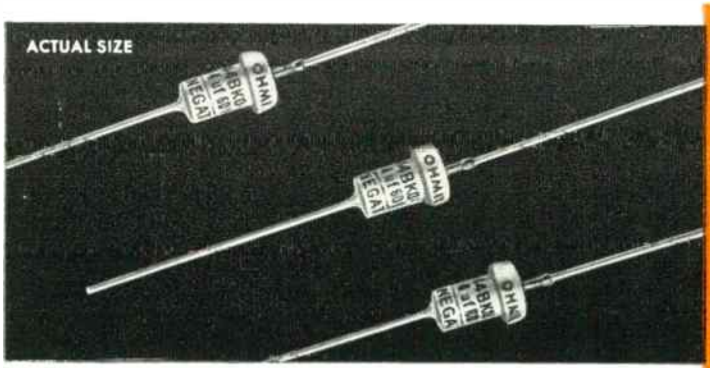
OHMITE®

TAN-O-MITE®

Tantalum Capacitors

Ohmite's supremely equipped laboratory has been approved for official ASESA qualification testing purposes! Here, Ohmite's tantalum electrolytic capacitors are tested on the same type of equipment as used by the military. Furthermore, rigid quality control standards assure 100% testing of every capacitor for its rated parameters—capacitance value, power factor, and leakage current. To meet and surpass the severe military tests, it is apparent that Ohmite capacitors must be more than adequate for any demanding application.

More tantalum capacitor styles are under development at the Ohmite laboratories. Watch for them.



SINTERED SLUG TYPE: MIL-C-3965B, Grade 1, 2, or 3 (Case Size T1, Styles CL44 Uninsulated, CL45 Insulated)

The DC leakage current is less than 0.01 microamperes/mfd-volt at 25°C. The DC surge voltage rating is 115% of the rated DC working voltage.

The anode, a porous slug of sintered tantalum, is sealed into a fine silver case which serves as the cathode and as a container for the wet electrolyte. Axial leads are 1½" long and solderable.

Units are available for operating temperature ranges of −55°C to +85°C and −55°C to +125°C, polar applications only. MIL values available from stock. Other values (which meet MIL requirements) promptly made to order. Write for Bulletin 159

New Slug Type, Straight-Cylindrical Shape

There are no MIL specifications on this type unit at present, but it offers all the characteristics of the slug-type units above with *less* bulk and more convenient mounting. Write for Bulletin 1004

PLAIN FOIL TYPE: MIL-C-3965B (Case Sizes C1, C2, C3; Styles CL34 Uninsulated, CL35 Insulated)

These capacitors now *exceed* the maximum vibration requirements (Grade 3, 5 to 2000 cps) of MIL-C-3965B having been successfully tested at 30 g's, *twice* the required acceleration. They also pass the 50 g shock test of MIL-Std. 202A, Method 205.

DC leakage current is less than 0.035 microamperes/mfd-volt at 25°C; less than 0.20 microamperes/mfd-volt at 85°C (tested in MIL-approved fashion).

The DC surge voltage rating is 116% of the DC rated voltage and the power factor is substantially below the following limits (at 120 cps, 25°C):

Voltage Range	Power Factor
Less than 15-volt rating	15%
15-Volt rating and above	10%

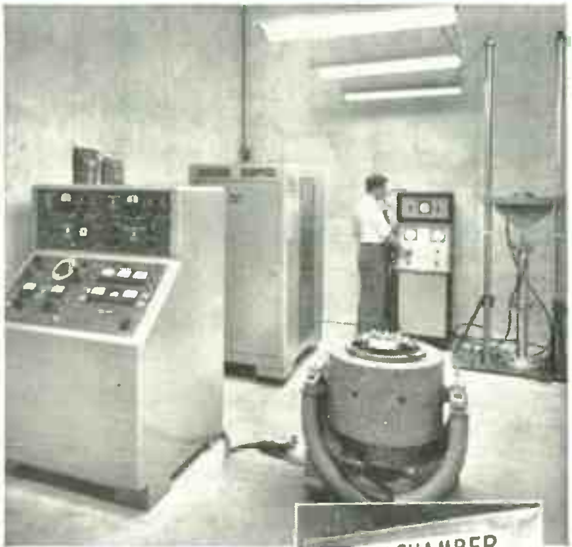
Supplied in polar and nonpolar units although MIL specifications now list only the former. Polar units are protected from current reversals up to 3.75 volts.

Operating temperature range is −55°C to +85°C. In addition to MIL units, many non-MIL values (which meet MIL requirements) are available from .25 to 140 microfarads and up to 150 working volts. MIL values in stock. Other values (which meet MIL requirements) made to order. Write for Bulletin 152

MIL Specifications

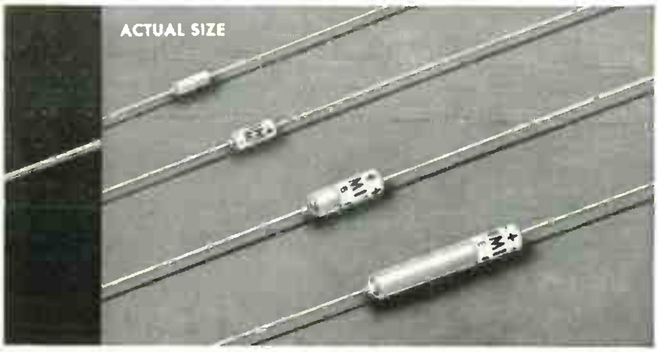
Load life and temperature cycling tests are made in this oven under controlled conditions.

Production measurement of capacitance, power factor, and leakage current.



Vibration, acceleration and shock tests for qualification approval purposes are made in this room.

Moisture resistance tests are conducted in this humidity chamber.



Ohmite also has extensive facilities for making such tests as low temperature exposure, high altitude performance, salt spray corrosion . . . almost everything needed to meet and surpass the most extreme operational and environmental conditions.

WIRE TYPE: No MIL Specification Covers This Type of Unit at Present

These subminiature Ohmite units offer amazingly high capacitance for their small size. Price and size advantages have made them widely used in noncritical, nonresonant low voltage, and transistorized circuitry. Compared to aluminum electrolytics, they offer small size, long shelf life, electrical stability, and superior performance under temperature extremes.

A specially processed tantalum wire serves as the anode. A silver case is the cathode and contains the electrolyte. The negative lead is connected directly to the end of the case. The open end of the case is sealed with a "Teflon" bushing, plus plastic embedment through which the welded anode lead wire projects.

Operating temperature range is −55°C to +85°C. Power factor is generally less than 50%. DC leakage current is less than .09 microamperes/mfd/volt for units of 0.5 mfd or more; less than 0.4 microamperes/mfd/volt for units under 0.5 mfd.

Eleven case sizes satisfy virtually any need. Capacitances from .01 to 80 mfds; voltage ratings to 150. Many stock sizes and values are available.

Write for Bulletin 148

Available from Ohmite Distributors or direct from the factory.

OHMITE®

OHMITE MANUFACTURING COMPANY
3617 Howard Street, Skokie, Illinois

RHEOSTATS RESISTORS RELAYS
TAP SWITCHES R.F. CHOKES
VARIABLE TRANSFORMERS
TANTALUM CAPACITORS
GERMANIUM DIODES

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)

unit for fan fold tape handling, and Model 420PR, using reels. The fan fold model offers easy tape handling and space-saving storage without rewinding.

These perforators use wire clutch drives for each punch. This provides a non-synchronous drive that can be operated at any speed if the minimum time interval between cycles is not less than $16\frac{2}{3}$ milliseconds.

The capstan is driven by a bi-directional tape transport for driving tape backward any distance to correct or modify tape already punched.

For more information, please write to the firm.

Cornell-Dubilier Appoints Craig

Herbert C. Criag has been appointed to the new position of Marketing Manager of Oil-Paper Capacitors, it was announced today by Octave Blake, president of **Cornell-Dubilier Electric Corp.**, South Plainfield, N. J.

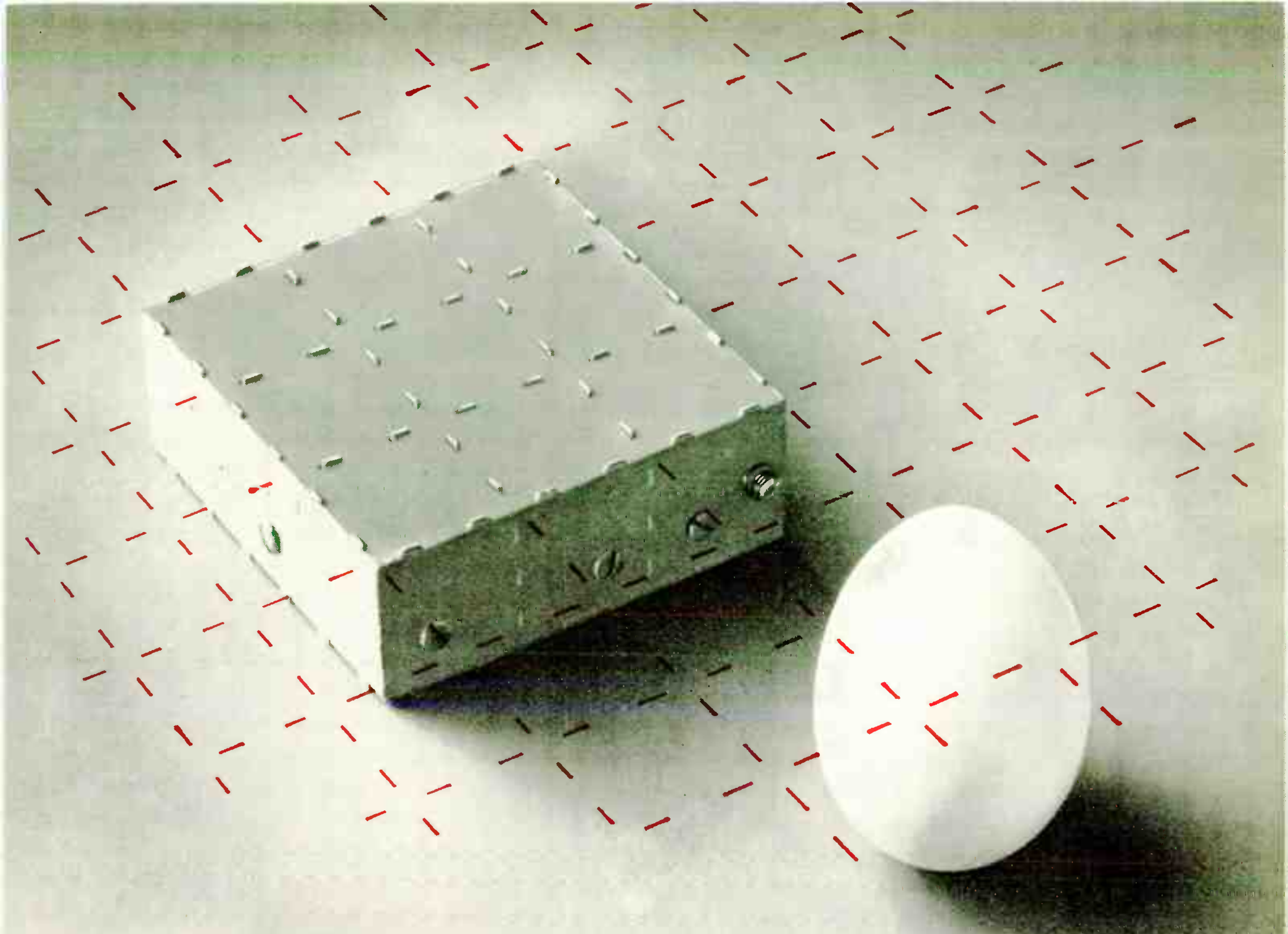
Criag will make his headquarters at the New Bedford, Mass., plant where he will be responsible for product planning, price policy, application engineering and sales promotion relating to all oil-paper capacitors. He will also be responsible for directing similar activities connected with such other items as high-voltage, energy-storage-type capacitors for the recently announced thermonuclear-fusion process for producing energy. This process involves the fusion of a plasma of hydrogen atoms at temperatures of more than one million degrees.

A graduate of Michigan State University, Craig has a B.S. degree in Physical Chemistry, was a research chemist and materials and process engineer for Westinghouse, a design engineer for Line Materials Industries, and a senior design and production engineer for General Electric.

Low Density Resistor Material

A new low density resistor alloy possessing high electrical resistivity—815 ohms/cmf at 20°C.—and a low temperature coefficient of resistance which is inherently controlled within guaranteed limits of ±0.00001 ohm/ohm/°C. (±10 ppm) has been developed by **Hoskins Mfg. Co.**, 4445 Lawton Ave. Detroit 8, Mich. for use in the manufacture of precision wire wound resistors and potentiometers.

(Continued on page 174A)



Better shape factor over wider frequency range

Daven's new EGG CRATE LC filters...

Center frequency: covers the range from 0.4 MC to 60.0 MC depending upon specific requirements. Center frequency stability: ±1.0 KC per MC from -55°C to +105°C. Shape factor: BW_{60}/BW_6 to 2.1. Shape factors can be modified for optimum time delay.

In addition to these outstanding specifications, Daven's new LC filters offer a unique type of construction which makes them the most rugged filters ever built. Small cells are welded together to form the partitioned shield compartment...making this the first filter with truly continuous mechanical and electrical bond...providing a high degree of inter-circuit shielding.

Daven's new LC filters are ideal for shaping the pass band of AM/FM or FM/FM data link receivers, double or single side band receivers and generators, direction finding receivers, communication and telemetering receivers, and spectrum analyzers. So versatile, in fact, that applications are almost limitless. Write today for complete, newly-published technical data.

THE DAVEN CO. LIVINGSTON, NEW JERSEY

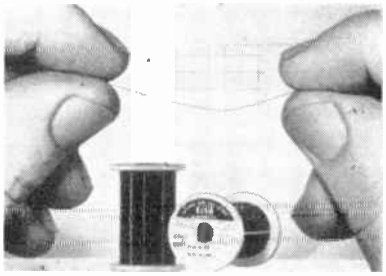
TODAY, MORE THAN EVER, THE DAVEN Ⓓ STANDS FOR DEPENDABILITY

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 172A)

Designated "Alloy 815-R," the material is basically a modified iron-chromium-aluminum composition containing small percentage of several other elements. These plus special melting and processing techniques combine to give it exceptionally uniform physical and electrical properties.



According to laboratory tests—as well as actual field application experience, its resistance to corrosion is said to be superior to other similarly composed materials, equivalent to 80-20 nickel-chromium 800 ohm resistor alloys. Its ductility and tensile and yield strengths also compare favorably with those properties in the nickel-chromium alloys—thus permitting use of the same winding tensions during fabrication operations without increased incidence of wire breakage. Moreover, tests conducted in accordance with MIL-R-19A Specifications—but at twice rated wattage—indicate that the wire possesses high resistance to wear and abrasion. Catalog available on request.

Texas Crystals Opens Florida Plant

The new plant of **Texas Crystals, a Div. of Westronix, Inc.,** has just been placed in full operation at 1000 Crystal Drive, Fort Myers, Fla. Texas Crystals is a processor and supplier of crystals for amateur, commercial and Citizen Band transmitters. The new plant provides a total working space of over 8,000 square feet. The new building provides much needed added facilities for crystal-sawing, processing, warehousing and shipping. Texas Crystals

(Continued on page 176A)

Use your IRE DIRECTORY! It's valuable!

NEWS New Products

(Continued from page 174A)

also announces it has just been awarded an Air Force sub contract to supply nearly 400,000 quartz blanks. Orders continue to be accepted for processing at both the firm's River Grove, Ill., location as well as the new Florida plant.

Guilfoy VP of Elion Instruments

Herbert A. Elion, President of **Elion Instruments, Inc.**, 430 Buckley St., Bristol, Penn., announced the appointment of Edwin J. Guilfoy as Vice President of Sales.

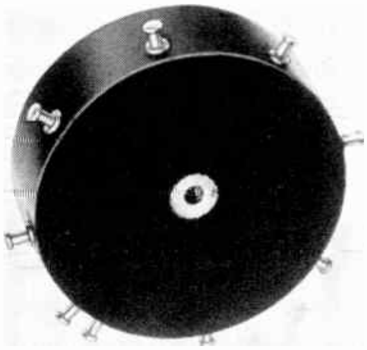
Guilfoy comes to Elion from Radio Corp. of America, where he held the title of Industrial Product Sales and Merchandising Manager. In that capacity he was responsible for sales, merchandising, advertising, budgets and related activities of the marketing of RCA industrial products.

Guilfoy has just completed 14 years with RCA. His experience had been in technical sales and merchandising management. Prior to that time, he was employed by the Abrasive Company as Manager of the Planning Department.

He attended Drexel Institute where he obtained a B.S. in Commerce and M.B.A. in Economics. He taught two years at Rutgers University and during World War II was a lieutenant in the U. S. Navy. He is a member of the Electron Microscope Society.

Toroidal Transformers

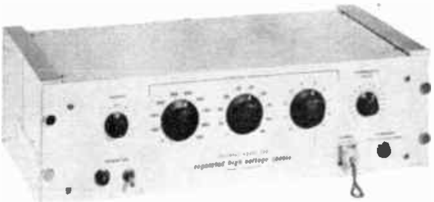
Toroidal transformers suitable for transistorized power supplies are now available in standard models from **Barker & Williamson Inc.**, Bristol, Pa. The transformers are encapsulated to meet all requirements of sealed construction and are designed to perform satisfactorily within a temperature range of −55° to 130°C.



The new line of standard models includes transformers of 25, 60 and 120 watt ratings and a 25 watts inverter which generates 26 and 115 volts 400 cps. The transformers operate on 12 to 14 vdc input. Prices range from $8.10 to $15.25.

(Continued on page 178A)

dial any output



from 0-1000 volts

with 1% accuracy

Keithley Regulated High-voltage Supply gives you new speed and accuracy for a wide range of tests. Its many uses include calibration of meters and dc amplifiers, supplying voltages for photo-multiplier tubes and ion chambers, as well as furnishing potentials for high resistance measurements.

Three calibrated dials permit easy selection of the desired output in one volt steps, at up to 10 milliamperes. Polarity is selectable. Other features include:

- **1%** accuracy above 10 volts.
- **Line regulation** 0.02%
- **Load regulation** 0.02%
- **Ripple** less than 3 mv RMS.
- **Stability:** within ± 0.02% per day.
- **Protective relays** disconnect output at 12 milliamperes.
- **Price:** $325.00.

Send for details about the Model 240 Supply.

KEITHLEY INSTRUMENTS, INC.
12415 Euclid Ave., Cleveland 6, Ohio

IMPULSE

A DIGEST OF NEW DEVELOPMENTS IN ELECTRONICS AND AUTOMATION

PUBLISHED BY ROME CABLE DIV. OF ALCOA, ROME, N. Y.
PIONEERS IN INSTRUMENTATION CABLE ENGINEERING

MAN AND THE MOON. Was Lunik's photo of the moon's backside merely the first step in a larger Russian plan? Speculation since the October 4 moon shot has it that a special moon "package" may have been dropped from the Russian moon station—capable of relaying data continually back to the satellite and then back to earth. This possibility has not yet been verified as this column goes to press. Nonetheless, the general implications are evident: the Russians fully intend to explore the planets and may be beginning now. Can they do it? Well . . . moon rocket control, according to the Reds' own opinion, requires 10 times greater precision than that needed to orbit an earth satellite. This seems to indicate a tremendous rate of achievement for the Russians—improving the precision of rocket control by at least 10 times in the last two years!

SMALL MATTER LOOMS BIG. Little things are becoming a bigger and bigger problem to the aerospace industries. The fact is, measurement techniques are not keeping up with the requirements of advance space projects. Aerospace equipment manufacturers lack the means of measuring, for example, 1/1,000,000 of an inch. Several manufacturers have entered a joint program designed to solve the problem and provide the answers that stand in the way of better calibration and improved standards. Any ideas?

HOW TO BE A CABLE EXPERT. Part of being an expert is knowing where to look for knowledgeable advice. When your problem is cable design or selection of cable materials, you can qualify as "expert" by calling on Rome Cable's specialists to solve your problem for you. In addition to solving the electrical problems inherent in the design of instrumentation and telemetering cables, Rome Cable engineers can also help you overcome certain environmental and physical handicaps under which your system must operate. *You* become the expert because you're *backed* by experts! Get acquainted with what Rome has to offer you by sending for the free booklet "Instrumentation Cables." Address IMPULSE, c/o Rome Cable Corporation, Dept. 1210, Rome, N. Y.

HOW MUCH FOR DEFENSE IN 1960? The Khrushchev visit of 1959 may cause reduced spending for defense in 1960. Speculators looked upon his visit to the U.S. early last fall as a turning point in the cold war. If predictions hold true, then the big electronic stock boom of 1959 will be over by 1960. As yet, there has been no indication of what this might mean to the large number of small firms that depend on R&D.

CABLEMAN'S CORNER. The subject of cable testing is an important one. This is the phase of production that determines whether or not the cable you are purchasing is in accordance with your standards and requirements. In the field of electronics and automation, cables are required to suit various stringent electrical, mechanical, and/or chemical environments. Many years of study and testing have gone into the design of test equipment to be used for these critical tests. It is not enough to know that a cable has been tested in a manner that is "essentially" the same as the required standard. Slight variations in equipment design or methods of tests can mean the difference between conformance and non-conformance. Make sure the test data you receive gives a true picture of the performance of your cable. When you need cable, call on a cable specialist. Our number is Rome 3000.

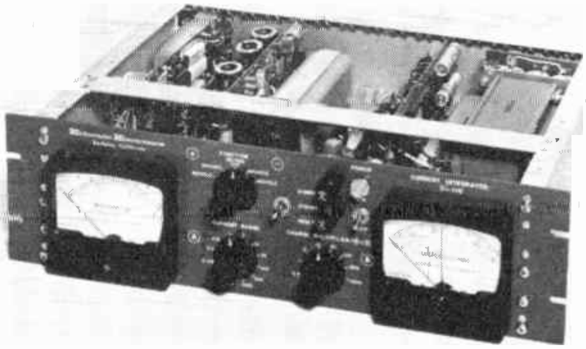
These news items represent a digest of information found in many of the publications and periodicals of the electronics industry or related industries. They appear in brief here for easy and concentrated reading. Further information on each can be found in the original source material. Sources will be forwarded on request.

(11)

(Continued from page 176A)

Current Integrator

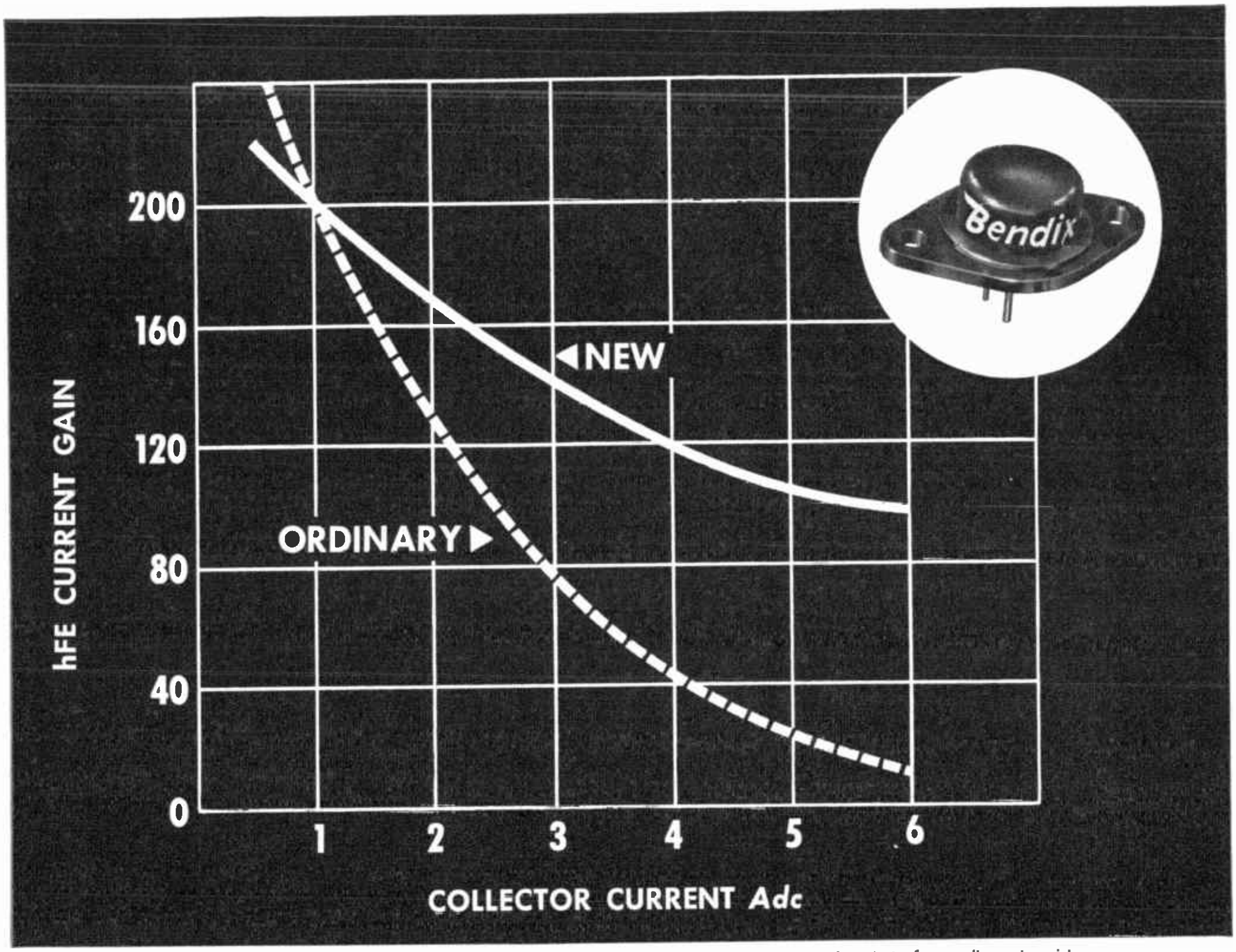
A new Current Integrator, featuring a wide current range and large integrating capacity, is now available from **Eldorado Electronics**, 2821 Tenth St., Berkeley, Calif.



By using the Model CI-110, the physicist or industrial control engineer can simultaneously measure (1) an instantaneous current and (2) the time integral (total charge) of this current. Any physical event which can be represented by an electric current can be measured and totaled with respect to time. Among the possible applications are: controlling exposure in photographic processing on the basis of total radiant energy (to eliminate the effect of brightness changes); accurately deposit a given thickness in electroplating (to eliminate the effect of fluctuations in plating current); and make precise measurements of magnetic flux.

The Model CI-110 measures currents from 0.003 microamperes to one milliampere. Total charges up to 50 ampere-seconds can be measured directly. Higher charges can be recorded with external control.

(Continued on page 180A)



Solid line indicates the low beta fall-off of one of the new Bendix transistors as compared to that of an ordinary transistor.

NEW BENDIX HIGH GAIN INDUSTRIAL POWER TRANSISTORS OFFER FLATTEST BETA CURVE

Now available—a new series of power transistors with the flattest beta curve in the industry, made possible by an exclusive Bendix process. This new series has very high current gains—up to 200 at 3 Adc—and a 10-ampere peak current rating.

Featuring ten-amp performance at a five-amp price, the 2N1136, A, B; 2N1137, A, B; and 2N1138, A, B series provide:

LOW BETA FALL-OFF → LESS DRIVE AND LESS DISTORTION
LOW SATURATION RESISTANCE → GREATER CIRCUIT EFFICIENCY
VOLTAGE BREAKDOWN RATINGS → ELIMINATION OF BURN-OUT
CURRENT GAIN MATCHING → OPTIMUM CIRCUIT PERFORMANCE

Ideally suited for use in static converters and regulators, these power transistors also have numerous applications in relay replacements and drivers for relays, magnetic clutches, solenoids and other loads requiring high current. In addition, their extremely high current gain and excellent hFE linearity make them practical and efficient television vertical output amplifiers and hi-fi amplifiers.

Current Gain hFE at Ic = 3 Adc	Maximum Voltage Rating		
	Vcb 60	Vcb 90	Vcb 100
	Vce 40	Vce 70	Vce 80
50-100	2N1136	2N1136A	2N1136B
75-150	2N1137	2N1137A	2N1137B
100-200	2N1138	2N1138A	2N1138B

For complete information, contact SEMICONDUCTOR PRODUCTS, BENDIX AVIATION CORPORATION, LONG BRANCH, NEW JERSEY, or the nearest sales office.

West Coast Sales Office: 117 E. Providencia Avenue, Burbank, California
Midwest Sales Office: 4104 N. Harlem Avenue, Chicago 34, Illinois
New England Sales Office: 4 Lloyd Road, Tewksbury, Massachusetts
Export Sales Office:
Bendix International Division, 205 E. 42nd Street, New York 17, New York
Canadian Affiliate:
Computing Devices of Canada, Ltd., P. O. Box 508, Ottawa 4, Ontario, Canada

SEMICONDUCTOR PRODUCTS
Red Bank Division
LONG BRANCH, N. J.

Bendix
AVIATION CORPORATION

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 178A)

The Model CI-110 utilizes modular packaging with all basic elements mounted on plug-in circuit cards. Standard instruments can be modified easily to meet unusual measuring requirements.

For more information, please write to the firm.

Solid State Power Packs

Electronic Research Associates, Inc., 67 Factory Place, Cedar Grove, N. J., announces the availability of a new two-color catalog sheet (catalog 116) which describes their new line of Transpac high voltage, miniaturized, solid state power packs.

These new units utilize ERA's Magitran principle which combines the properties of a special magnetic controller with the fast response characteristics and advantages of the transistor regulator. Highly regulated units are available covering the voltage ranges 150–310 vdc at current ratings up to 100 ma.

The catalog sheet provides descriptive material on these units, lists available model types, includes specification data, and current pricing information.

Magnet Material Bulletin

Alnico VII A is the subject of a new engineering bulletin just released by **The Indiana Steel Products Co.**, Valparaiso, Ind. Having the greatest coercive force in the Alnico family, this improved permanent magnet material is especially suited for core type meters and instruments.

The new bulletin lists magnetic and material characteristics of the oriented and non-oriented forms, as well as a demagnetization and energy product curve. Applications for Alnico VII A, which now develops an energy value of 2.8 million (oriented), are also discussed. For copies of the engineering bulletin, write to Indiana Steel. Request Form 351.

Pulse Transformer Catalog

"The Design and Usage of Miniature Pulse Transformers"—new, comprehensive catalogue published by **PCA Electronics, Inc.**, 16799 Schoenborn St., Sepulveda, Calif., covers history of low-level pulse transformers, their chief differences compared to other transformer types, methods of measurement and theory of application. Also included are data on pulse transformer equivalent circuit, transformer polarization, methods of degaussing a core, manufacturing procedures, style and packaging. Some of PCA's 2,000 transformers of standard design are listed together with their circuit diagrams, pulse width charts, electrical specifications and case types.

(Continued on page 182A)

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 180A)

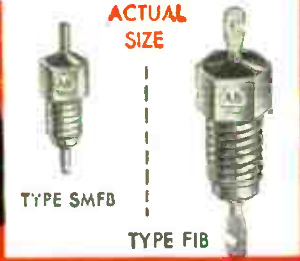
Overload Circuit Breakers

For positive protection of current sensitive components common to transistor circuitry, Series 500-1 electro-magnetic circuit breakers provide tripping action within 25 milliseconds on overloads of 150% of rated current. These miniature, hermetically sealed circuit breakers developed by **Airpax Electronics Incorporated,** Cambridge, Md.—Fort Lauderdale, Fla., are available with current ratings from 50 ma to 10 amperes. Only slightly larger than an on-off switch, they will replace fuses, overload relays and operational switches. Designed for use at dc (50 volts maximum) and ac (120 volts RMS maximum, 60 or 400 cps), these breakers can be supplied in ganged assemblies to save space and to automatically trip two or three circuits when an overload occurs in any one breaker. For further information contact the firm. Literature available on request.

Size 8 Synchro

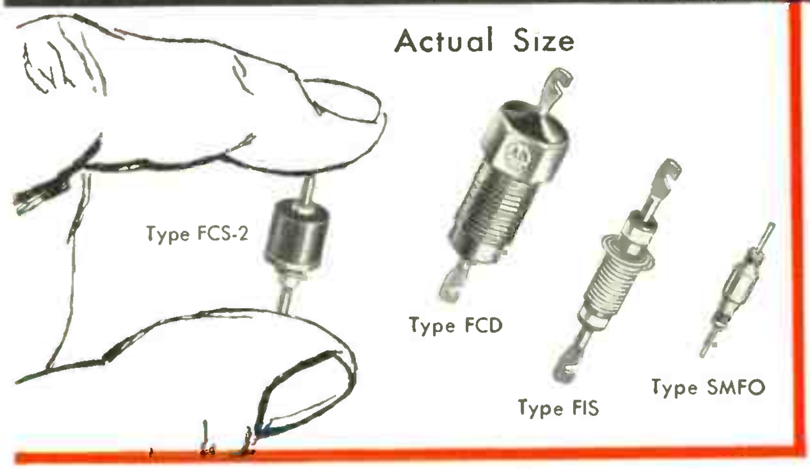
A new 400 cps synchro control transformer with minimum error variations from −55°C to +125°C has been developed by **John Oster Mfg. Co., Avionic Div.,** 1 Main St. Racine, Wis. Type 4227-01 has stainless steel housing. All molded parts are made from special material with high impact strength, excellent dimensional stability, superior heat resistance and good electrical properties. Input voltage is 11.8 volts, input current 0.030 amperes, input watts 0.073 watts, output

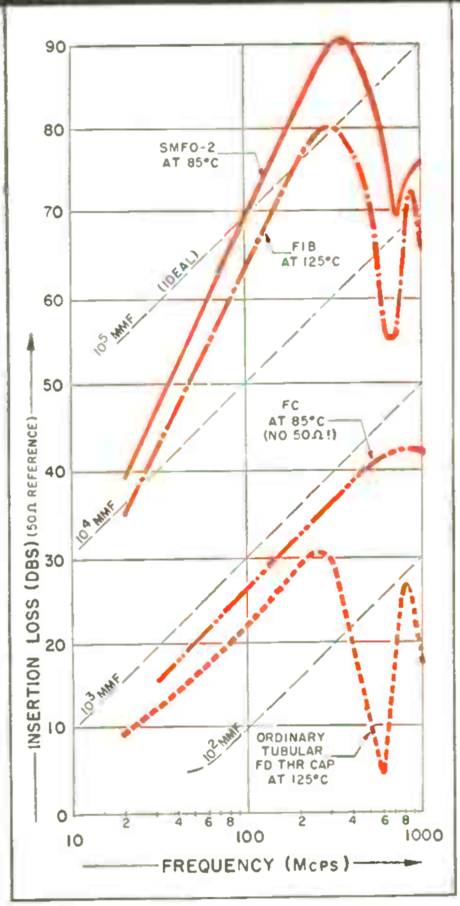
(Continued on page 184A)



Broad Band High Frequency Filters

Allen-Bradley cascaded ceramic feed-thru filters provide effective filtering up to and beyond 5,000 MCS





Here's an entirely new concept in ultra-high frequency filtering—Allen-Bradley's new ceramic feed-thru filters. Their high insertion loss—up to 60 db—effectively prevents feedback and radiation from low power circuits operating in the frequency range from 50 mcs to 5000 mcs.

Astounding in performance, these new A-B filters are actually superior to the theoretical *ideal* capacitor over a wide frequency range. Note, in the graph at right, their effective filtering increases with frequency—and they have none of the undesirable resonance characteristics of standard tubular capacitors. In addition, A-B filter elements provide far greater effective capacitance values than practical with conventional capacitor designs. Filters are available in voltage ratings up to 500 v DC at 125°C. Send for Technical Bulletin 5410.

Allen-Bradley Co., 114 W. Greenfield Ave., Milwaukee 4, Wis.
In Canada: Allen-Bradley Canada Ltd., Galt, Ont.

ALLEN-BRADLEY

Quality ELECTRONIC COMPONENTS

NEWS New Products

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

(Continued from page 182A)

voltage 22.5 volts phase shift 8.5° (lead), rotor resistance 316 ohms, stator resistance 67 ohms, Z_{ro} 500+j1937 ohms, Z_{so} 79+j350 ohms, Z_{rs} 594+j182 ohms, null voltage 30 mv and maximum error from E.Z±7 minutes.

Thorpe GM Of Avnet

The appointment of Harold S. Thorpe as General Manager of the **Avnet Electronics Corp. of Northern Calif.**, was announced by Robert H. Avnet, president of the firm.

In making the announcement, Avnet stated that, "The Avnet Electronics Corp. of Northern Calif. located in Sunnyvale, Calif. is a subsidiary of the Avnet Electronics Corp.

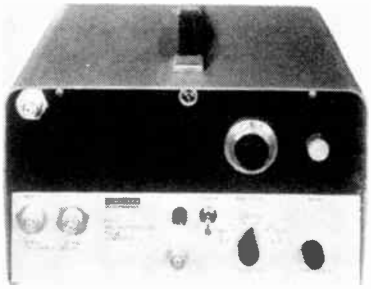
"One of Thorpe's first tasks will be the development and training of personnel necessary to bring the staff of the facility to full strength."

Tuning Fork Frequency Generator

Model TFGX Precision Frequency Generator developed by **Philamon Laboratories, Inc.**, 90 Hopper St., Westbury, N. Y., incorporates the Philamon Tuning Fork Oscillator and additional components to provide frequencies from 60 to 4000 cps and a variety of output waveforms and voltages. Operating from 28 volt dc to furnish 6 volt P/P square wave or pulse into equal or more than 7.5K load or alternately a 1.0 RMS filtered sine wave into equal or more than 10K with less than 1% distortion. Binary Frequency Dividers offer lower frequencies with overall accuracy and stability ranges from 0.001 to 0.05%. Potted in silicone rubber and hermetically sealed for ultimate reliability and maximum life.

Millimicrosecond Pulse Generator

An entirely new transistorized Pulse Generator with risetimes said to be faster than those of any other commercially available instrument has been announced by **Lumatron Electronics, Inc.**, 68 Urban Ave., Westbury, L. I., N. Y.

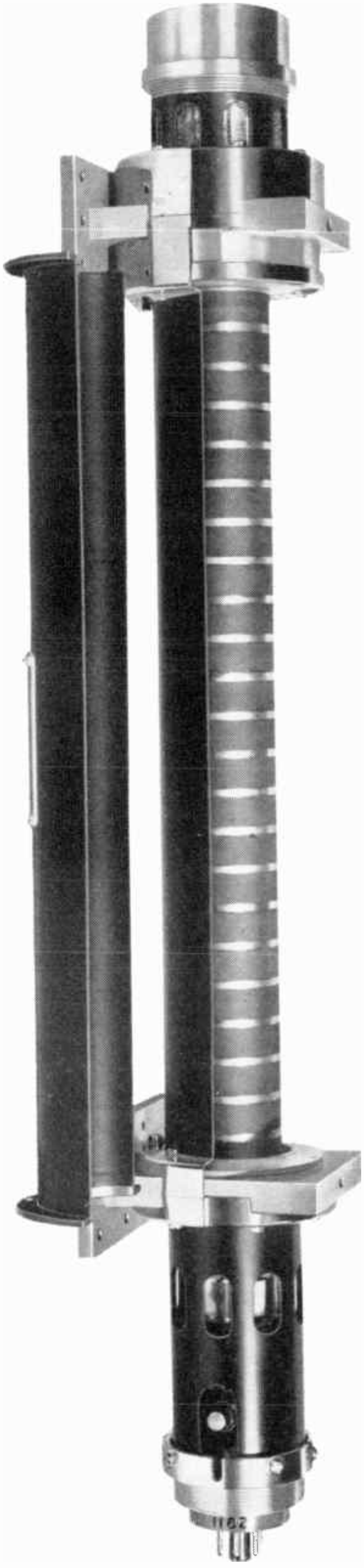


Risetime: Better than .4 millimicroseconds. Repetition Rates: (a) Variable—20 to 300 cycles; (b) Fixed—at 60 cycles, 120 cycles; (c) Externally triggered. Calibrated Pulse Amplitude: 0 to 100 volts, plus or minus. Calibrated Pulse Widths: 1.7, 5, 10 and 20 millimicroseconds (any other pulse widths can be produced by use of added cable). Construction: Mercury switch and cavity assembly replaceable as a unit; Size, 12¾" wide×7" high×10½" deep.

The Model PG-3 is priced at $450. F.O.B. Westbury, New York. The instrument is in production and available for 30 60 day delivery.

(Continued on page 186A)

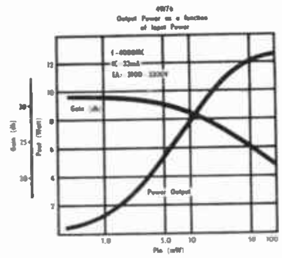
package-type TWT power amplifiers with NEC's new long life cathode

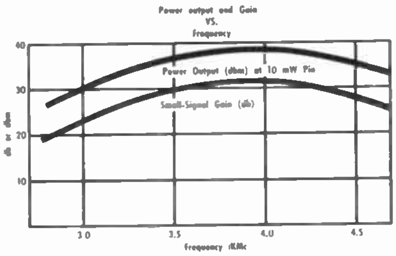


Production of traveling wave tubes at NEC began seven years ago and introduction of the package-type three years later. As chief supplier to Japan's complex network of microwave communications, NEC has become the world's largest maker of TW tubes. With the high development costs amortized and large manufacturing capacity, NEC is now able to supply these tubes at well below usual prices.

NEC's new doped nickel cathode core material, a 10-year development, increases both emission and tube life. It has been thoroughly field-proven in disc-sealed planar triodes for 2000-mc equipment of a large U.S. systems manufacturer (name on request). With its cooler operating temperature, evaporation rate of oxide is less than any other known core materials. This extends tube life up to 50%.

Designers will appreciate the compactness these tubes will give to their systems and operators the reliability and economy. Tubes connect to standard IEC waveguide flanges and can be shipped from stock. For specifications sheets, please write to Tokyo.





4W76

The 4W76 operates in the 4000-mc band and has nominal saturated power output of 10 watts. High amplification over a wide range of power levels results in small-signal gain of approx. 30 db. The band width at half-power points is 1400 mc, but the tube can be used in the frequency range of 2800 to 5000 mc.

Typical Operating Characteristics at 4000 mc

First Anode Voltage	2,640 V	Saturated Power Output	12.5 watts
Helix Voltage	3,220 V	Small-Signal Gain	32 db
Helix Current	0.7 mA	Noise Figure	approx. 25 db
Collector Current	33 mA	VSWR	less than 2 to 1 (from 3500 to 4300 mc)
Focusing Electrode Voltage	-40 V		

NEC TRAVELING-WAVE AMPLIFIERS

PERMANENT MAGNET FOCUSED AMPLIFIERS

4W75	4000-mc band	1.5 watts	8W75	7000-mc band	1.5 watts
4W76	" "	5-10 watts	8W76	" "	5-10 watts
6W50	6000-mc band	5-10 watts	11W17	11000-mc band	1.0 watt

ELECTROMAGNET FOCUSED AMPLIFIERS

4W85	4000-mc band	0.1 watt	4W72A	4000-mc band	1.5 watts
4W86	" "	1.0 watt	7W52	6000-mc band	5-10 watts

Advantages of package type

- NO focusing or impedance matching at installation
- NO dummy space for removal
- NO power source or current stabilizer for electromagnet

NEC Nippon Electric Company Ltd. Tokyo, Japan

COMPONENTS / SYSTEMS

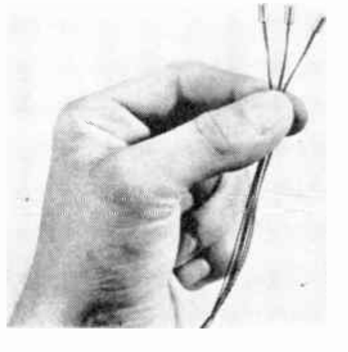
NEWS New Products

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

(Continued from page 184A)

Resistance Thermometer

A resistance Thermometer, half the size of an ordinary pencil eraser and designed for accuracy and high temperature operation, is now available for delivery from **Minco Products, Inc.**, 740 Washington Ave., N., Minneapolis 1, Minn.



The Model S-22 provides reliable operation from $-100°$F. to $+500°$F. It has a resistance of 470 ohms at 32°F., which varies at the rate of about 1 ohm per degree F. The small mass and size .156" diameter × 0.281" long results in rapid response to transients and changes in temperature. The platinum sensing element is potted for maximum environmental capabilities, dielectric and mechanical strength. The attachment method of the lead wires provides a minimum of 5 lbs. pull strength. The stainless steel case will withstand a minimum of 5 lbs. compressive force from a rigid load.

Calibration accuracies of $\pm\frac{1}{4}$%, $\frac{1}{2}$% and 1% are available from stock. Curves, points, and/or equations are available with each unit. Further information and prices are available from the firm.

Power Supply

The Daven Co., 528 Mt. Pleasant Ave., Livingston, N. J., announces a new voltage regulated power supply, Spec 7197. Its input voltage is 115 vac ± 10%, 60 cps ± 1 cps; output voltage is adjustable from 2.5 to 13.0 vdc; load current is 0 to 10.0 amperes continuous duty.

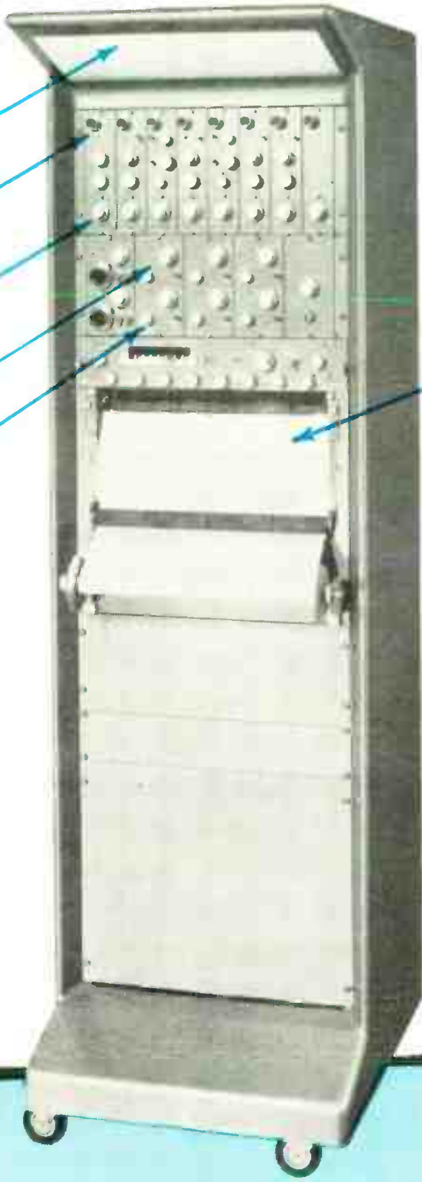
Dynamic regulation: with a ± 90% load change and a step function of ± 5% line change; slow regulation: no load to

(Continued on page 190A)

OFFNER ALL TRANSISTOR

TYPE R DYNOGRAPH

the most versatile . . . most sensitive direct writing unit available



Illuminated canopy

Type 9800 series input couplers provide all input, control and balance functions. Input available both front and rear.

Type 481 Preamplifier provides sensitivities from one microvolt to 5 volts per mm.

Type 482 power amplifiers—may be used without preamplifiers for up to 10 mv/cm sensitivity

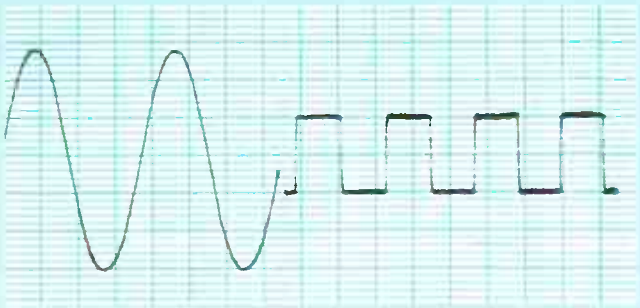
Zero suppression control

504-A paper drive—speeds from 1 to 250 mm/sec. Electrical speed shift 1 to 250 mm per minute available. Zero weave high precision drive, 850 ft. capacity (heat or electric) 1500 ft. (ink). Front loading, with full unobstructed record visible from front.

Combining all these features . . .

- *stable d-c sensitivity of one microvolt per mm*
- *true differential input*
- *high input impedance*
- *response to beyond 150 cps.*
- *reluctance, differential transformer, strain gage with a-c or d-c excitation, thermocouples, etc., used with all preamplifiers*
- *deflection time less than 1.5 milliseconds (2.5 ms with preamplifiers)*
- *fixed precision calibration*
- *instant warm-up*
- *precision source for d-c and 400 cycle excitation, self-contained*
- *zero suppression, twenty times full scale, both directions*

FULL SCALE, UNRETOUCHED CHARTS PRODUCED ON THE TYPE R DYNOGRAPH



EXTREME SENSITIVITY

10 Microvolt RMS Sine Wave — **Ten Microvolt D-C Square Wave**

Four recording media. Heat or electric rectilinear—ink or electric curvilinear. Readily convertible.

All these features . . . plus 8 channels in only 35″ of rack space. Whatever your application for direct writing records . . . you should investigate the ability of the Offner Type R Dynograph to do the job *better* and more *simply*. Using transistor circuits* developed and tested for over three years in thousands of channels of Offner equipment, the Type R Dynograph has already proved its superiority in practically every respect to *any other* direct writing oscillograph. Write on your company letterhead for literature giving details and specifications.

**Patents granted and pending*

OFFNER ELECTRONICS INC.
3912 River Road, Schiller Park, Ill.
(Suburb of Chicago)

1. Within the control blockhouse, ⅞″ 50 ohm Styroflex® cable starts its run to the Titan launch Complex 16 at Cape Canaveral . . .

2. Following along the domed wall of the concrete blockhouse inside a protective case . . .

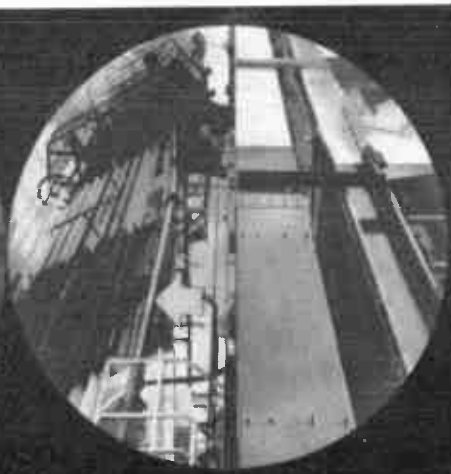
3. To the conduit that carries the high frequency cable through the massive concrete wall.

Styroflex® Coaxial Cable

helps put the USAF Titan ICBM into space

The selection of Styroflex® air dielectric cable for use in the missile field was based on its superior electrical properties, uniformity, rugged physical qualities, long lengths that can be pulled up a tower without splicing and the elimination of radiation always present in braided coaxial cables. ■ Already proven in scores of applications, including broadcast, radar, missile tracking and tropospheric systems, Styroflex® cable has a long record of successes since its introduction in Europe in 1937. ■ Next time you have requirements for a high frequency cable with low attenuation and an extremely low inherent noise level, check the qualifications of Styroflex®. Just write Phelps Dodge.

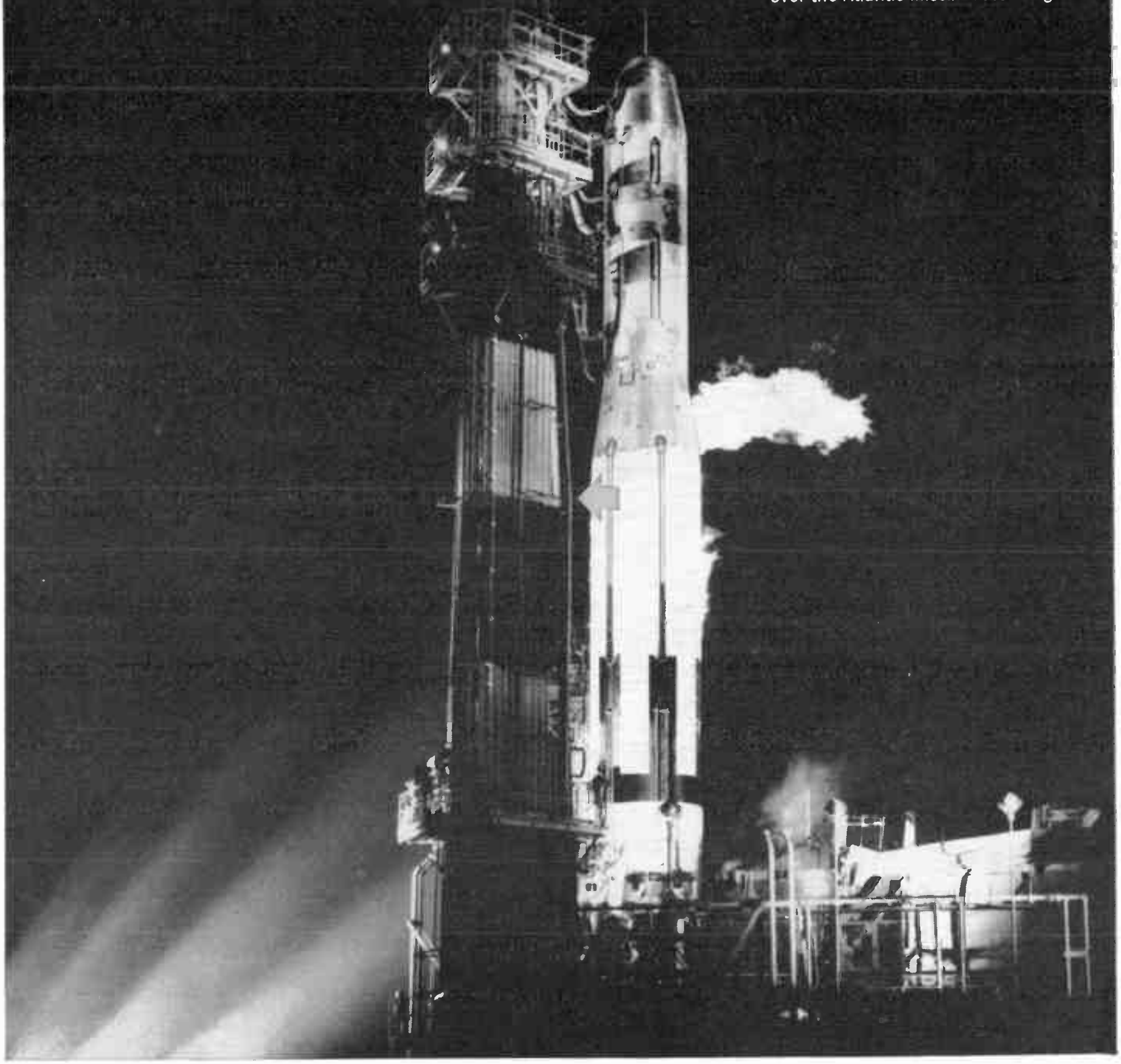
PHELPS DODGE COPPER PRODUCTS
CORPORATION
300 Park Avenue, New York 22, N. Y.



4. Here, the Styroflex® cable from the blockhouse enters the lower deck of the Titan launch Complex 16 . . .

5. Then begins to rise perpendicularly through the lower portion of the launch deck . . .

6. Climbs the side of the umbilical tower and helps send The Martin Company's Titan on a fast trip over the Atlantic Missile Test Range!



NEWS New Products

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

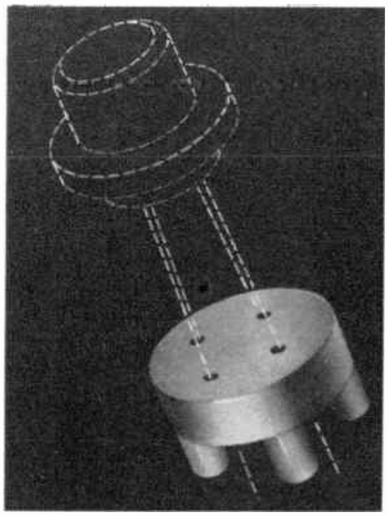
(*Continued from page 186A*)

full load and ±10% line change; total regulation: slow, dynamic and ripple will not exceed ±1%; ripple voltage; will not exceed 10 mv maximum. The power supply is short circuit protected over the entire range and will meet all electrical specifications up to 35°C. The dc output voltage sensing available.

In operation, a Variac, ganged together with a voltage adjust potentiometer, reduces the dissipation on the pass transistors. A magnetic type circuit breaker is set to trip for any overloads in the power supply in excess of 15 amperes. The input transformer supplies ac to the rectifiers and filter with the low voltage high current necessary for the regulated output.

Transistor Mounts

A line of injection-molded plastic transistor mounting pads, designed to eliminate difficulties incurred in mounting transistors on etched circuit boards, is being manufactured by **Regan Plastics Corp.**, 532 W. Windsor Rd., Glendale 1, Calif.

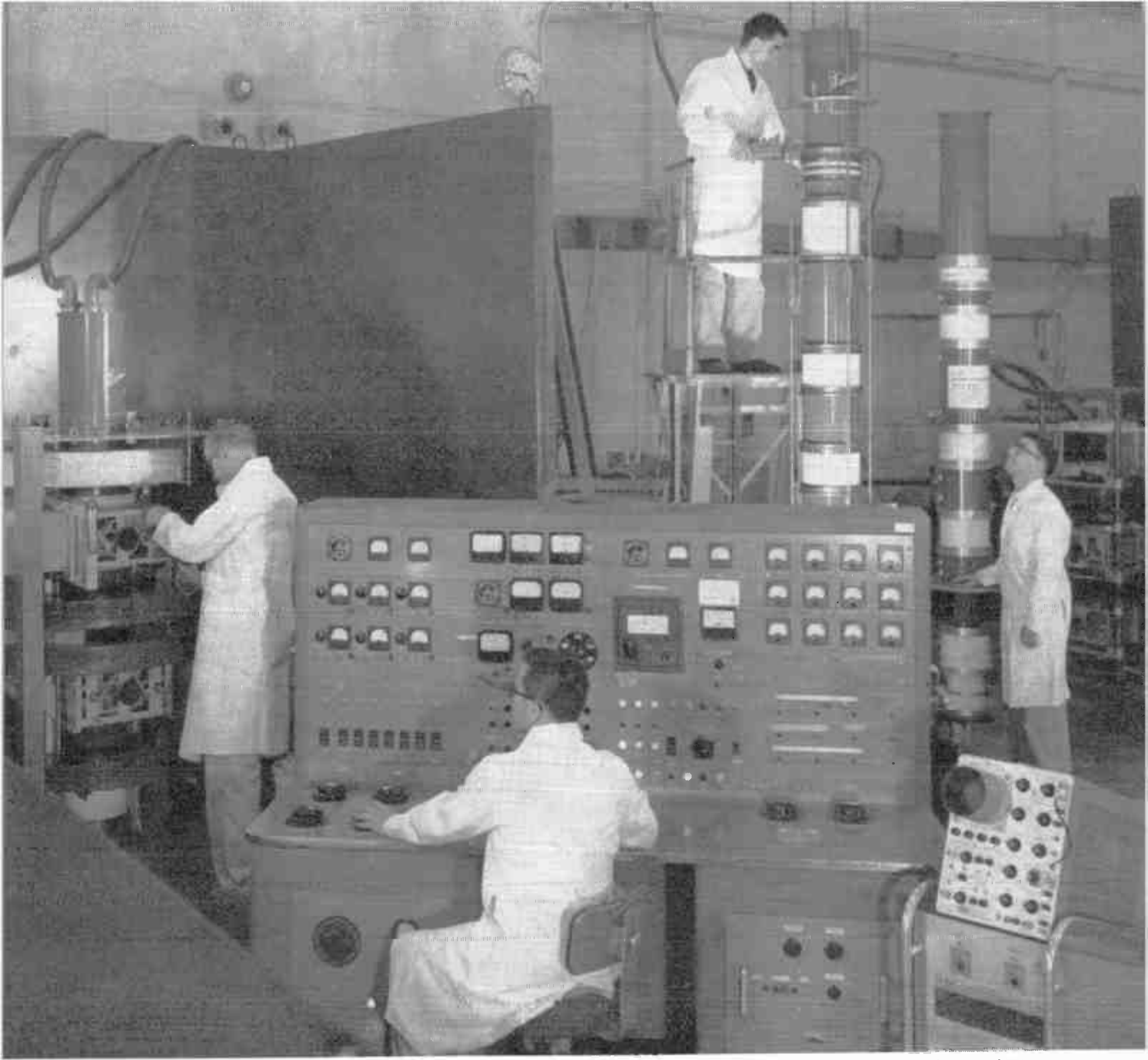


These pads provide stable mounting for all Jetec #12 and #30 transistor cases. They are elevated on four integrally molded feet, minimizing surface irregularities of the board, and allowing an insulating air space. This air space protects the transistor from heat damage during soldering, and prevents the entrapment of moisture at the base of the transistor.

Easy to install, Regan mounting pads are slipped over the transistor leads, which in turn are passed through the board grommets. Leads are then trimmed and formed on the reverse side, ready for hand or dip soldering. Regan pads further simplify soldering, since no jig is necessary for component spacing.

Three designs are currently available in heat resistant nylon, for either commercial or military use. For more detailed information and prices, write to Regan.

(*Continued on page 192A*)



Portion of Eimac's extensive super-power klystron production and test area

MORE EIMAC KLYSTRONS PRODUCED FOR UHF SUPER-POWER RADAR THAN ALL OTHER TUBES COMBINED

A decade ago Eimac decided that negative-grid tubes were impractical to generate high power at UHF. Instead, Eimac developed external cavity klystrons and opened the upper spectrum to high power propagation. With high power at UHF new applications and systems have been made possible.

Custom, laboratory-made tubes can't begin to meet the demands of systems such as UHF space radar. In keeping with its pioneering tradition, Eimac was a leader in developing super-power, long-pulse klystrons for this system—and followed through with quantity production.

This combination of development and production has placed Eimac klystrons in more tropospheric communications and UHF super-power radar transmitters than all other makes of final amplifier tubes combined. And Eimac will continue to convert its developments to production to meet the increasing demand.

For high power at ultra-high frequencies, investigate the many advantages of Eimac external cavity amplifier klystrons.

EITEL-McCULLOUGH, INC.

Eimac San Carlos • California

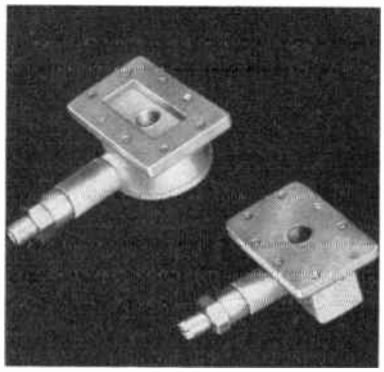
NEWS New Products

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

(Continued from page 190A)

Microwave Cavities

Portchester Instrument Corp., 114 Wilkins Ave., Port Chester, N. Y., offer for immediate delivery a series of cast invar cavities. These cavities, tunable over a 10% frequency range, are available for frequencies from 5929–7750 mc, and can be designed in frequencies from L band through Ku band.

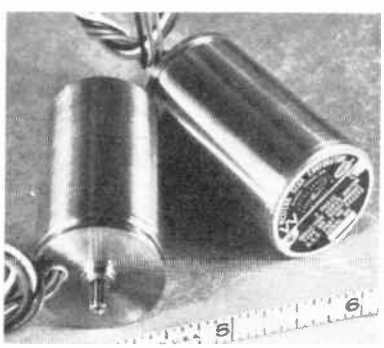


The cavities, when used with a waveguide discriminator, are capable of controlling the frequency of transmitting klystrons to conform to FCC specifications. The technique of invar casting and glass sealing has succeeded in reducing the price of cavities considerably without sacrifice in stability.

Address all inquiries to the firm.

Servo Brake

Western Gear Corp., Electro Products Div., 132 W. Colorado Blvd., Pasadena, Calif., announces the design and manufacture of a new servo motor brake combination. Engineering specifications include: stall torque motor alone 0.48 ounce/inches; brake range available 0.03–0.1 ±0.01 ounce/inches; size 1 inch diameter times 1.85 inches long; designed to meet MIL-E-5272; 115 volts, 400 cps, both phase; nominal impedance 1450 plus J1600; and the brake does not change linearity of the speed torque characteristic.



(Continued on page 194A)

THE PRACTICAL LOW COST ANSWER TO MULTI-CHANNEL OSCILLOGRAPHIC RECORDING PROBLEMS ··· SANBORN INTRODUCES THE FIRST OF THE 950 SERIES ··· THE 958-1500 SYSTEM ··· FOR FLOATING OR GROUNDED INPUTS FROM 10 μV TO 2 MV PER DIVISION ··· ALL-TRANSISTORIZED ELECTRONICS MOUNTED BEHIND A SINGLE 7" HIGH PANEL ··· FLUSH-FRONT RECORDER WITH 9 ELECTRICALLY-CONTROLLED CHART SPEEDS ··· IMPROVED, RUGGED GALVANOMETERS ··· CLEAR, INKLESS TRACES ··· RECTANGULAR COORDINATE RECORDINGS ··· ALL IN A SYSTEM DESIGNED SPECIFICALLY TO PROVIDE GREATER ECONOMY AND ACCURACY WHEN SYSTEM FLEXIBILITY IS NOT

Additional features of the "950" Series include: common power supply, built-in MOPA, front and rear inputs, easily serviced plug-in circuit cards, adaptability for use with other readout devices, availability in 4-, 6- or 8-channel models. When many channels are constantly in use for floating or grounded high gain inputs the simplified 958-1500 design assures dependable operation, yet at much lower "per channel" cost.

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

SPECIFICATIONS—Model 958-1500 System

INPUT	100,000 ohms, all ranges, floating and guarded.
LINEARITY	±0.4%
SENSITIVITY	10, 20, 50, 100, 200, 500, 1000 and 2000 uv per chart div
COMMON MODE REJECTION	100 db, min. dc
FREQUENCY RESPONSE	0-100 cps within 3 db at 10 div peak to peak, 0-50 cps within 3 db at 50 div peak to peak.
NOISE	¼ div peak to peak maximum.

(All data subject to change without notice)

SANBORN COMPANY

INDUSTRIAL DIVISION 175 Wyman Street, Waltham 54, Massachusetts

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

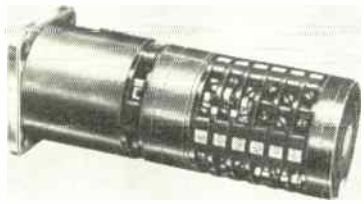
(Continued from page 192A)

The unit has a stainless housing, end bells and shaft, and can be produced with size 10, size 11, or one inch mounting. A design including an integrally mounted gear box with ratios up to 8,000 to one is available. The gear box will add approximately one inch to the overall length of the unit.

For further details write to the firm.

Rotary Solenoid Drive

The American Solenoid Co., 124 Madison Ave., Union, N. J., manufacturers of the "Blue Line"—a complete line of manually operated rotary switches for power and motor circuits, announces their new Rotary Solenoid Drive Unit Attachment.



Switches are assembled by this company with standard stock components which include the basic "stage" unit and one or more of a wide variety of accessories and attachments as required. Designed for use in conventional stepping arrangements or master slave operations, the unit incorporates a feature which allows manual operation in the event of power failure.

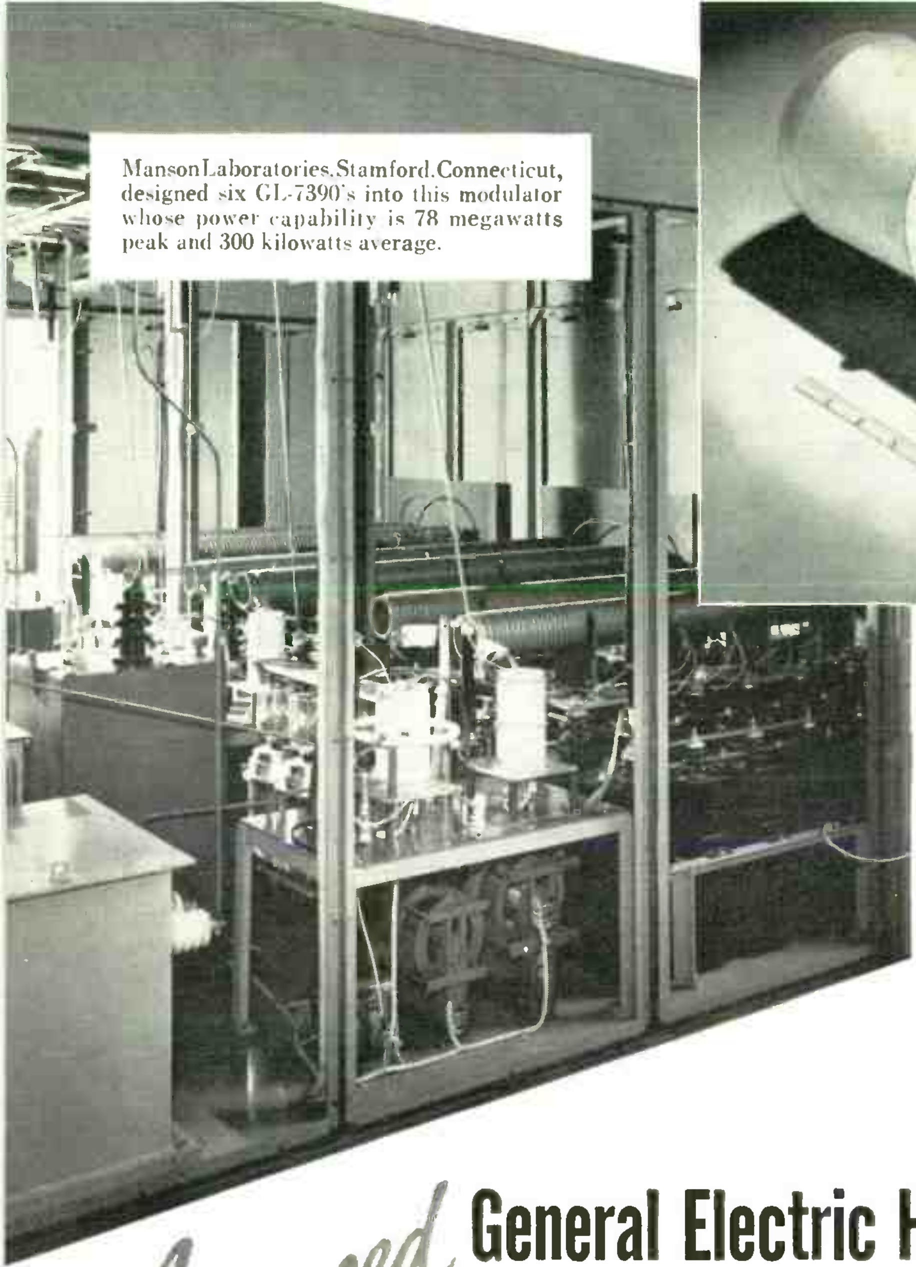
Like all other standard attachments, the solenoid drive unit is quickly available from stock and has the patented bayonet feature. Thus, this drive is attached quickly and economically without extra tools. Amperage ranges is stated at 10 to 200 as standard.

For complete information, write to the firm.

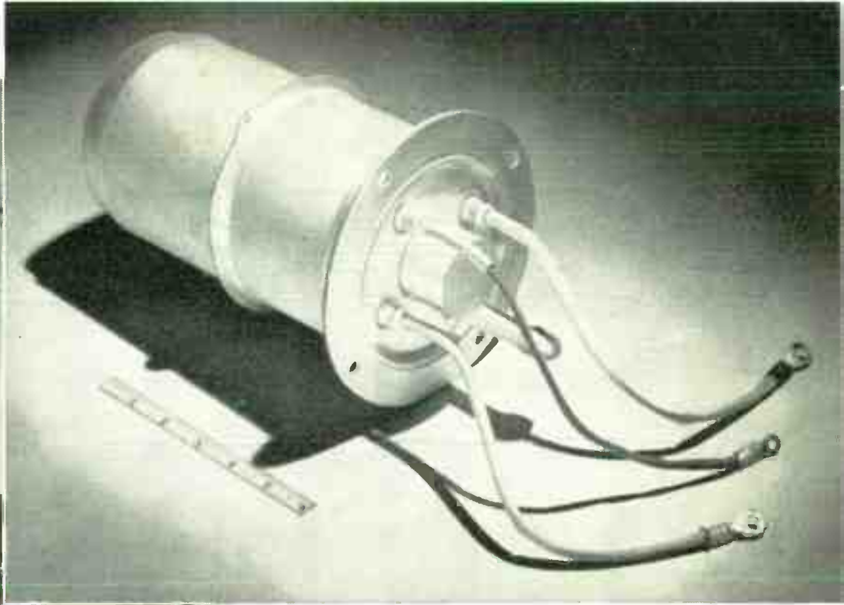
Sweep Generator

Jerrold Electronics Corp., Industrial Products Div., Philadelphia 32, Pa., is producing its new Model 707 high output, flat sweep frequency generator. Claiming a ±0.05 flatness over its highest octave of coverage, the new unit is available with plug-in oscillator heads covering any portion of the spectrum from ½ to 250 mc.

(Continued on page 196A)



ManSon Laboratories, Stamford, Connecticut, designed six GL-7390's into this modulator whose power capability is 78 megawatts peak and 300 kilowatts average.



Below are shown the approximate envelope sizes and power outputs of two thyratrons now in use in high-power radar, as compared to the new General Electric tube.

Type 1257	Type 5948	New G-E Development (GL-7390)
8½"x20"	5"x16"	6"x11"
Avg. Power 33KW Peak Power 33MW	Avg. Power 12.5KW Peak Power 12.5MW	Avg. Power 66KW Peak Power 33MW

CHARACTERISTICS:

Peak Anode Voltage	33 KV
Average Anode Current	4 amperes
Peak Anode Current	2,000 amperes
Anode Dissipation Factor	30 x 10^9

Advanced General Electric Hydrogen Thyratron Available NOW from Stock!

The new General Electric GL-7390 hydrogen thyratron, which has the highest known power handling capability of any hydrogen thyratron now available, can be shipped immediately from stock. Designed for high-power radar pulse modulators, the GL-7390 features metal-ceramic construction for great mechanical ruggedness, smaller size for important space savings, and ability to switch extremely high average and peak power.

The external anode and grid construction allows direct convection cooling of the anode and grid. Reduced anode and grid temperatures during operation minimize the possibility of arc-back and/or grid emission.

Ceramic-metal construction provides a rugged envelope which enables the GL-7390 to withstand shock and vibration conditions beyond the limits of glass designs. The anode and grid are in the form of solid metal cups solidly brazed to the ceramic body. This is a far stronger design than conventional glass seals and lead supports.

The metal-ceramic construction allows close, accurate, and rigidly fixed spacings of the anode and grid. The result is very reliable high-voltage operation. Application assistance available from your regional General Electric power tube office. *Power Tube Department, General Electric Company, Schenectady 5, New York.*

Progress Is Our Most Important Product

GENERAL ELECTRIC

9545-8481-23

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 194A)

The 707 features an all-electronic saturable reactor, permitting a maximum deviation of approximately $4\frac{1}{2}$ to 1 with sweep rates adjustable from one sweep every two minutes up to 60 sweeps per second.

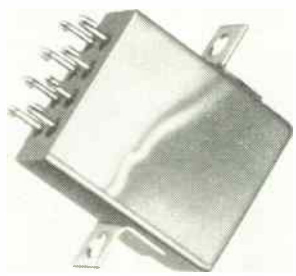
Direct coupled circuits in the 707's monitoring detector and deviation drive circuits permit use of the instrument with "X–Y" recorders where permanent records for inspection or reporting purposes are desired. Sawtooth or pyramid sweep shapes may be selected by adjustment of a front panel control. The output exceeds +20 dbm and is monitored by a front panel meter.

The instrument is designed to operate from a 115 volt 50/60 cps supply built on a 13 by 17-inch chassis for cabinet or 19-inch rack mounting. The 707 will sell for $795.

For additional information, contact the firm.

Latching Relay

An electrical latching model is the latest in a series of miniature relays developed by **Babcock Relays, Inc.,** 1640 Monrovia Ave., Costa Mesa, Calif. Incorporating military specifications, the BR-7A latching relay is intended for application in time delay and sequencing circuitry.



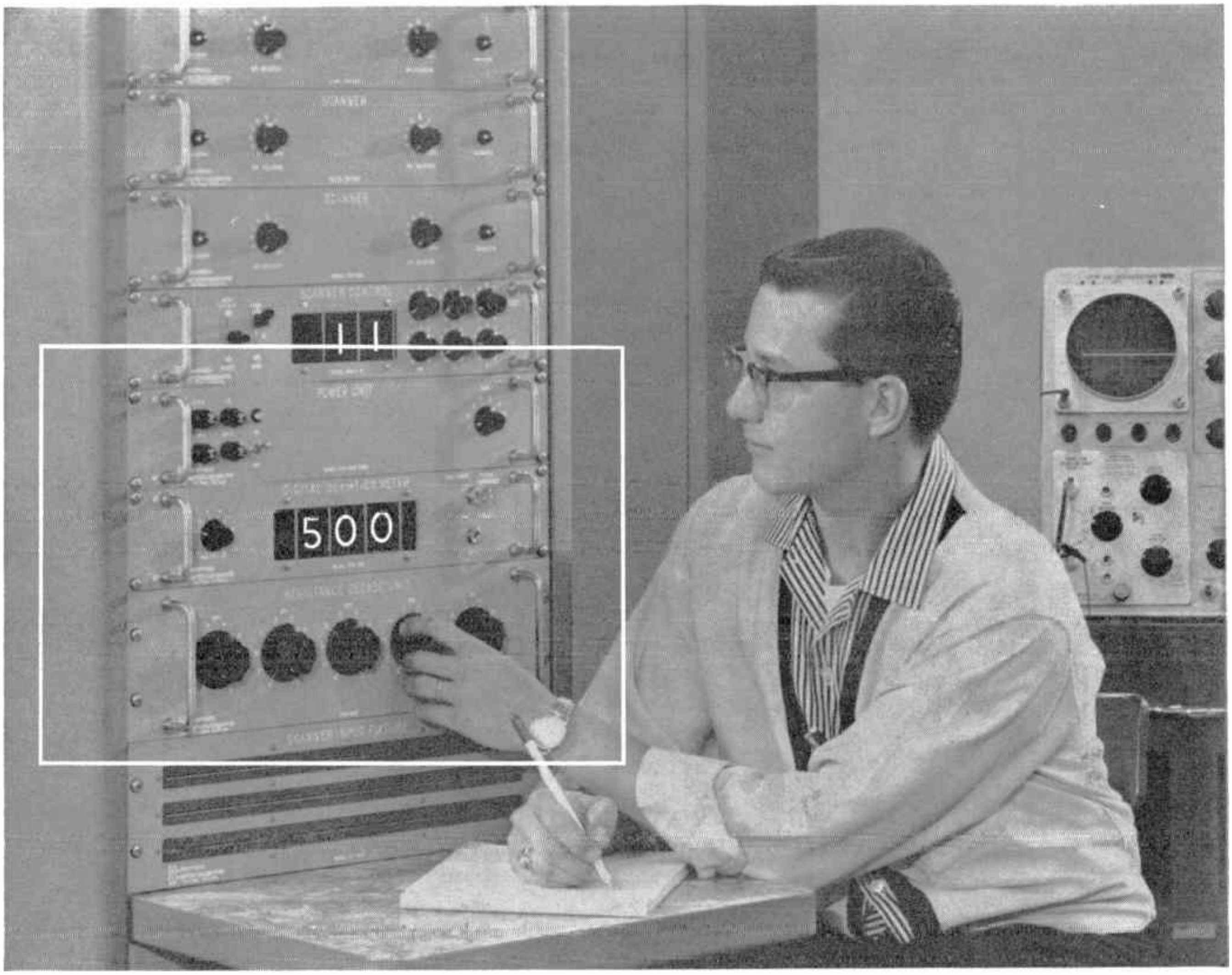
The new model utilizes the standard BR-7 header with an additional set of single pole, normally open contacts which ground one coil terminal to the case when the relay is energized. The 1A set of contacts is rated at 1 ampere inductive and is

(Continued on page 198A)

Use your

IRE DIRECTORY!

It's valuable!



New EI Deviation Ohmmeter...

5 digit accuracy with 3 digits!

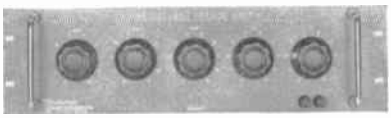
Automatically measures resistance, presents results in percentage of deviation from nominal value!

Testing of resistors can now be done 40% faster and with almost complete freedom from human error. The new EI Model DDM-300 Deviation Ohmmeter measures around any selected nominal value and presents the information in digital form as a direct percentage of this value. Possibility of ambiguous readings, reading the wrong range, or mistakes in calculating deviation from an absolute value, is eliminated.

Speeds testing – Using three digits, instead of five, and elimination of time consuming range changing, reduce the average reading time by 40%. Average balance time is one second.

0.01% accuracy—The instrument reads ±5% of the total resistance to an absolute accuracy of 0.01%. The chosen ±5% range is measured to a full, three-digit accuracy. Essentially, reading accuracy is limited only by the absolute accuracy of the nominal value.

The new instrument is particularly suited for automatic measurement of temperature co-efficients. Contact closures are available for go/no go and tolerance sorting operations.



Values can be referenced to a single resistor of known value or set with the optional precision resistance decade unit (shown above).

If you are testing resistors in quantity, you will want complete information on this extremely versatile instrument. Get in touch with your nearest EI field engineer today.

BRIEF SPECIFICATIONS

Display:
3 digits, over range indication.

Resistance range:
10 ohms to 10 megohms.

Accuracy:
The instrument reads ±5% of the total resistance to an absolute accuracy of 0.01%.

Speed:
1 second, average.

Engineers: Many challenging positions are open. For details contact Mr. Carl Sibelius.

Electro Instruments, Inc.

EI 3540 AERO COURT
SAN DIEGO 11, CALIF.

Digital instruments for measuring AC/DC voltages, AC/DC ratios, resistance, capacitance, and frequency • X-Y recorders and accessories • DC amplifiers

NEWS New Products

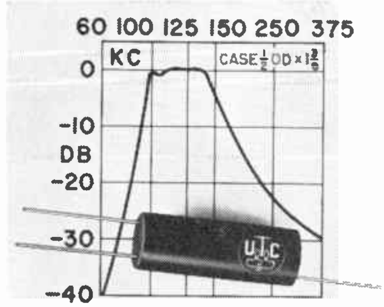
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 196A)

intended only as a holding contact for coil power.

The BR-7A is available with 2 form "C" contacts (DPDT) rated at 10 amperes resistive for BR-7AX, 5 amperes resistive for BR-7AY, and dry circuit to 2 amperes resistive for BR-7AZ. Pull-in power of the various types is: BR-7AX, 500 mw; BR-7AY, 225 mw; BR-7AZ, 150 mw.

High Frequency Filters



Type HF Minifilters ultra-miniaturized high frequency types weighing one-third ounce, with dimensions ½" diameter

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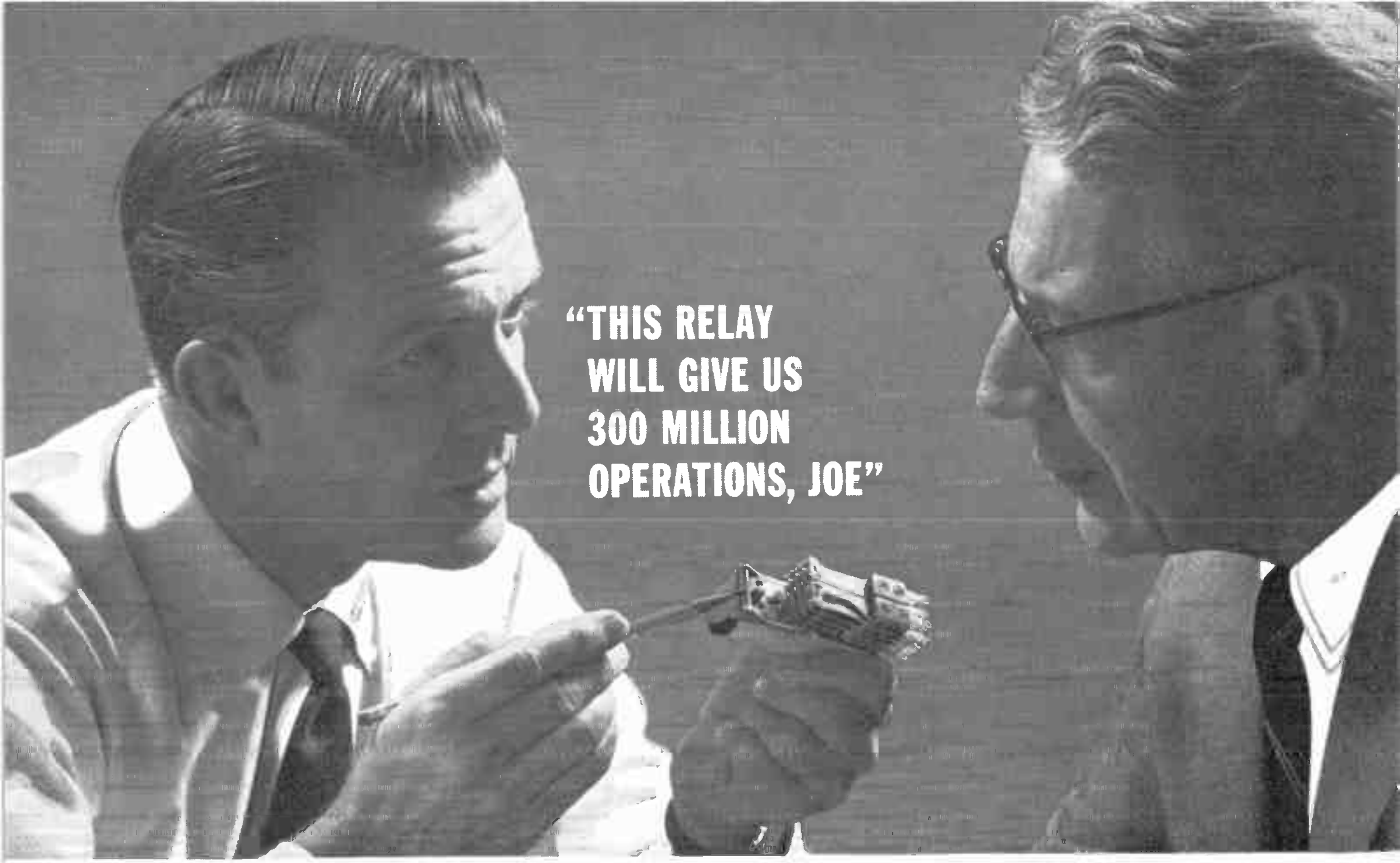
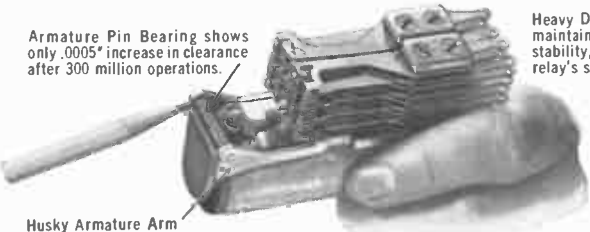


HERE'S WHY P&B TELEPHONE TYPE RELAYS GIVE YOU

reliable performance over long life



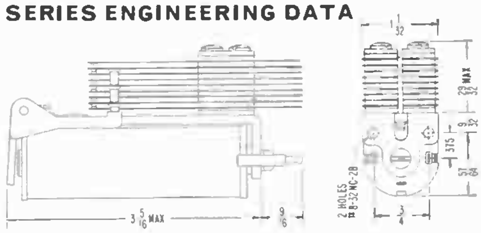
BS SERIES TELEPHONE TYPE

Measure the thickness of the BS series armature arm. You will find the cross section area is greater than ordinary relays of this type. Here is the kind of quality that spells dependability.

Observe that the stainless steel hinge pin runs the full width (not just half) of the armature, providing optimum bearing surface. This pin, operating in a stainless steel sleeve, shows only minimal wear during nearly a *third of a billion operations.*

Best of all, P&B quality costs no more. A whole new plant is being devoted to the production of high performance telephone type relays. Your nearest P&B sales engineer will be happy to discuss your relay problems. Call him today.

BS SERIES ENGINEERING DATA



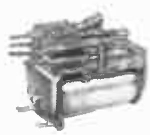
GENERAL:
Breakdown Voltage: 1000 volts rms 60 cy. min. between all elements.
Ambient Temperature: −55° to +85° C. +125° C available on special order.
Weight: 9 to 16 ozs.
Terminals: Pierced solder lugs; Coil: One #16 AWG wire Contacts: Two #18 AWG wires
Enclosures: Dust covered or sealed

CONTACTS:
Arrangements: DC—up to 28 springs AC—up to 24 springs
Material: 1/16" dia. twin palladium. Up to 1/4" dia. single silver. Other materials on special order.
Load: 4 amps at 115 volts, 60 cycle resistive
Pressure: 15 grams minimum

COILS:
Resistance: 100,000 ohms maximum Current: 10 amps maximum
Power: DC—50 Milliwatts per movable arm. Greater sensitivity on special order. AC—17.9 volt-amps.
Duty: Continuous
Treatment: Centrifugal impregnation
Voltages: DC—up to 300 volts with series resistor. AC—up to 250 volts, 60 cy.

MOUNTING: Two #8-32 tapped holes 3/4" o.c. Other mountings on special order.

GS SERIES—Excellent sensitivity: 50 mw per movable arm minimum (DC). For applications requiring many switching elements in small space.



LS SERIES—Medium coil relay with short springs and light weight armature for fast action, reliability and long life.

TS SERIES—Short coil relay is available in AC and DC versions. Long life construction. Can be supplied (DC) with up to 20 springs (10 per stack).

P&B STANDARD RELAYS ARE AVAILABLE AT YOUR LOCAL ELECTRONIC PARTS DISTRIBUTOR

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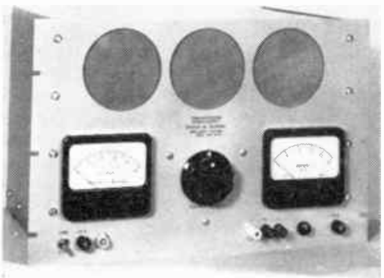
NEWS New Products

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

(Continued from page 198A)

by $1\frac{3}{8}''$ long, are available from **United Transformer Corp.**, 150 Varick St., New York 13, N. Y. Type HF are available in band pass types for frequencies in the range of 50 kc to 1 mc, low pass types with cutoff frequencies from 75 kc to 1 mc, and in high pass types with cutoff in the range of 100 kc to 1 mc. The unit illustrated is a 15,000 ohm band pass filter with a center frequency of 120 kc, flat within 3 db from 95 kc to 145 kc. Attenuation is greater than 40 db below 60 kc and greater than 30 db above 375 kc. Typical is a 1500 ohm unit having 150 kc cutoff. Attenuation is less than 3 db below 150 kc and greater than 45 db above 250 kc. Typical of a high pass HF Minifilter is a 7500 ohm unit with 150 kc cutoff. Attenuation is less than 3 db above 150 kc and greater than 45 db below 85 kc.

Transistorized DC Supply



Universal Electronics Co., 1720 Twenty Second St., Santa Monica, Calif., announces a new line of completely transistorized power supplies having large output current capacity. Typical model LQ5-32-25 has a voltage range of 5 to 32 volts dc at 0 to 25 amperes. Specifications are as follows: regulation—line or load, 50 mv, ripple—less than 2 millivolts, response time —less than 50 microseconds, output impedance—less than 0.005 ohm, rack mounting size—$12\frac{1}{4}''\times19''\times14\frac{3}{4}''$ deep.

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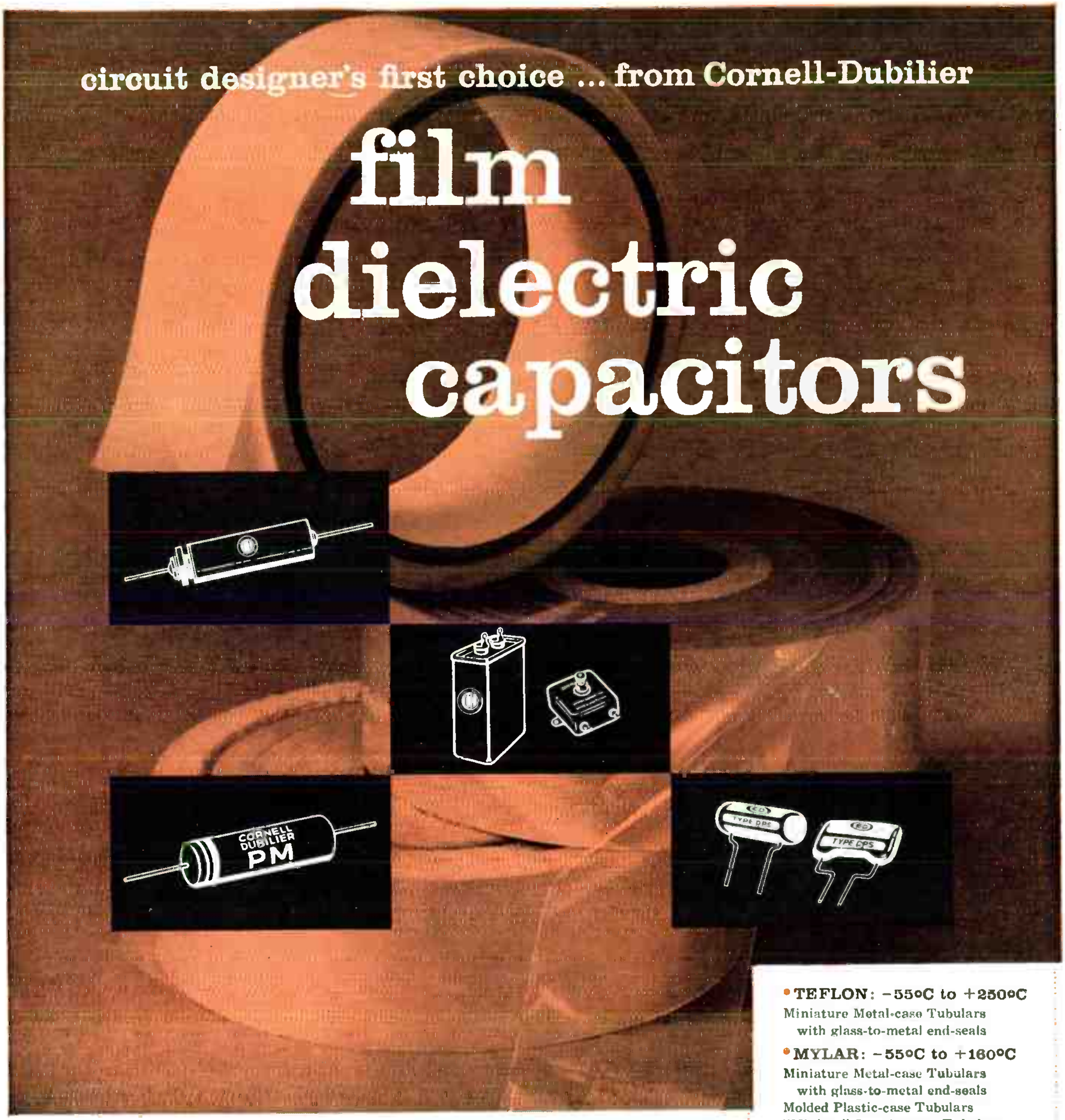

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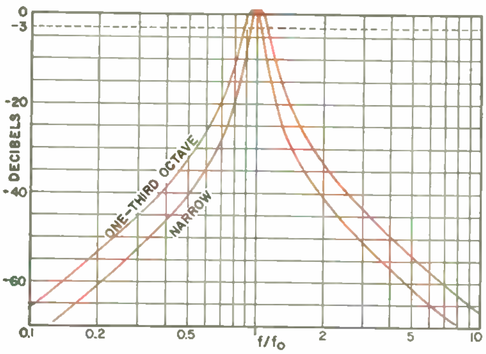


★ THREE BANDWIDTHS:
"**Narrow**" for measurement of single-frequency components of spectrum — bandwidth is 8% of center frequency at 3-db points.
"**One-Third Octave**" for analysis of noise — bandwidth is 1.26 times center frequency.
"**All-Pass**" for convenient adjustment of overall level — flat from 2.5c to 25 kc ±2 db.

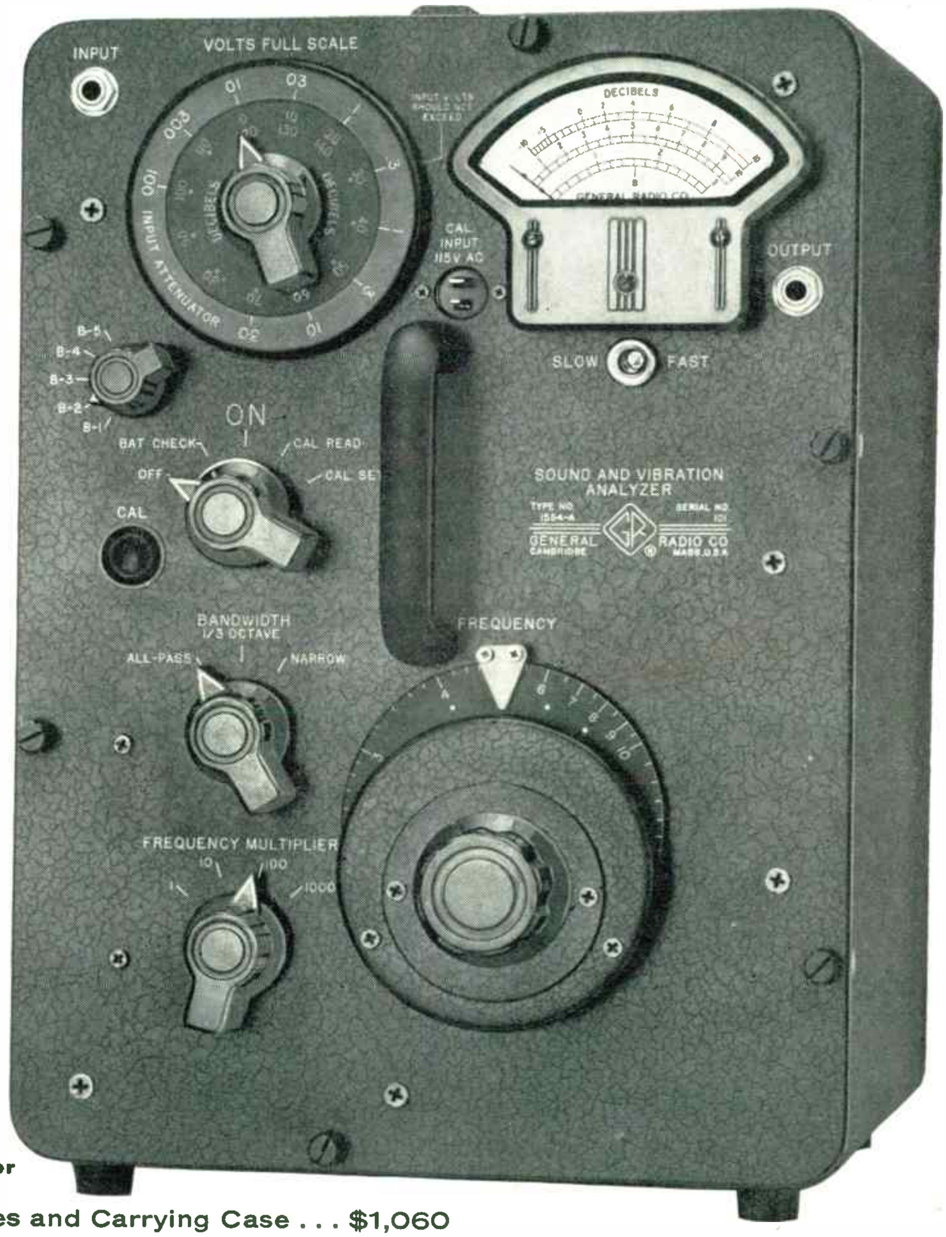
★ Wide Range: Continuously tunable from 2.5 to 25,000 cps.

★ Both Analyzer characteristics fall off at 12-db per octave far from center frequency.

★ Adjustable decibel dial may be set so Analyzer is direct reading in sound-pressure level, one-third octave-band level, or any other desired level — may be calibrated using 115v line to make Analyzer direct reading in input volts.

★ Portable and battery-operated for field use.

★ New G-R Type 1521-A Recorder can be linked to the Analyzer to provide a graphic record of the frequency spectrum under study.



Type 1554-A
Sound and Vibration Analyzer

Complete with Accessory Cables and Carrying Case . . . $1,060

This Analyzer is a tunable voltmeter whose bandwidth is a constant percentage of the frequency to which the instrument is tuned. It will measure amplitudes and frequencies of the individual components of sounds and vibrations. Frequency-selective amplifiers employing RC feedback networks can be either synchronously or stagger tuned to give "Narrow" or "One-Third Octave" bandwidth respectively. The amplifiers can also be switched to "All-Pass" operation to measure the entire input signal. Transistors have been used extensively throughout this instrument.

The instrument is designed for use with a Sound-Level or Vibration Meter. When component levels exceed 50 db, measurements may be made directly with accessory microphones.

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